Current Control System with High-Frequency Signal Injection for Dual Three-Phase Permanent Magnet Synchronous Motor

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Dual three-phase permanent magnet synchronous motors have the characteristics of multi-phase and permanent magnet synchronous motors, whose double- and decoupled-winding models based on the vector space decomposition have been reported. High-frequency signal injection methods enable position sensorless control, parameter identification, and the search for maximum torque per ampere operating points. However, the current control system with high-frequency signal injection for a dual three-phase permanent magnet synchronous motor has not yet been sufficiently discussed. This study compares the double- and decoupled-winding models for the current control system with the high-frequency signal injection and proposes a current control system with high-frequency signal injection. Furthermore, this study proposes a position sensorless control method at low speed based on the high-frequency voltage injection as an application of the proposed current control system with the high-frequency signal injection. Experiments verify the effectiveness of the proposed current control system with the high-frequency signal injection and the proposed position sensorless control method.

Keywords: dual three-phase motor, permanent magnet synchronous motor, high-frequency signal injection, position sensorless control,

1. Introduction

A permanent magnet synchronous motor (PMSM) is an electric motor that is composed of a rotor consisting of a permanent magnet and three-phase stator windings generating rotating magnetic fields. PMSMs have certain advantages such as a small size, a high output density, and high efficiency; therefore, they are widely used in industry, domestic appliances, and mobility. Because PMSMs are expected to be applied in electric ships, high-power industrial applications, and electric aircraft, multi-phase PMSMs with more than three-phase stator windings are gaining attention. This is because multiphase stator windings enable the splitting of the motor power, higher power, and fault tolerance. There are diverse types of multi-phase PMSMs, such as five-phase, seven-phase, and dual three-phase PMSMs. The PMSM drive system requires an inverter whose phase number is equal to the stator phase number; hence, a five-phase PMSM requires a five-phase inverter, whereas a seven-phase PMSM requires a seven-phase inverter. However, although a dual three-phase PMSM (DTP-PMSM), which has two sets of three-phase stator windings, is a six-phase PMSM, DTP-PMSM can be driven comparatively easily by two three-phase inverters, not by a six-phase inverter.

Therefore, many studies have evaluated DTP-PMSM for the control of multi-phase PMSM. There are two approaches, as shown in Figs. 1 and 2 in terms of the mathematical modeling of the DTP-PMSM in previous studies. The first approach is the use of the double-winding model. Using this approach, as shown in Fig. 1, by applying the separate Park-Clark transformation into each of the two sets of three-phase windings, two dq-axis coordinates are obtained and controlled by two independent current controllers. Although this approach is based on the physical characteristics of the DTP-PMSM, it is necessary to apply decoupling

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control against mutual interference between the A-set and B-set. As shown in Fig 2, the second approach is the use of a decoupled-winding model based on the vector space decomposition (VSD)\(^{(25-27)}\). By applying VSD, two sets of dq-axis coordinates are transformed into a set that contains the fundamental component and a set that contains certain order harmonics, which results in decoupling the mutual interference.

When considering the control systems for the DTP-PMSM, in addition to the aforementioned discussion of the current control systems, such as the three-phase PMSM, it is necessary to consider the position sensorless control\(^{(18-21)}\), parameter identification\(^{(24-26)}\), and maximum torque per ampere (MTPA) control\(^{(29)}\). For example, the literature reports position sensorless control methods for the DTP-PMSM\(^{(25-29)}\). These control methods are realized by equipping the current control system with high-frequency signal injection.

First, the injected high-frequency signal is an inevitable disturbance to the current control system for added values such as the position estimation, which considerably deteriorates the current control performance. Previous studies on the three-phase PMSM considered solutions based on filters\(^{(18-20)}\) or the motor mathematical model\(^{(23)}\). This study focuses on the possibility of constructing two types of the current control system based on the “double-winding model” and the “decoupled-winding model” by taking advantage of the degree of freedom due to the multi-phase stator windings.

In other words, this study considers the current control systems with the high-frequency signal injection for both the double- and decoupled-winding models and proposes a current control system with the high-frequency signal injection. Furthermore, the study proposes a position sensorless control method at low speed based on the high-frequency voltage injection method, whose transient response is superior to those in the literature\(^{(25-27)}\) as an application of the proposed current control system with high-frequency signal injection.

In previous studies, there have been no comparative analyses of the current control systems with the high-frequency signal injection based on both the double- and decoupled-winding models. Therefore, the comparative analyses, the proposal of the current control system with the high-frequency signal injection based on the analytical results, and the proposal of the low-speed position sensorless control method by the high-frequency voltage injection are all novel and original in this study.

This paper is divided into the following sections. Sections II and III consider the double-winding model current control system and the decoupled-winding model current control system with the high-frequency signal injection, respectively. They clarified the relationship between the signal injection methods and torque ripples on each model and the influence of the high-frequency signal injection method on the current control system. Based on this discussion, Section III proposes a current control system with high-frequency signal injection. Section IV proposes a high-frequency voltage injection method as an application for the position sensorless control. Section V verifies the effectiveness of the proposed current control system with the high-frequency signal injection and proposed position sensorless control. Lastly, Section VI summarizes the conclusion of the study.

