Parameters identification for Interior Permanent Magnet Synchronous Motor driven by sensorless control

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Abstract

In this paper, a new parameter identification method for Interior Permanent Magnet Synchronous Motor (IPMSM) driven by sensorless control is proposed. Noise and vibration in mechanical plant becomes larger if rotor position estimation error occurs. Furthermore, the rotor position estimation error cause noise and vibration. When a load torque and a reference motor speed are constant, and phase voltages and phase currents in multiple stationary states are measured, motor parameters can be identified even if an estimated rotor position does not correspond with the rotor position. Furthermore, this paper proposes a new sensorless control method that derives the position by solving motor voltage equation. It is shown that the method is able to estimate the true position even if the motor is in acceleration. Numerical simulations with implementation of Pulse Width Modulation (PWM) inverter are illustrated to verify the effect of the methods.

Key words: parameter identification, sensorless control, interior permanent magnet synchronous motor, motor control.

1 Introduction

Interior Permanent Magnet Synchronous Motor (IPMSM) is well known because of its high efficiency, high output and wide speed range. Therefore, it is widely put into the practical use such as compressor, and is expected to be applied to Electric Vehicle (EV). Sensorless controls that estimate rotor angle of Permanent Magnet Synchronous Motor (PMSM) are widely studied to raise reliability and to make cost cutting because they can drive IPMSM without using motor angular sensors [1]−[7]. Most of the sensorless control requires motor parameters such as winding resistance, permanent magnet flux linkage and inductances. However, the parameters have data spread among products or vary across the ages. Therefore, there is a problem that the difference between the real parameters and the initially adjusted parameters for the sensorless controller makes the position estimation performance worse. In order to deal with this problem, some research projects developed parameter identification methods that identify motor parameters [8]−[13]. However, there is no report concerning identification method that theoretically assure to identify all the parameters in motor voltage equation without using the information of angular sensors or estimated motor position.

Air-conditioner, refrigerator, and electric vehicle might have operating conditions in which load torque and speed are constant for a while. In this case, that is, under the condition that the load torque and the speed are constant, some of the transient terms in the motor voltage equation are simplified. Then, this condition might help to identify the motor parameters without using the rotor position.

This paper proposes a new parameter identification method that does not require the information about the rotor angle. This method can identify the motor parameters under the following two conditions even if the estimated rotor angle derived by sensorless control does not correspond with the real angle. The first condition is that a load torque and a reference speed are constant. The second condition is that we can derive phase currents and phase voltages in multiple stationary states in which current phase and so on are different.

The conventional sensorless controls require an assumption that the motor speed is constant in order to make rotor position estimation error zero [7]. The assumption becomes the obstacle to achieve quick acceleration and deceleration. Noise and vibration in mechanical plant becomes larger if rotor position estimation error occurs. Furthermore, the rotor position estimation error cause noise and vibration. Therefore, the main applications of the sensorless drives are limited for air-conditioner, refrigerator and so on. This paper proposes a new sensorless method that does not require the assumption. This method acquires the rotor angle by solving the motor voltage equation. The calculation load of the method is larger, but the estimated angle corresponds with the real angle even if the motor is accelerated or decelerated. Therefore, the applications of the sensorless drive might spread.

In the second section, a new sensorless motor parameter identification method is proposed. In the third section, a new sensorless control that does not require the assumption of the motor speed to be constant is proposed. In the forth section, the effectiveness of the methods are verified using numerical simulations with implementation of Pulse Width Modulation (PWM) inverter.

2 Sensorless motor parameters identification

A new motor parameter identification method is proposed here, and it is presented that the method is available without using the information of the rotor angle.

2.1 IPMSM model

Motor coordinate system is shown in Fig. 1. θ is rotor angle. The subscript U, V, W mean quantities on fixed 3 phase axis, α, β mean quantities on orthogonal fixed 2 phase axis, and d, q mean quantities on orthogonal 2 phase axis synchronized with rotor angle θ [2]. γ, δ mean quantities on orthogonal 2 phase axis synchronized with estimated rotor angle $\hat{\theta}$ derived by sensorless drive.

θ_e is yielded by

$$\theta_e = \theta - \hat{\theta}. \tag{1}$$

dq-axis corresponds with $\gamma\delta$-axis when $\theta_e = 0$.

IPMSM model on $\alpha\beta$ fixed coordinate is yielded by the following equations [7].

$$s_{\alpha\beta} = \dot{\Phi} \tag{2}$$

$$s_{\alpha\beta} = [s_\alpha \ \ s_\beta]^{\mathrm{T}} = v_{\alpha\beta} - R_a i_{\alpha\beta} - L_q \dot{i}_{\alpha\beta} \tag{3}$$

$$\Phi = (\phi_a + (L_d - L_q) i_d) \psi(\theta) \tag{4}$$

$$\psi(\theta) = \begin{bmatrix} \cos(\theta) \\ \sin(\theta) \end{bmatrix} \tag{5}$$

where T denotes the transposition of matrix, $v_{\alpha\beta} = [v_\alpha \ \ v_\beta]^{\mathrm{T}}$, $i_{\alpha\beta} = [i_\alpha \ \ i_\beta]^{\mathrm{T}}$ are the voltage and current vectors on $\alpha\beta$-axis, respectively. R_a is the winding resistance on UVW-axis. ϕ_a is the permanent magnet flux linkage. L_d, L_q are the dq-axis inductances. $\dot{f}$ means the derivation of f. dq-axis voltage v_{dq} and current i_{dq} are derived by the following rotational transformation of coordinate system [2].

$$i_{dq} = [i_d \ \ i_q]^{\mathrm{T}} = C(\theta) i_{\alpha\beta}, \tag{6}$$

$$v_{dq} = [v_d \ \ v_q]^{\mathrm{T}} = C(\theta) v_{\alpha\beta}, \tag{7}$$

$$C(\theta) = \begin{bmatrix} \cos(\theta) & \sin(\theta) \\ -\sin(\theta) & \cos(\theta) \end{bmatrix}. \tag{8}$$

$\gamma\delta$-axis voltage $v_{\gamma\delta}$ and current $i_{\gamma\delta}$ are derived by the following rotational transformation of coordinate system [2].

$$i_{\gamma\delta} = [i_\gamma \ \ i_\delta]^{\mathrm{T}} = C\left(\hat{\theta}\right) i_{\alpha\beta}, \tag{9}$$

$$v_{\gamma\delta} = [v_\gamma \ \ v_\delta]^{\mathrm{T}} = C\left(\hat{\theta}\right) v_{\alpha\beta}. \tag{10}$$

From now, a motor model on $\gamma\delta$-axis is considered. By multiplying $C(\hat{\theta})$ to $s_{\alpha\beta}$ from the left, we get

$$s_{\gamma\delta} = C\left(\hat{\theta}\right) s_{\alpha\beta}. \tag{11}$$

Substituting (3), (9) and (10) to (11), we get

$$\begin{aligned} s_{\gamma\delta} &= C\left(\hat{\theta}\right) v_{\alpha\beta} - R_a C\left(\hat{\theta}\right) i_{\alpha\beta} - L_q C\left(\hat{\theta}\right) \dot{i}_{\alpha\beta} \\ &= v_{\gamma\delta} - R_a i_{\gamma\delta} - L_q C\left(\hat{\theta}\right) \dot{i}_{\alpha\beta}. \end{aligned} \tag{12}$$

In order to represent $C(\hat{\theta})\dot{i}_{\alpha\beta}$ in (12) by $i_{\gamma\delta}$, (9) is differentiated by time.

$$\begin{aligned} \dot{i}_{\gamma\delta} &= \dot{C}\left(\hat{\theta}\right) i_{\alpha\beta} + C\left(\hat{\theta}\right) \dot{i}_{\alpha\beta} \\ \therefore C\left(\hat{\theta}\right) \dot{i}_{\alpha\beta} &= \dot{i}_{\gamma\delta} - \dot{C}\left(\hat{\theta}\right) i_{\alpha\beta} \\ &= \dot{i}_{\gamma\delta} - \dot{\hat{\theta}} \begin{bmatrix} -\sin\left(\hat{\theta}\right) & \cos\left(\hat{\theta}\right) \\ -\cos\left(\hat{\theta}\right) & -\sin\left(\hat{\theta}\right) \end{bmatrix} i_{\alpha\beta} \\ &= \dot{i}_{\gamma\delta} + \dot{\hat{\theta}} J C\left(\hat{\theta}\right) i_{\alpha\beta} \end{aligned} \tag{13}$$

where J is

$$J = \begin{bmatrix} 0 & -1 \\ 1 & 0 \end{bmatrix}. \tag{14}$$

Substituting (9) to (13), we get

$$C\left(\hat{\theta}\right)\dot{i}_{\alpha\beta} = \dot{i}_{\gamma\delta} + \hat{\omega}Ji_{\gamma\delta} \quad (15)$$

where $\hat{\omega} = \dot{\hat{\theta}}$. Substituting (15) to (12), $s_{\gamma\delta}$ is represented with $v_{\gamma\delta}$ and $i_{\gamma\delta}$ as

$$s_{\gamma\delta} = v_{\gamma\delta} - R_a i_{\gamma\delta} - L_q\left(\dot{i}_{\gamma\delta} + \hat{\omega}Ji_{\gamma\delta}\right). \quad (16)$$

Next, $C(\hat{\theta})\dot{\Phi}$ is represented with $\psi(\theta_e)$ instead of $\psi(\theta)$. From (4), we get

$$\begin{aligned}\dot{\Phi} &= (\phi_a + (L_d - L_q) i_d)\dot{\psi}(\theta) + (L_d - L_q)\dot{i}_d\psi(\theta) \\ &= \omega(\phi_a + (L_d - L_q) i_d) J\psi(\theta) \\ &\quad + (L_d - L_q)\dot{i}_d\psi(\theta) \quad (17)\end{aligned}$$

where $\omega(= \dot{\theta})$ is motor speed. By multiplying $C(\hat{\theta})$ to (17) from the left, we get

$$\begin{aligned}&C\left(\hat{\theta}\right)\dot{\Phi} \\ &= \omega(\phi_a + (L_d - L_q) i_d) C\left(\hat{\theta}\right) J\psi(\theta) \\ &\quad + (L_d - L_q)\dot{i}_d C\left(\hat{\theta}\right)\psi(\theta) \\ &= \omega(\phi_a + (L_d - L_q) i_d) \\ &\quad \times \begin{bmatrix} -\cos\left(\hat{\theta}\right)\sin(\theta) + \sin\left(\hat{\theta}\right)\cos(\theta) \\ \sin\left(\hat{\theta}\right)\sin(\theta) + \cos\left(\hat{\theta}\right)\cos(\theta) \end{bmatrix} \\ &\quad + (L_d - L_q)\dot{i}_d \\ &\quad \times \begin{bmatrix} \cos\left(\hat{\theta}\right)\cos(\theta) + \sin\left(\hat{\theta}\right)\sin(\theta) \\ -\sin\left(\hat{\theta}\right)\cos(\theta) + \cos\left(\hat{\theta}\right)\sin(\theta) \end{bmatrix} \\ &= \omega(\phi_a + (L_d - L_q) i_d)\begin{bmatrix} -\sin\left(\theta - \hat{\theta}\right) \\ -\cos\left(\theta - \hat{\theta}\right) \end{bmatrix} \\ &\quad + (L_d - L_q)\dot{i}_d \begin{bmatrix} \cos\left(\theta - \hat{\theta}\right) \\ \sin\left(\theta - \hat{\theta}\right) \end{bmatrix}. \quad (18)\end{aligned}$$