2. Double-winding model current control system with high-frequency signal injection

2.1 Double-winding model

Fig.3 shows the equivalent circuit of the double-winding model on the rotational reference when the resistance is omitted. There are two sets of stator windings: sets A and B. Sets A and B set voltages on...
the dq-axis as $u_{dqA} = [v_{dA}, v_{qA}]^T$, and $u_{dqB} = [v_{dB}, v_{qB}]^T$, respectively, and $A$- and $B$-set currents on the dq-axis as $i_{dqA} = [i_{dA}, i_{qA}]^T$, and $i_{dqB} = [i_{dB}, i_{qB}]^T$, respectively. Based on the equivalent circuit, the double-winding model of the DTP-PMSM is shown in Eq.(1) \(^{(6/12/17)}\). The parameters in Eq.(1) are listed in Table 1 under the assumption that the number of windings of the two sets are similar \((R = R_A = R_B = L_{dA} = L_{dB} = L_{qA} = L_{qB})\). In reality, the motor parameters of the two sets might not match completely. However, provided it is designed to have the same number of turns with the same material, the parameters of the two sets are assumed to be equivalent. In fact, previous studies are also based on the same assumption \(^{(6/13/17)/29}\). The torque equation of the double-winding model is given by Eq.(2).

2.2 Signal injection methods on double-winding model

In high-frequency signal injection methods, a high-frequency current is excited by injecting a high-frequency voltage into any axis of both sets A and B. Therefore, a high-frequency current generated by the own set is opposite to that of the other set. Consequently, the torque ripple is greater than that under the one-set injection because the phase of the component excited by the own set is opposite to that of the other set. However, the torque ripple in the in-phase injection is greater than that in the one-set injection if the amplitude of the injected voltage is the same. The in-phase $q$-axis injection generates a larger torque ripple than the in-phase $d$-axis injection owing to the relationship between the magnet torque and reluctance torque, similar to the one-set injection. However, the opposite phase injection is preferable for in-phase injection from the perspective of torque ripple suppression.

The opposite phase injection injects high-frequency voltages with the same phase into any axis of both sets A and B. The opposite phase injection injects high-frequency currents with opposite phase into both sets A and B. The amplitudes of the high-frequency currents in the two sets are equal in the same manner as in the in-phase injection. By applying the high-frequency voltage with the opposite phase, the components excited by the own set and by the other set have the same phase, which results in a higher amplitude of the high-frequency current than the one-set injection. This implies that the high-frequency voltage amplitude can be decreased compared with the one-set injection. Additionally, by applying the opposite phase high-frequency voltage into the A and B sets, the high-frequency current phases in the two sets are opposite. Consequently, the high-frequency components of the torque are canceled by each other, which suppresses the torque ripples. Therefore, the opposite phase injection simultaneously enables torque ripple suppression and high-frequency signal injection. This is realized regardless of the injected axis; both the opposite phase $d$-axis injection and opposite phase $q$-axis injection can suppress the torque ripples. This means that the opposite phase injection has a degree of freedom, as opposed to the one-set and in-phase injection.

2.3 Double-winding model current control system with signal injection

The high-frequency currents fed back to the current controllers deteriorate the current control of the fundamental component. The double-winding model current control system is then equipped with band elimina-
Table 2. Torque ripple due to signal injection on double-winding model

| Type                  | Name          | Injected voltage $v_{ABh}$ | High-frequency current $i_{ABh}$ | Torque ripple $T_h$                                      |
|-----------------------|---------------|---------------------------|----------------------------------|--------------------------------------------------------|
| A-set injection       |               | $[0, 0, 0]$               | $\begin{bmatrix} \frac{L_{d}v_{h}}{(L_{d}^2 - M_{d}^2)p} \\ 0 \\ 0 \end{bmatrix}$ | $(L_{d} - L_{q} + M_{d} - M_{q})i_{q} \frac{v_{h}}{(L_{d} + M_{d})p}$ |
| One-set injection     |               | $[0, 0, 0]$               | $\begin{bmatrix} 0 \\ 0 \\ 0 \end{bmatrix}$ | $(L_{d} - L_{q} + M_{d} - M_{q})i_{q} \frac{v_{h}}{(L_{d} + M_{d})p}$ |
| B-set injection       |               | $[0, 0, 0]$               | $\begin{bmatrix} 0 \\ 0 \\ 0 \end{bmatrix}$ | $(L_{d} - L_{q} + M_{d} - M_{q})i_{q} \frac{v_{h}}{(L_{d} + M_{d})p}$ |
| In-phase injection    |               | $[v_{h}, v_{h}, 0]$       | $\begin{bmatrix} \frac{L_{d}v_{h}}{(L_{d}^2 - M_{d}^2)p} \\ 0 \\ 0 \end{bmatrix}$ | $2(L_{d} - L_{q} + M_{d} - M_{q})i_{q} \frac{v_{h}}{(L_{d} + M_{d})p}$ |
| Two-set injection     |               | $[0, 0, 0]$               | $\begin{bmatrix} 0 \\ 0 \\ 0 \end{bmatrix}$ | $2(K_c + (L_{d} - L_{q} + M_{d} - M_{q})i_{d}) \frac{v_{h}}{(L_{d} + M_{d})p}$ |
| Opposite phase injection |             | $[v_{h}, -v_{h}, 0]$     | $\begin{bmatrix} 0 \\ 0 \\ 0 \end{bmatrix}$ | $0$ |

The transfer function from the current command to the actual current is expressed by the following equation: It should be noted that $\zeta$ is a filter coefficient (elimination bandwidth) that determines the gain of the center frequency $\omega_c$.