Substituting (1), (5) and (14) to (18), we get the following equation that represents $C(\hat{\theta})\dot{\Phi}$ with $\psi(\theta_e)$ instead of $\psi(\theta)$.

$$\begin{aligned}C\left(\hat{\theta}\right)\dot{\Phi} &= \omega(\phi_a + (L_d - L_q) i_d) J\psi(\theta_e) \\ &\quad + (L_d - L_q)\dot{i}_d\psi(\theta_e). \quad (19)\end{aligned}$$

By rewriting (2), (11), (16) and (19), we get a $\gamma\delta$-axis motor model as

$$s_{\gamma\delta} = C\left(\hat{\theta}\right)\dot{\Phi} \quad (20)$$

$$s_{\gamma\delta} = v_{\gamma\delta} - R_a i_{\gamma\delta} - L_q\left(\dot{i}_{\gamma\delta} + \hat{\omega}Ji_{\gamma\delta}\right) \quad (21)$$

$$\begin{aligned}C\left(\hat{\theta}\right)\dot{\Phi} &= \omega(\phi_a + (L_d - L_q) i_d) J\psi(\theta_e) \\ &\quad + (L_d - L_q)\dot{i}_d\psi(\theta_e). \quad (22)\end{aligned}$$

A torque equation is as follows [2]

$$T = p(\phi_a + (L_d - L_q) i_d) i_q \quad (23)$$

where T is a motor torque, and p is the number of pole pairs.

2.2 Properties in stationary state

Properties of signals in stationary state are presented when the load torque and the reference speed are constant. The motor torque T and the motor speed ω have the following relation[2]

$$\omega = \frac{1}{J_m s + D_m}(T - T_L) \quad (24)$$

where J_m and D_m are the moment of inertia and the viscous friction, respectively. We make assumptions that both of the sensorless control system and motor control system are stable, and a constant $\hat{\omega}$ can be detected. Then, a synchronized driving is realized because the motor control system is stable, and then, it follows

$$\hat{\omega} = \omega. \quad (25)$$

From (24), the motor torque T is controlled to be constant, and from (23), v_{dq} and i_{dq} become constant in stationary state. And then, the stationary value of $\theta_e(= \theta - \hat{\theta})$ become constant because the sensorless control is stable. From (6), (7), (9) and (10), it follows

$$i_{\gamma\delta} = C\left(\hat{\theta} - \theta\right) i_{dq}, \quad (26)$$

$$v_{\gamma\delta} = C\left(\hat{\theta} - \theta\right) v_{dq} \quad (27)$$

and then, $v_{\gamma\delta}$ and $i_{\gamma\delta}$ also become constant.

2.3 R_a identification

The winding resister R_a identification in stationary state is presented.

2.3.1 Basic equation of R_a identification

In stationary state, because i_d is constant, we get $\dot{i}_d = 0$, and from (17), we get

$$\dot{\Phi} = \omega(\phi_a + (L_d - L_q) i_d) J\psi(\theta). \quad (28)$$

From (2), (28), (5) and (14), we get

$$\|s_{\alpha\beta}\| = \omega\left(\phi_a + (L_d - L_q)\, i_d\right) \quad (29)$$
$$\frac{s_\alpha}{\|s_{\alpha\beta}\|} = -\sin(\theta) \quad (30)$$
$$\frac{s_\beta}{\|s_{\alpha\beta}\|} = \cos(\theta) \quad (31)$$

where $\|\cdot\|$ means 2 norm. From (23) and (29), we get

$$\omega T = p\|s_{\alpha\beta}\| i_q. \quad (32)$$

Substituting (6) into (32), we get

$$\omega T = p\|s_{\alpha\beta}\| \left[-\sin(\theta) \quad \cos(\theta)\right] i_{\alpha\beta}. \quad (33)$$

Substituting (30) and (31) into (33), from (3), we get

$$\begin{aligned} \omega T &= p\|s_{\alpha\beta}\| \left[\frac{s_\alpha}{\|s_{\alpha\beta}\|} \quad \frac{s_\beta}{\|s_{\alpha\beta}\|}\right] i_{\alpha\beta} \\ &= p{s_{\alpha\beta}}^{\mathrm{T}} i_{\alpha\beta}. \end{aligned} \quad (34)$$

From (34), (3) and (25), it is verified that T can be estimated using $v_{\alpha\beta}$, $i_{\alpha\beta}$ and $\hat{\omega}$ in stationary state even if θ and $\hat{\theta}$ are unknown. Because $C(\hat{\theta})^{\mathrm{T}}C(\hat{\theta})$ becomes an unit matrix, from (34), we get

$$\begin{aligned} \omega T &= p{s_{\alpha\beta}}^{\mathrm{T}} \left(C\left(\hat{\theta}\right)^{\mathrm{T}} C\left(\hat{\theta}\right)\right) i_{\alpha\beta} \\ &= p\left(C\left(\hat{\theta}\right) s_{\alpha\beta}\right)^{\mathrm{T}} C\left(\hat{\theta}\right) i_{\alpha\beta}. \end{aligned} \quad (35)$$

Substituting (9), (11) and (21) to (35), we get

$$\begin{aligned} \omega T &= p{s_{\gamma\delta}}^{\mathrm{T}} i_{\gamma\delta} \\ &= p\left(v_{\gamma\delta} - R_a i_{\gamma\delta} - L_q \dot{i}_{\gamma\delta} + \hat{\omega} J i_{\gamma\delta}\right)^{\mathrm{T}} i_{\gamma\delta} \\ &= p\left(v_{\gamma\delta} - R_a i_{\gamma\delta} - L_q \dot{i}_{\gamma\delta}\right)^{\mathrm{T}} i_{\gamma\delta} + p\hat{\omega} {i_{\gamma\delta}}^{\mathrm{T}} J^{\mathrm{T}} i_{\gamma\delta}. \end{aligned}$$

Substituting ${i_{\gamma\delta}}^{\mathrm{T}} J^{\mathrm{T}} i_{\gamma\delta} = i_\gamma i_\delta - i_\delta i_\gamma = 0$ and $\dot{i}_{\gamma\delta} = [0 \;\; 0]^{\mathrm{T}}$, we get the following basic equation to identify R_a

$$\omega T = p\left(v_{\gamma\delta} - R_a i_{\gamma\delta}\right)^{\mathrm{T}} i_{\gamma\delta}. \quad (36)$$

From (36) and (25), it follows that the torque T can be estimated using $v_{\gamma\delta}$, $i_{\gamma\delta}$ and $\hat{\omega}$ in stationary state even if the rotor position estimation error θ_e is unknown.

2.3.2 R_a identification at standstill

Because the motor speed ω is 0 at standstill, R_a can be identified by substituting $v_{\gamma\delta}$ and $i_{\gamma\delta}$ to the following equation that is derived by substituting $\omega = 0$ to (36), and by solving for R_a.

$$\hat{R}_a = \frac{{v_{\gamma\delta}}^{\mathrm{T}} i_{\gamma\delta}}{{i_{\gamma\delta}}^{\mathrm{T}} i_{\gamma\delta}} \quad (37)$$

where the superscript ˆ means the estimated value.

2.3.3 R_a identification at the operating state

Two stationary states $i (= 1,\ 2)$ are considered. In the stationary state 1 and 2, it is assumed that the motor torques $T^{(1)}$ and $T^{(2)}$ are equal, and the current phases $\beta^{(1)}$ and $\beta^{(2)}$ or the rotational speeds $\omega^{(1)}$ and $\omega^{(2)}$ are different. The superscript (i) means the signal in the stationary state i. From (25), the estimated motor speed $\hat{\omega}^{(i)}$ corresponds with the motor speed $\omega^{(i)}$. In the stationary state 1, from (36), we get

$$T^{(1)} = \frac{p}{\hat{\omega}^{(1)}}\left(v_{\gamma\delta}^{(1)} - R_a i_{\gamma\delta}^{(1)}\right)^{\mathrm{T}} i_{\gamma\delta}^{(1)}. \quad (38)$$

In the stationary state 2, from (36), we get

$$T^{(2)} = \frac{p}{\hat{\omega}^{(2)}}\left(v_{\gamma\delta}^{(2)} - R_a i_{\gamma\delta}^{(2)}\right)^{\mathrm{T}} i_{\gamma\delta}^{(2)}. \quad (39)$$

When $T^{(1)} = T^{(2)}$, by solving (38) and (39) for R_a, we get

$$\hat{R}_a = \frac{v_{\gamma\delta}^{(1)\mathrm{T}} i_{\gamma\delta}^{(1)} \hat{\omega}^{(2)} - v_{\gamma\delta}^{(2)\mathrm{T}} i_{\gamma\delta}^{(2)} \hat{\omega}^{(1)}}{i_{\gamma\delta}^{(1)\mathrm{T}} i_{\gamma\delta}^{(1)} \hat{\omega}^{(2)} - i_{\gamma\delta}^{(2)\mathrm{T}} i_{\gamma\delta}^{(2)} \hat{\omega}^{(1)}} \quad (40)$$

that can identify R_a using $\gamma\delta$-axis voltage $v_{\gamma\delta}^{(i)}$, $\gamma\delta$-axis current $i_{\gamma\delta}^{(i)}$ and estimated motor speed $\hat{\omega}^{(i)}$.

2.4 L_d and ϕ_a identification

Here, we present a L_d and ϕ_a identification method using signals in stationary state when a $\hat{L}_q$ is given. It is assumed that $\hat{R}_a$ is already identified to be R_a, that is, it follows $\hat{R}_a = R_a$. By the method, it is presented that ϕ_a and L_d can be identified to be true values when $\hat{L}_q = L_q$. Because L_q is unknown, $s_{\gamma\delta}$ in (21) is estimated using $\hat{L}_q$ instead of L_q as

$$\hat{s}_{\gamma\delta} = v_{\gamma\delta} - \hat{R}_a i_{\gamma\delta} - \hat{L}_q \hat{\omega} J i_{\gamma\delta} \quad (41)$$

where $\dot{i}_{\gamma\delta} = [0 \;\; 0]^{\mathrm{T}}$ is used because of the stationary state. $\hat{\theta}_e$ is derived as

$$\hat{\theta}_e = \tan^{-1}\left(-\frac{\hat{s}_\gamma}{\hat{s}_\delta}\right). \quad (42)$$