$$i_{dqAB} = \frac{P(s)C(s)}{1 + P(s)C(s)H(s)}i_{dAB}^{*} + \frac{P(s)}{1 + P(s)C(s)H(s)}v_{h}(6)$$

$P_f$ is the filter coefficient (cut-off depth) that determines the sharpness of the cut-off characteristic, and $d_f$ is the filter coefficient (cut-off depth) that determines the gain of the center frequency $\omega_c$.
\[
\begin{align*}
P(s) &= \frac{1}{L_d s + R} C(s) = \frac{\omega_c L_d s + \omega_c R}{s} \\
H(s) &= \frac{s^2 + 2\zeta\omega_h s + \omega_h^2}{s^2 + 2\omega_c s + \omega_c^2}.
\end{align*}
\]

Hence, the closed-loop transfer function from the current command to the actual current is as follows:

\[
W(s) = \frac{P(s)C(s)}{1 + P(s)C(s)H(s)} \frac{\omega_c s^2 + 2\zeta\omega_h \omega_c s + \omega_h^2 \omega_c}{s^3 + (2\omega_h + \omega_c) s^2 + (\omega_h^2 + 2\zeta\omega_h \omega_c) s + \omega_h^2 \omega_c}.
\]

(7)

The characteristic equation of the current control system including BEF is

\[
s^3 + (2\omega_h + \omega_c) s^2 + (\omega_h^2 + 2\zeta\omega_h \omega_c) s + \omega_h^2 \omega_c = 0
\]

(8)

The poles of this characteristic equation under the condition that \(\zeta = 1.0, \delta_f = 0, f_b = 100 - 1000 \text{ Hz}, \omega_h = 2\pi f_b, \) and \(\omega_c = 2000 \text{ rad/s}\) are illustrated in Fig.5. In Fig.5, the real root is distant from the right half-plane. The complex conjugate roots are near the right half-plane, which is related to destabilization. Furthermore, as \(\omega_h\) decreases, they approach the right half-plane. In the high-frequency signal injection methods, the injected frequency needs to be set lower to guarantee the amplitude of the high-frequency current because the amplitude is inversely proportional to the injected frequency. However, Fig.5 illustrates that the lower injected frequency leads to further destabilization. Thus, the BEF process deteriorates the control performance of the current control system.

A simulation was carried out to confirm the current control performance deterioration. Fig.6 shows the A/B-sets currents after the BEF processing when applying a 10 A current step into the q-axis of the A and B sets. The rotation speed was 50 rpm, the cut-off frequency of the current controller was 2000 rad/s, the cut-off frequency of the BEF was 800 Hz, and the cut-off width \(\zeta = 1.0\). Fig.6 shows that the overshoot reached a maximum of 1.28 times of the command. Thus, the q-axis current transient response of the current step deteriorates.

Therefore, in the double-winding model current control system with the high-frequency signal injection, the BEF, which is indispensable to the current feedback, deteriorates the current response. Additionally, the control system design needs to consider the current response deterioration, which becomes complicated.

3. Decoupled-winding model current control system with high-frequency signal injection

3.1 Decoupled-winding model

Based on the vector space decomposition(VSD)\(^{(10)}\), a six-dimensional motor model can be decomposed into three orthogonal coordinate systems: the \(a\beta\)-axis coordinate system, \(z_1z_2\)-axis coordinate system, and \(o_1o_2\)-axis coordinate system, as expressed in Eq.(15).

\[
\begin{bmatrix} F_a & F_\beta & F_{z1} & F_{z2} & F_{o1} & F_{o2} \end{bmatrix}^T = T_6 \cdot \begin{bmatrix} F_{aA} & F_{aA} & F_{aA} & F_{aB} & F_{aB} \end{bmatrix}^T \cdots (15)
\]

where \(F\) represents a variable. By applying the VSD-based transformation, the harmonic components are mapped onto different coordinate systems. The fundamental and \((12k \pm 1)\)th, \(k=1,2,3\) harmonic components are mapped onto the \(a\beta\)-axis coordinate system, the \((12k-6\pm 1)\)th, \(k=1,2,3\) harmonic components are mapped onto the \(z_1z_2\)-axis coordinate system, and the \((3(2k-1)-1)\)th, \(k=1,2,3\) harmonic components are mapped onto the \(o_1o_2\)-axis coordinate system.

Transforming by the following rotational matrix, variables on the \(a\beta\)-axis coordinate are mapped onto the rotational coordinate system \((dqS\)-axis coordinate system), which is synchronous with the rotor of the DTP-PMSM.

\[
\begin{bmatrix} F_{ds} & F_{dq} \\ F_{qS} & F_{qP} \end{bmatrix} = T_{dqS} \cdot \begin{bmatrix} F_a & F_\beta \end{bmatrix} \quad T_{dqS} = \begin{bmatrix} \cos \theta_{ce} & \sin \theta_{ce} \\ -\sin \theta_{ce} & \cos \theta_{ce} \end{bmatrix} \cdots (17)
\]

Additionally, the rotational transformation matrix below maps the variables on the \(z_1z_2\)-axis coordinate onto another rotational coordinate system \((dqD\)-axis coordinate system), which is synchronous with the rotor of the DTP-PMSM.

\[
\begin{bmatrix} F_{ds} & F_{dq} \\ F_{qS} & F_{qP} \end{bmatrix} = T_{dqD} \cdot \begin{bmatrix} F_o \end{bmatrix} \quad T_{dqD} = \begin{bmatrix} \cos \theta_{ce} & \sin \theta_{ce} \\ -\sin \theta_{ce} & \cos \theta_{ce} \end{bmatrix} \cdots (18)
\]

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The relationship between the double- and decoupled-winding models is discussed. The transformation matrix from the double-winding model in Eq.(1) to the decoupled-winding model in Eq.(9) is as follows;

\[
T_{VSD} = \begin{bmatrix} 1 & 0 & 1 & 0 \\ 0 & 1 & 0 & 1 \\ -1 & 0 & 1 & 0 \\ 0 & -1 & 0 & 1 \end{bmatrix}
\]

The transformation matrix from the decoupled-winding model in Eq.(9) to the double-winding model in Eq.(1) is as follows;

\[
T_{VSD}^{-1} = \begin{bmatrix} 1 & 0 & -1 & 0 \\ 0 & 1 & 0 & -1 \\ 1 & 0 & 1 & 0 \\ 0 & 1 & 0 & 1 \end{bmatrix}
\]

It should be noted that the coefficients in Eqs.(19) and (20) are different, which treats the current and voltage invariably between the double- and decoupled-winding models.

### 3.2 Signal injection methods on decoupled-winding model

When injecting the high-frequency voltage \(v_{sdh} = [v_{dsh}, v_{qsh}, v_{dDh}, v_{qDh}]^T\), the decoupled-winding model high-frequency current equation is expressed as Eq.(12). The torque equation is then expressed as shown in Eq.(14). The decoupled-winding model signal injection methods include summation- and difference-set injections. Although they inject one set of the summation- or difference-sets, they physically inject high-frequency signals into both sets A and B, which are classified into two-set injections. Table 3 summarizes the signal injection methods on the decoupled-winding model. Based on the table, this section discusses the summation- and difference-set injections.

The summation-set injection injects a high-frequency voltage into any axis of the summation-set. It only generates the summation-set current because the summation- and difference-sets are independent in the decoupled-winding model. The summation-set current contributes to the output torque; hence, the torque ripple is generated by any injected...
Table 3. Torque ripple due to signal injection on decoupled-winding model

| Type                        | Name                             | Injected voltage $v_{S_Dh}$ | High-frequency current $i_{S_Dh}$ | Torque ripple $T_h$ |
|-----------------------------|----------------------------------|-----------------------------|-----------------------------------|---------------------|
| Summation-set injection     |                                  | $v_h$                       | $i_{S_Dh}$                        | $2(L_{ds} - L_{qs})i_{qS}v_h$                                                                 |
| Two-set injection           |                                  | $v_h$                       | $i_{S_Dh}$                        | $2(K_e + (L_{ds} - L_{qs})i_{dS})v_h$                                                      |
| Difference-set injection    |                                  | $v_h$                       | $i_{S_Dh}$                        | $0$                                                              |

The difference-set injection injects the high-frequency voltage into any axis of the difference set. It only generates the difference-set current owing to the independence of the summation- and difference-sets. The difference-set current fundamental component is controlled to 0 A, and it does not generate torque ripple regardless of the injected axis. Hence, the difference-set injection has the degree of freedom of the signal injection.

Therefore, as a signal injection method for the decoupled-winding model, the difference-set injection is superior to the summation-set injection from the perspective of the torque ripple suppression.

3.3 Decoupled-winding model current control system with signal injection Fig.8 shows the decoupled-winding model current control system when applying the difference-set injection. As shown in the figure, while the difference-set current control system necessitates the BEF to eliminate the high-frequency current contained in the motor output current, the summation-set current control system does not need the BEF. This is because the summation-set is independent of the difference-set, and there is no inter-set interference. Hence, the high-frequency current does not appear in the summation-set, and it is possible to avoid the transient response deterioration due to the BEF. Although the transient response in the difference-set deteriorates owing to the BEF, it does not lead to any serious problems because the difference-set current command is set to 0 A and the difference set is not used for the transient response. Thus, it is unnecessary to design the filter by considering the transient performance deterioration, which simplifies the control system design in the decoupled-winding model current control system with high-frequency signal injection.
Therefore, we propose a current control system with a difference-set high-frequency signal injection as the current control system for a DTP-PMSM with high-frequency signal injection.

A simulation was performed to verify the performance of the proposed current-control system. Fig. 9 shows the summation- and difference-sets feed-back current when applying the 10 A current step into the q-axis of the summation-set. The simulation conditions were the same as in Section II. The rotation speed was 50 rpm, the cut-off frequency of the current controller was 2000 rad/s, the cut-off frequency of the BEF was 800 Hz, and the cut-off current controller was 2000 rad/s. The simulation conditions were the same as in Section III, the opposite phase d-axis injection in the double-winding model does not appear, which confirms the stable transient response. Therefore, the proposed decoupled-winding current control system enables a stable current control performance.

3.4 Comparison between double-winding model and decoupled-winding model

Three points summarize the comparison between the double- and decoupled-winding models as the current control system with the high-frequency signal injection in Sections II and III.

- Both models can suppress the torque ripple due to the high-frequency current. The opposite phase injection in the double-winding model and difference-set injection in the decoupled-winding model can suppress the torque ripple. Additionally, both the d-axis and q-axis injection suppress the torque ripple; hence, there is a degree of freedom in the choice of the injected axis.