Using $\hat{\theta}_e$ derived by (42), $\hat{v}_{\gamma\delta}$ and $\hat{i}_{\gamma\delta}$ are derived as

$$\hat{v}_{\gamma\delta} = C\left(\hat{\theta}_e\right) v_{\gamma\delta} \quad (43)$$
$$\hat{i}_{\gamma\delta} = C\left(\hat{\theta}_e\right) i_{\gamma\delta}. \quad (44)$$

Substituting $v_{\gamma\delta}$ and $i_{\gamma\delta}$ in (41), $\bar{s}_{\gamma\delta}$ is calculated using $\hat{v}_{\gamma\delta}$ and $\hat{i}_{\gamma\delta}$ as

$$\bar{s}_{\gamma\delta} = \hat{v}_{\gamma\delta} - \hat{R}_a \hat{i}_{\gamma\delta} - \hat{L}_q \hat{\omega} J \hat{i}_{\gamma\delta}. \quad (45)$$

From (21), (41), $\hat{R}_a = R_a$ and (25), it follows $\hat{s}_{\gamma\delta} = s_{\gamma\delta}$ if $\hat{L}_q = L_q$. Then using (20) and (22), $\hat{\theta}_e$ derived by (42) corresponds with θ_e because it follows $\dot{i}_d = 0$ in stationary state. Using $\hat{\theta}_e = \theta_e$, from (43) and (44), we get

$$\hat{v}_{\gamma\delta} = v_{dq} \quad (46)$$
$$\hat{i}_{\gamma\delta} = i_{dq}. \quad (47)$$

In the above calculations, L_d and ϕ_a are not used. From (20), (22) and (45), by adopting an error term $\bar{e}_\delta$ that becomes zero when $\hat{L}_q = L_q$, it follows

$$\bar{s}_\delta = \omega \left(\phi_a + (L_d - L_q)\, i_d\right) + \bar{e}_\delta. \quad (48)$$

Therefore, we get

$$\bar{s}_\delta + \omega L_q i_d = \begin{bmatrix} \omega & \omega i_d \end{bmatrix} \begin{bmatrix} \phi_a \\ L_d \end{bmatrix} + \bar{e}_\delta. \quad (49)$$

When $\hat{L}_q = L_q$, it is possible to set $\hat{i}_\gamma$ correspond with i_d. By adopting an error term $\bar{e}_{\delta 1}$ that becomes zero when $\hat{i}_\gamma = i_d$, it follows

$$\bar{s}_\delta + \hat{\omega} \hat{L}_q \hat{i}_\gamma = \begin{bmatrix} \hat{\omega} & \hat{\omega} \hat{i}_\gamma \end{bmatrix} \begin{bmatrix} \phi_a \\ L_d \end{bmatrix} + \bar{e}_{\delta 1} \quad (50)$$

$$\bar{e}_{\delta 1} = \omega \left(i_d - \hat{i}_\gamma\right) + \bar{e}_\delta. \quad (51)$$

From (50), using γ-axis stationary current $\hat{i}_\gamma^{(i)}$ and stationary motor speed $\hat{\omega}^{(i)}$, ϕ_a and L_d are identified using the least square method as follows

$$\begin{bmatrix} \hat{\phi}_a \\ \hat{L}_d \end{bmatrix} = \left(\Omega^{\mathrm{T}} \Omega\right)^{-1} \left(\Omega^{\mathrm{T}} Y\right) \quad (52)$$

$$\Omega = \begin{bmatrix} \hat{\omega}^{(1)} & \hat{\omega}^{(1)} \hat{i}_\gamma^{(1)} \\ \hat{\omega}^{(2)} & \hat{\omega}^{(2)} \hat{i}_\gamma^{(2)} \\ \vdots & \vdots \end{bmatrix} \quad (53)$$

$$Y = \begin{bmatrix} \bar{s}_\delta + \hat{\omega}^{(1)} \hat{L}_q \hat{i}_\gamma^{(1)} \\ \bar{s}_\delta + \hat{\omega}^{(2)} \hat{L}_q \hat{i}_\gamma^{(2)} \\ \vdots \end{bmatrix}. \quad (54)$$

The true values of ϕ_a and L_d are derived because it follows $\bar{e}_{\delta 1} = 0$ when $\hat{L}_q = L_q$.

2.5 L_q identification

$\bar{e}_{\delta 1}$ in (51) becomes zero when $\hat{L}_q = L_q$. If the number of stationary state i is large enough, it follows $\bar{e}_{\delta 1} = 0$ in all the stationary state $1 \sim i$ only when $\hat{L}_q = L_q$. Therefore, the following residual error e has the same property as $\bar{e}_{\delta 1}$.

$$e = Y - \Omega \begin{bmatrix} \hat{\phi}_a \\ \hat{L}_d \end{bmatrix}. \quad (55)$$

Using e, $\hat{L}_q$ is identified so that the following cost function J_{Lq} is minimized by one variable optimization such as iteration.

$$J_{Lq}\left(\hat{L}_q\right) = e^{\mathrm{T}} e. \quad (56)$$

L_q can be identified because non-zero function J_{Lq} becomes the minimum when $\hat{L}_q = L_q$.

From now, another cost functions are presented. Torque can be estimated by substituting $\hat{\omega}^{(i)}$, $\hat{i}_{\gamma\delta}^{(i)}$, $\hat{v}_{\gamma\delta}^{(i)}$ and $\hat{R}_a$ in (36). $\bar{L}_q^{(i)}$ that is an estimated L_q is derived by substituting the estimated torque, $\hat{\phi}_a$ and $\hat{L}_d$ into the torque equation (23). The number of $\bar{L}_q^{(i)}$ is i because $\bar{L}_q^{(i)}$ can be derived at each stationary state i. When $\hat{L}_q = L_q$, it is considered to follow $\hat{\phi} = \phi$, $\hat{L}_d = L_d$ and $\bar{L}_q^{(i)} = L_q$. Therefore, the following two cost functions are introduced.

$$J_1\left(\hat{L}_q\right) = \sum_i \sum_j \left(\bar{L}_q^{(i)} - \bar{L}_q^{(j)}\right)^2 \quad (57)$$

$$J_2\left(\hat{L}_q\right) = \sum_i \left(\hat{L}_q - \bar{L}_q^{(i)}\right)^2. \quad (58)$$

L_q can be identified because $J_1(\hat{L}_q)$ and $J_2(\hat{L}_q)$ are non-negative functions, and become zeros that are the minimum value when $\hat{L}_q = L_q$.

One variable optimization requires larger calculation load than the motor controller. However, once the calculation was done, recalculation will not be required for a time. Therefore, it does not require high performance CPU if the calculation is done while motor is not driven, or while the CPU is not busy by setting the priority of the calculation low.

3 Sensorless drive

Here, we propose a novel sensorless drive that makes the position estimation error zero quickly. Conventional methods require the assumption that ω is constant [7]. One of the features of our method is that the assumption is not required.

Substituting (8) into (6), we get

$$i_d = \begin{bmatrix} \cos(\theta) & \sin(\theta) \end{bmatrix} i_{\alpha\beta}. \quad (59)$$

Differentiating with time, we get

$$\dot{i}_d = \omega \begin{bmatrix} -\sin(\theta) & \cos(\theta) \end{bmatrix} i_{\alpha\beta} + \begin{bmatrix} \cos(\theta) & \sin(\theta) \end{bmatrix} \dot{i}_{\alpha\beta}. \quad (60)$$

Substituting (5) and (14), we get

$$\dot{i}_d = \omega (J\psi(\theta))^{\mathrm{T}} i_{\alpha\beta} + \psi(\theta)^{\mathrm{T}} \dot{i}_{\alpha\beta}. \quad (61)$$

Substituting (3), (17) and (61) into (2), we get

$$x_{\alpha\beta} = \omega y_{\alpha\beta} \quad (62)$$

$$\begin{aligned} x_{\alpha\beta} &= [x_\alpha \ \ x_\beta]^{\mathrm{T}} \\ &= v_{\alpha\beta} - R_a i_{\alpha\beta} - L_q \dot{i}_{\alpha\beta} \\ &\quad - (L_d - L_q) \psi(\theta)^{\mathrm{T}} \dot{i}_{\alpha\beta} \psi(\theta) \end{aligned} \quad (63)$$

$$\begin{aligned} y_{\alpha\beta} &= [y_\alpha \ \ y_\beta]^{\mathrm{T}} \\ &= (\phi_a + (L_d - L_q) i_d) J\psi(\theta) \\ &\quad + (L_d - L_q)(J\psi(\theta))^{\mathrm{T}} i_{\alpha\beta} \psi(\theta). \end{aligned} \quad (64)$$

The following cost function $J_\theta(\theta)$ is introduced as

$$J_\theta(\theta) = \|x_\alpha y_\beta - x_\beta y_\alpha\| \quad (65)$$

$$z = \frac{1}{T_f s + 1}(x_\alpha y_\beta - x_\beta y_\alpha) \quad (66)$$

where s is a differential operator, and $T_f > 0$. By using (66), high frequency noize in $\dot{i}_{\alpha\beta}$ used to get $x_{\alpha\beta}$ in (63) is filtered out. The position is estimated so as to search for the $\hat{\theta}$ that minimizes $J_\theta(\hat{\theta})$ as

$$\hat{\theta} = \arg\min_{\hat{\theta}} J_\theta\left(\hat{\theta}\right). \quad (67)$$

This one variable optimization requires much more load than conventional sensorless controller. However, this demerit will be resolved in the near future because the CPU power has been improved more and more in recent days. When $\hat{\theta} = \theta$, from (62), (63) and (64), the non-negative function $J_\theta(\hat{\theta})$ follows

$$J_\theta\left(\hat{\theta}\right) = 0 \quad (68)$$

that is the minimum value. Therefore, from (65) and (68), using a constant k,

$$x_{\alpha\beta} = k y_{\alpha\beta} \quad (69)$$

follows. From (62) and (69), we get

$$\omega = k. \quad (70)$$

Therefore, the motor speed $\hat{\omega}$ can be estimated as

$$\hat{\omega} = s_1 \frac{\|x_{\alpha\beta}\|}{\|y_{\alpha\beta}\|} \quad (71)$$

$$s_1 = sign(x_\alpha + x_\beta)\, sign(y_\alpha + y_\beta)$$

where $sign(x)$ denotes the sign of x.

Fig. 2 shows a block diagram of the method. The subscript $*$ means a reference signal, and PI means PI controller. β^* is the reference current phase, and has the relation as $\beta^* = \tan^{-1}(-i_\gamma{}^*/i_\delta{}^*)$. $\tan(\beta^*)$ is related with motor efficiency and maximum speed [2].

4 Numerical simulations

The effectiveness of the parameter identification method was verified using numerical simulations with implementation of PWM inverter, and it was shown that the sensorless control using the identified parameters was able to keep driving if the speed or the load torque were changed quickly.