- The decoupled-winding model is superior in terms of the current control performance. In the double-winding model, both sets A and B require a BEF to eliminate the high-frequency current from the current control feedback loop. In the decoupled-winding model, however, the BEF is unnecessary in the summation-set, which plays a role in controlling the fundamental current because inter-set interference does not occur.

- It is easier to design the control system of the decoupled-winding model. In the double-winding model, it is necessary to consider the transient response deterioration when designing the current control system. In contrast, in the decoupled-winding model, the BEF is unnecessary in the summation-set, which simplifies the design of the current control system.

Hence, the decoupled-winding model is superior to the double-winding model in terms of the signal injection in a DTP-PMSM. Therefore, the proposed current control system with the high-frequency signal injection is suitable for a DTP-PMSM.

4. Application into position sensorless control

Sections II and III compare the double- and decoupled-winding models for the current control system with the high-frequency signal injection and propose a current control system with the high-frequency signal injection suitable for the DTP-PMSM. This section proposes a position sensorless control method with a stable transient response as an application example of the proposed current control system with the high-frequency signal injection.

4.1 High-frequency current equation on estimated rotational coordinate

Transforming the estimated rotational coordinates after applying the high-frequency voltage into Eqs. (1) and (9), the high-frequency current equations on the estimated rotational coordinate are obtained as Eqs. (21) and (22). \( \Delta \theta \) represents the position estimation error, and \( \dot{\theta}_{re} \) represents the estimated position, where \( \Delta \theta = \dot{\theta}_{re} - \dot{\theta}_{re} \). The following discussion assumes the injection of a sinusoidal voltage \( V_h \sin(\omega_h t) \), where \( V_h \) is the amplitude and \( \omega_h \) is the injected frequency. Based on the conclusions in Sections II and III, the opposite phase d-axis injection in the double-winding model and the difference-set d-axis injection in the decoupled-winding model are applied to suppress the torque ripple.

When applying the opposite phase d-axis injection in the double-winding model, the high-frequency voltage is \( v_{ySBH} = [V_h \sin(\omega_h t), 0, -V_h \sin(\omega_h t), 0]^T \), and the high-frequency current formulas are represented as follows:

\[
\begin{align*}
{i_y}_{ABh} &= -\left( K_1 - K_2 \right) \sin \left( 2 \Delta \theta \right) \frac{V_h \cos(\omega_h t)}{K_{\omega h}} + \left( K_2 - K_4 \right) \cos \left( 2 \Delta \theta \right) \frac{V_h \cos(\omega_h t)}{K_{\omega h}} \\
{i_d}_{ABh} &= \left( K_2 - K_4 \right) \sin \left( 2 \Delta \theta \right) \frac{V_h \cos(\omega_h t)}{K_{\omega h}} \\
{i_y}_{BHh} &= \left( K_1 - K_3 \right) + \left( K_2 - K_4 \right) \cos \left( 2 \Delta \theta \right) \frac{V_h \cos(\omega_h t)}{K_{\omega h}} \\
{i_d}_{BHh} &= -\left( K_2 - K_4 \right) \sin \left( 2 \Delta \theta \right) \frac{V_h \cos(\omega_h t)}{K_{\omega h}}
\end{align*}
\]

The high-frequency current on the estimated rotational coordinates is generated in both the \( \gamma \)-axis and \( \delta \)-axis in the A and B sets, as expressed in the equations above, which necessitates filters in all axes of the current control system.

In contrast, when applying the difference-set d-axis injection in the decoupled-winding model, the high-frequency voltage is \( v_{ySDh} = [0, 0, V_h \sin(\omega_h t), 0]^T \), and the high-frequency current formulas are presented as follows:

\[
\begin{align*}
{i_y}_{SBh} &= 0 \\
{i_d}_{SBh} &= 0 \\
{i_y}_{BDh} &= -\frac{L_{\omega D} + L_{1D} \cos \left( 2 \Delta \theta \right)}{L_{1D}^2 - L_{1D}^2 \omega_h} \frac{V_h \cos(\omega_h t)}{K_{\omega h}} \\
{i_d}_{BDh} &= -\frac{L_{1D} \sin \left( 2 \Delta \theta \right)}{L_{1D}^2 - L_{1D} \omega_h} \frac{V_h \cos(\omega_h t)}{K_{\omega h}}
\end{align*}
\]

As the equations above show, the high-frequency currents on the estimated rotational coordinate are generated in the difference-set, but not in the summation-set. Thus, as discussed in Section III, the summation-set current control system in the decoupled-winding model does not require any filters.

In both the double- and decoupled-winding models, the \( \gamma \)-axis high-frequency current contains offset terms in addition to the terms correlated to the position error, whereas the \( \delta \)-axis high-frequency current contains only terms correlated to the position error. Hence, the \( \delta \)-axis current can be used when injecting a high-frequency voltage into the \( d \)-axis.

4.2 Position estimation system

Fig. 10 illustrates the configuration of the position estimation system. In the position estimation system, by applying the heterodyne process to the high-frequency current, the correlational signals expressed in Eqs. (30) and (31) are obtained. The phase lock
\[
\begin{align*}
[i_{AB}] & = \frac{1}{K_p} \begin{bmatrix} K_1 + K_2 \cos (2 \Delta \theta) & -K_2 \sin (2 \Delta \theta) & K_3 + K_4 \cos (2 \Delta \theta) & -K_4 \sin (2 \Delta \theta) & -K_3 \sin (2 \Delta \theta) & K_4 \cos (2 \Delta \theta) \end{bmatrix} \cdot \begin{bmatrix} i_{YAB} \\ i_{YAB} \\ i_{YB} \\ i_{YB} \end{bmatrix} \\
\end{align*}
\]

\[
K = -L_0^3 + 2 L_2 L_1^2 + 2 L_0^2 M_0^2 + 2 L_0^2 M_1^2 - 8 L_0 L_1 M_0 M_1 - L_1^4 + 2 L_1^2 M_0^2 + 2 L_1^2 M_1^2 - M_0^4 + 2 M_0^2 M_1^2 - M_1^4
\]

\[
L_0 = \frac{L_0 - L_1}{2}, \quad L_1 = \frac{L_0 + L_1}{2}, \quad M_0 = \frac{M_0 + M_1}{2}, \quad M_1 = \frac{M_0 - M_1}{2}
\]

\[
\begin{align*}
[i_{SH}] & = \begin{bmatrix} L_{SH} \sin(2 \Delta \theta) \\ L_{SH} \sin(2 \Delta \theta) \\ L_{SH} \sin(2 \Delta \theta) \\ L_{SH} \sin(2 \Delta \theta) \end{bmatrix} \\
[i_{DH}] & = \begin{bmatrix} L_{DH} \sin(2 \Delta \theta) \\ L_{DH} \sin(2 \Delta \theta) \\ L_{DH} \sin(2 \Delta \theta) \\ L_{DH} \sin(2 \Delta \theta) \end{bmatrix}
\end{align*}
\]

\[
L_{DS} = \frac{L_{DS} + L_{qS}}{2}, \quad L_{IQ} = L_{DS} - L_{qS}, \quad L_{OD} = \frac{L_{OD} + L_{qD}}{2}, \quad L_{ID} = \frac{L_{OD} - L_{qD}}{2}
\]

\[
\begin{align*}
\omega_r & = K_p \frac{i_{SH}}{s} + \frac{1}{s} \frac{d \theta_r}{dt} \\
\end{align*}
\]

4.3 Position sensorless current control system  Figs. 11 and 12 show the position sensorless current control system based on the double- and decoupled-winding models, respectively. The double-winding model position sensorless current control system shown in Fig.11 is presented in (26) and the feedback current requires the BEF because the high-frequency current is generated in both the A and B sets. In contrast, the decoupled-winding model position sensorless current control system shown in Fig.12 requires the BEF only in the difference-set. This is because the difference-set injection generates a high-frequency current not in the summation-set, but only in the difference-set. Therefore, as discussed in Section III, the decoupled-winding model position sensorless current control system is superior to the counterpart of the double-winding model in terms of the transient characteristics.

The decoupled-winding model position sensorless current control system shown in Fig.12 is the proposed method in this study. The proposed method has two characteristics compared with (26)–(29). First, the proposed method can suppress the torque ripple due to the high-frequency current, similar to the methods presented in (26)–(29). Second, the proposed method has a stable transient characteristic, as opposed to the methods in (26)–(29). Previous studies (26–29) applied position sensorless control in the double-winding model, necessitating filters in both the A and B sets to eliminate the high-frequency components from the feedback currents. A previous study (29) did not mention the overall position sensorless current control system, although it applies position sensorless control in the decoupled-winding model. According to the transient response experimental result, however, there are overshoots, which imply that it uses a filter even in the summation-set. Hence, the stable transient characteristic of the method proposed in this study is posited as a novel benefit.

Therefore, the proposed position sensorless control can achieve torque ripple suppression and a stable transient response, in addition to the position estimation.

4.4 Adaptable speed range of high-frequency voltage injection method  The position sensorless control method based on the high-frequency voltage injection is widely known to be available at low speeds. This study also derives the high-frequency current equations expressed by Eqs.(3) and (12) by approximating the resistance and speed terms as 0. This subsection discusses the speed adaptability of the proposed low-speed position sensorless control method.

The high-frequency current equation of the decoupled-winding model considering the resistance and speed terms is expressed as follows:

\[
\begin{align*}
\end{align*}
\]
By transforming into the estimated rotational coordinates, Eq.(35) can be obtained.

Applying the difference-set d-axis injection, the difference-set δ-axis current is used as the correlation signal.

\[
i_{D\delta}(\Delta \theta) = \frac{L_{1D} \sin(2\Delta \theta) s - \omega_r L_{1D} - \omega_r L_{1D} \cos(2\Delta \theta)}{K_D} v_{Dh}
\]

Approximating \( \Delta \theta \approx 0 \), the following equation is obtained.

\[
i_{D\delta}(\Delta \theta) = \frac{2L_{1D} s - \omega_r L_{1D} - \omega_r L_{1D}}{K_D} v_{Dh}
\]

\[
= \frac{2L_{1D} s - \omega_r L_{1D}}{K_D} v_{Dh} \left( \Delta \theta - \frac{\omega_r L_{1D}}{2L_{1D}s} \right) \cdot (36)
\]

Substituting \( s = j \omega_h \),

\[
i_{Dh}(\Delta \theta) = \frac{2L_{1D} j \omega_h}{K_D} v_{Dh} \left( \Delta \theta - \frac{\omega_r L_{1D}}{2L_{1D}j \omega_h} \right) \cdot (37)
\]

When the position estimator converges to the phase of \( i_{Dh} = 0 \), the constant estimation error \( \Delta \theta_{\text{const}} \) occurs as follows.

\[
\Delta \theta_{\text{const}} = \frac{L_{1D} \omega_r}{2 \omega_h L_{1D}} = -\frac{L_{1D} \omega_r}{\omega_h (L_{1D} - L_{eqD})} \quad \ldots \ldots \quad (37)
\]

Eq.(37) implies that this constant error is proportional to the speed and difference-set d-axis inductance and is inversely proportional to the injected frequency and difference-set saliency.