4.1 Simulation setting

The IPMSM parameters employed here were from the reference [7].

$$L_d = 3.5[\mathrm{mH}]$$

$$L_q = 6.3[\mathrm{mH}]$$

$$R_a = 0.143[\Omega]$$

$$\phi_a = 0.176[\mathrm{Vs}]$$

$$J_m = 0.00018[\mathrm{Nms}^2]$$

$$D_m = 0.0[\mathrm{Nms}].$$

The specifications of driving area for this motor were as the minimum speed $15 \times 2\pi$[rad/s], the maximum speed $120 \times 2\pi$[rad/s], and the maximum load torque 15[Nm]. The sampling time was set as 0.1[ms].

In order to compensate the voltage drops caused by switching devices in PWM inverter, phase voltages are adjusted using dead time compensation [14].

4.2 Sensorless parameter identification results

Table 1 showed the differences on percentage between the true and identified parameters at various driving conditions. The current phase β was set as 20, 30, 40[deg] at each driving condition, and parameters were identified using the phase current $i_{\alpha\beta}$ and the phase voltage $v_{\alpha\beta}$ detected at each stationary state. The driving condition was set as combination of motor speeds ($20 \times 2\pi$[rad/s], $120 \times 2\pi$[rad/s]), load torques (1[Nm], 15[Nm]) and position estimation errors θ_e(2, 30[deg]). Table 1 showed that the maximum absolute value of the difference between the true and identified parameters was less than 0.03%. Therefore, the effectiveness of the method was verified,

Table 1: Motor parameters identification errors in various operating conditions

θ_e[deg]	ω[rad/s]	T_L[Nm]	R_a	L_q	L_d	ϕ_a
2	$20 \times 2\pi$	1	-1.3×10^{-12} %	$+3.0 \times 10^{-2}$ %	-3.0×10^{-4} %	-1.4×10^{-4} %
30	$20 \times 2\pi$	1	-1.3×10^{-12} %	$+3.0 \times 10^{-2}$ %	-3.0×10^{-4} %	-1.4×10^{-4} %
2	$120 \times 2\pi$	1	-8.6×10^{-12} %	$+3.0 \times 10^{-2}$ %	-3.0×10^{-4} %	-1.4×10^{-4} %
30	$120 \times 2\pi$	1	-8.6×10^{-12} %	$+3.0 \times 10^{-2}$ %	-3.0×10^{-4} %	-1.4×10^{-4} %
2	$20 \times 2\pi$	15	-1.6×10^{-12} %	$+3.0 \times 10^{-2}$ %	-2.6×10^{-2} %	-2.4×10^{-2} %
30	$20 \times 2\pi$	15	-1.6×10^{-12} %	$+3.0 \times 10^{-2}$ %	-2.6×10^{-2} %	-2.4×10^{-2} %
2	$120 \times 2\pi$	15	-1.6×10^{-12} %	$+3.0 \times 10^{-2}$ %	-2.6×10^{-2} %	-2.4×10^{-2} %
30	$120 \times 2\pi$	15	-1.1×10^{-11} %	$+3.0 \times 10^{-2}$ %	-2.6×10^{-2} %	-2.4×10^{-2} %

In Fig. 3 the solid line showed the cost function $J_{Lq}(\hat{L}_q)$ in (56) used for identifying $\hat{L}_q$ at the driving condition of the speed $20 \times 2\pi$[rad/s], the load torque 1[Nm] and the position estimation error $\theta_e = 30$[deg]. For the purpose of reference, $J_1(\hat{L}_q)$ in (57) and $J_2(\hat{L}_q)$ in (58) were shown as dotted line and dashed line, respectively. All of the cost functions $J_{Lq}(\hat{L}_q)$, $J_1(\hat{L}_q)$ and $J_2(\hat{L}_q)$ have the minimum zero at $\hat{L}_q = L_q$, which matched theoretically.

4.3 Sensorless drive results

The proposed sensorless control was performed using the identified parameters. Fig. 4 (a) showed the estimated speed $\hat{\omega}$ (dashed line), the real speed ω (solid line) and the α-axis current i_α (dotted line). Fig. 4 (b) showed the position estimation error θ_e. In order to verify the case the reference speed changed quickly, the reference speed was changed like a step function at 0.15[s] from $20 \times 2\pi$[rad/s] to $120 \times 2\pi$[rad/s], and at 0.25[s] from $120 \times 2\pi$[rad/s] to $20 \times 2\pi$[rad/s] when sensorless driving was being done. Furthermore, in order to check the case the load torque changed quickly, the load torque was changed like a step function at 0.3[s] from 1[Nm] to 15[Nm]. From Fig. 4 (a), $\hat{\omega}$ almost corresponded with ω. From Fig. 4 (b), the maximum absolute value of θ_e was about less than 2[deg], and the sensorless driving was able to continue.

Consequently, in the case of these simulations, the proposed method was able to identify motor parameters effectively, and good sensorless driving was achieved using the identified parameters.

5 Conclusion

In this paper, we proposed a novel parameter identification method available for sensorless driving, and a novel sensorless drive in which motor position was derived by solving the motor voltage equation. It was verified that the identification method was able to identify the motor parameters when a load torque and a reference motor speed were constant, and phase voltages and phase currents in multiple stationary states were measured, even if the estimated rotor position did not correspond with the rotor position. Noise and vibration in mechanical plant became larger if rotor position estimation error occurs. Furthermore, the rotor position estimation error caused noise and vibration. It was also verified that the sensorless drive required large calculation load but was able to make the position estimation error small even if the motor was in acceleration or the load torque was changed quickly.