To verify the validity of the derived constant estimation error, Eq.(37), a simulation was carried out. Calculating the speed that makes the estimation error \( -5 \) deg under several injected frequencies and inductance using Eq.(37), simulation confirmed the estimation error under the speed. In this simulation, the initial speed was 10 rpm, and the acceleration rate was 100 rpm/s from 1.5 s.

Fig.13 and Table 4 present the simulation wave forms and results, respectively. Table 13 expresses the injected frequency, the rate of the set value against the nominal value \( L_{eqD} \), the mechanical angular speed that makes \( \Delta \theta_{\text{const}} = -5 \) deg calculated by Eq.(37), and the simulation estimation error under the speed in Fig.13, from the left.

The simulation results show that the estimated errors at the speed calculated from Eq.(37) were \( -5 \) to \( -8 \) deg, which confirms that the theoretical equation approximately matches the simulation results. The difference from the theoretical equation is inferred to be caused by the approximation of \( \Delta \theta \approx 0 \).

Therefore, it is confirmed that the estimation error due to the speed depends on the inductance and injected frequency, as shown in Eq.(37). In practice, the estimation error is based on Eq.(37) and determines the adaptable speed range of the high-frequency voltage injection method.

The constant error of the decoupled-winding model is derived in this study; however, that of the double-winding model can be derived in the same way. In the double-winding model, the constant error due to the application of the opposite phase d-axis injection is the same as in Eq.(37).

5. Experiments

Experiments were performed to verify the effectiveness of the proposed current control system with the high-frequency signal injection and proposed position sensorless control for low speed. In the following experiments, instead of the DTP-PMSM, a dual three-phase wound-field synchronous motor was used. Although the target motor is different from the
The torque command was 7.58 Nm in this experiment because it is symmetrical with the reluctance torque. Compared with the injection method. This is because the salient motor has a larger magnet torque than the reluctance torque. The opposite phase d-axis injection, the in-phase injection and summation-set injection suppressed the torque ripples in both the d-axis and q-axis injections. However, it was confirmed that the opposite phase injection and summation-set injection doubled the torque ripple. Hence, from the perspective of the torque ripple suppression, the d-axis injection is preferable to the q-axis injection; particularly, the opposite phase injection and summation-set injection are appropriate as the signal injection methods.

### 5.2 Transient response experiment

The next experiment compared the current control systems with the high-frequency signal injection. The opposite phase d-axis injection was applied as the double-winding model signal injection and the summation-set d-axis injection as the decoupled-winding model signal injection; the 40 A current step under the MTPA control was applied at 0.05 s.

The experimental results are shown in Figs. 15 and 16. Fig. 15 shows the A/B-set feedback current after processing the BEF and the output torque. Fig. 16 shows the summation-set feedback current, the difference-set feedback current af-

Table 5. Experimental condition

| Parameter                  | Value         |
|----------------------------|---------------|
| DC link voltage            | $V_{DC}$ 12 V |
| Inverter carrier frequency | $f_c$ 10 kHz  |
| Winding resistance of rotor| $R_f$ 2.075 Ω |
| Self-inductance of rotor   | $L_f$ 0.24 H  |
| Mutual-inductance of rotor | $M_f$ 2.90 mH |
| Filed current              | $i_f$ 4.0 A   |
| Rotation Speed             | $\omega_{tm}$ 50 rpm (6.67 Hz) |
| Injected frequency         | $f_h$ 800 Hz  |
| Injected voltage           | $V_h$ 4 V     |
| Cut-off width of BEF       | $\zeta$ 1.0   |
| Cut-off depth              | $d_f$ 0       |

Fig. 14. Experimental result of torque ripple

5.1 Evaluation on torque ripple

The first experiment verified the influence of the signal injection methods on the torque ripple. The signal injection methods discussed in Sections II and III were applied under MTPA control with a 40 A current amplitude. The observed torque is calculated by the torque equations and currents. The B-set injection is omitted in this experiment because it is symmetrical with the A-set injection. The torque command was 7.58 Nm in this experiment.

Fig. 14 summarizes the torque ripple ratio against the torque fundamental component. According to the figure, it is confirmed that the torque ripples by the q-axis injections were greater than those by the d-axis injections in any injection method. This is because the salient motor has a larger magnet torque than the reluctance torque. Compared with the A-set injection, the in-phase injection and summation-set injection doubled the torque ripple. However, it was confirmed that the opposite phase injection and the difference-set injection suppressed the torque ripples in both the d-axis and q-axis injections. These experimental results match those in Tables 2 and 3 in Sections II and III, respectively.

PMSM in terms of the generation of the magnetic flux, it is equivalent to the DTP-PMSM under the condition of the constant field current.

The drive system of the target motor was a PE-Expert 4 (Myway Plus Corporation). The PE-Expert 4 system included a DSP board (MWPE4-C6657) that used a DSP(TMS320C6657) manufactured by Texas Instruments(TI) and the PEV board (MWPE4-PEV), which includes the A/D conversion function and gate signal output function. The voltage was supplied to the target motor by two voltage source inverters made by the Myway Plus Corporation (MWINV-7R006A) for stator windings and a voltage source inverter (MWINV-5R022) for rotor windings. The loading apparatus against the target motor was a servo motor and a servo driver manufactured by the YASKAWA Electric Corporation. In the experiments, the loading apparatus maintained the rotational speed of the target motor constant. The experimental condition is presented in Table 5.