The characteristics of IPMSM corresponds with SPMSM if $L_d = L_q$, or synchronous reluctance motor if $\phi_a = 0$ [2]. Therefore, proposed method can be applied to SPMSM and synchronous reluctance motor.

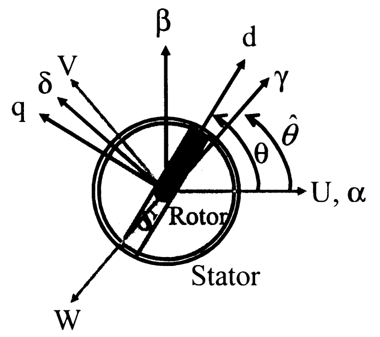


Figure 1: Motor coordinate system

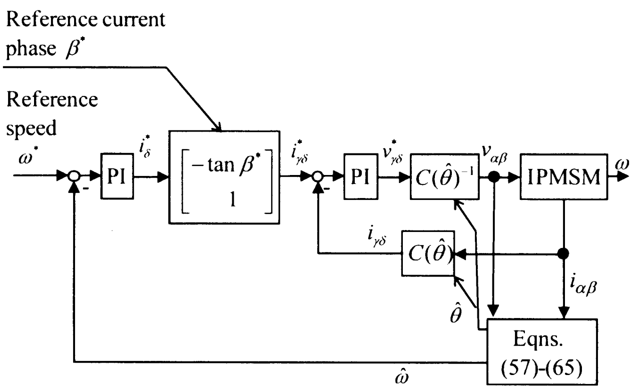


Figure 2: Control block diagram

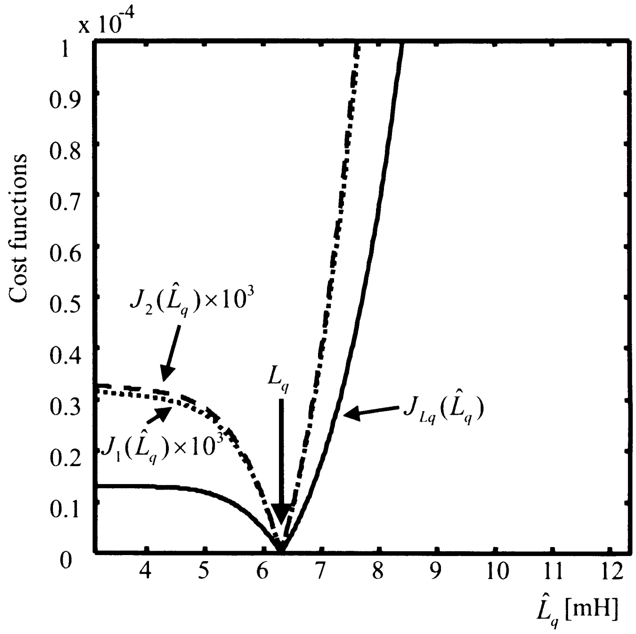
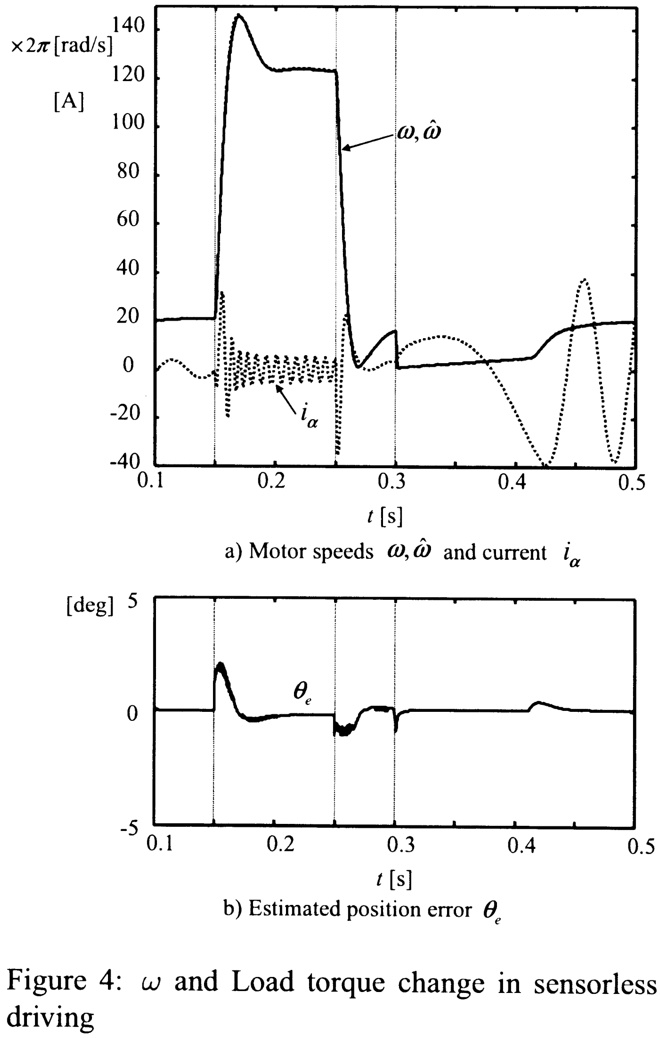


Figure 3: $\hat{L}_q$ and the cost function $J_{Lq}(\hat{L}_q)$



a) Motor speeds $\omega, \hat{\omega}$ and current i_α

b) Estimated position error θ_e

Figure 4: ω and Load torque change in sensorless driving

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