### Table 4. Results of acceleration simulation

| $f_h$ [Hz] | $L_{QD}/L_{DQ}$ | $\omega_{rpm}$ [rpm] using Eq.(37) | Result [deg] |
|------------|-----------------|-----------------------------------|--------------|
| 800        | 1.0             | 164                               | -6.66        |
| 600        | 1.0             | 122                               | -5.07        |
| 1000       | 1.0             | 205                               | -7.69        |
| 800        | 1.1             | 232                               | -6.95        |
| 800        | 0.9             | 95                                | -6.30        |

Fig. 13. Wave forms of acceleration simulation

![Graph](image_url)
term processing the BEF, and the output torque. As shown in Fig.15, the overshoot and fluctuations were observed in the transient response at the current step when applying the signal injection method in the double-winding model, which suggests that the current control performance deteriorated. Consequently, the output torque also deteriorated with the torque overshoot. When applying the signal injection method to the decoupled-winding model, as shown in Fig.16, the summation-set current had a stable one-order delay response without the overshoot. Thus, the torque response in the last row is also a one-order delay characteristic.

Hence, the proposed current control system based on the decoupled-winding model has a more stable control performance than its counterpart based on the double-winding model.

5.3 Position sensorless control experiment Finally, the position sensorless control at a low speed based on the high-frequency voltage injection was verified experimentally. The experiments compared the A-set d-axis injection, opposite phase d-axis injection as the methods on the double-winding model, and difference-set d-axis injection as that on the decoupled-winding model.

First, the correlational signals were observed to verify the principle of the position estimation. The correlational signals were observed under MTPA control with a 10 A current command while applying the position error from \(-\pi\) to \(\pi\). Figs.17(a), 17(b), and 17(c) show the results. According to the figures, it is confirmed that the high-frequency components in both the double- and decoupled-winding models are correlated with the position error.

Secondly, the position estimation performances at the steady-state are compared. Figs.18(a), 18(b), and 18(c) show the estimation results under the MTPA control with a 20 A current command, by A-set d-axis injection, the opposite phase d-axis injection, and difference-set d-axis injection, respectively. According to the figures, any method stably estimated the position and speed. In terms of the torque ripple, while the A-set d-axis injection generated 0.29 Nm ripple (peak-to-peak), the opposite phase d-axis injection and the difference-set d-axis injection decreased to 0.01 Nm. These results support that position estimation and torque ripple suppression can be realized simultaneously by the discussed signal injection methods.

The last experiments compared the transient performance of the three methods. Figs.19(a), 19(b), and 19(c) show the transient estimation results by the A-set d-axis injection, the opposite phase d-axis injection, and the difference-set d-axis injection, respectively. According to the figures, in any method, although the transient position error reached 15 deg, it converged to 0 deg. It is confirmed that any methods stably estimated even at the transient-state. For the torque response, there were overshoots in the A-set d-axis injection method and the opposite phase d-axis injection method based on the double-winding model. The overshoots reached 4.29 Nm and 4.45 Nm, respectively, which were 29.6% and 38.1% of the input torque step. However, the different-set d-axis injection method did not result in any overshoot and exhibited a first-order delay characteristic. Therefore, the proposed high-frequency voltage injection method based on the decoupled-winding model (the proposed sensorless control) achieves stable transient performance in addition to the position estimation.

6. Conclusion

This study presents a comparison between the double- and decoupled-winding models as the current control system with high-frequency signal injection for a dual three-phase permanent magnet synchronous motor (DTP-PMSM). Based on the comparison, we proposed a current control system that injects high-frequency signals into the difference-set on the decoupled-winding model.

In terms of the torque ripple due to the high-frequency signal injection, both models can suppress the ripple similarly. This study mathematically analyzed that the opposite phase injection in the double-winding model and the difference-set injection in the decoupled-winding model realize torque ripple suppression.

However, in terms of the current control performance of the current control system with the signal injection, the decoupled-winding model is superior to the double-winding model. The double-winding model current control system necessitates BEFs in both sets A and B to eliminate the high-frequency components from the feedback currents. The decoupled-winding model current control system, however, does not necessitate the BEF in the summation-set current control feedback that controls the fundamental component.

Additionally, in terms of the control system design, the decoupled-winding model is superior to the double-winding model. This is because the decoupled-winding model current control system only needs to consider the current controller, whereas the double-winding model current control system needs to consider the transient performance deterioration due
Current Control System with High-Frequency Signal Injection for Dual Three-Phase Permanent Magnet Synchronous Motor (Koji Imai et al.)

The above three points were verified by experiments in addition to the mathematical analyses. Therefore, it is concluded that it is better to use the proposed current control system for position estimation, parameter identification, and search for MTPA operating points.

This study also proposed a position sensorless control for low speed based on the difference-set high-frequency voltage injection method as an application example of the proposed current control system. Although both the double- and decoupled-winding models can achieve position sensorless control and torque ripple suppression, the proposed method was confirmed to have stable transient characteristics experimentally.

Fig. 17. Correlational signals

Fig. 18. Estimation results at steady-state

Fig. 19. Estimation results at transient-state

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