Stationary Reference Frame Position Sensorless Control Based on Stator Flux Linkage and Sinusoidal Current Tracking Controller for IPMSM

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Generally, standard direct torque control (DTC) does not require fine current responses from the current controller because the torque and flux of a DTC-based drive are controlled by a closed-loop system without current loops. It is often difficult for the standard DTC to achieve the desired current regulation and fine harmonic current disturbance suppression performance. In order to overcome these problems, this paper proposes a new stationary reference frame position sensorless control system based on stator flux linkage estimation, and a new sinusoidal current tracking controller. In this paper, the current control performance of IPMSMs using the proposed sinusoidal current tracking controller is evaluated through numerical simulation and experiments. Finally, using the proposed sinusoidal current tracking controller, the effectiveness of the proposed stationary reference frame position sensorless control system with the stator flux control method is confirmed through experimental results.

Keywords: IPMSM, position sensorless control, disturbance suppression characteristics, stationary reference frame, sinusoidal tracking current controller

1. Introduction

Internal permanent magnet synchronous motors (IPMSMs) have been applied for motors in electric vehicles and various industrial applications. Recently, to reduce the size and cost of the motor drive systems, position sensorless control methods have been studied extensively. Direct torque control (DTC) has often used to its position sensorless control based on stator flux estimation^{(1)–(4)}. DTC has the merits of both fast torque response and low sensitivity to inductance variations. Hence, DTC has been used in many types of AC motor.

DTC has many regulation methods of stator flux and motor torque. DTC can have a fast torque response. The standard type of DTC uses both a switching table of space voltage vector and hysteresis comparator of torque and flux⁽¹⁾. Other type of DTC is called Reference Flux Vector Calculator DTC (RFVC-DTC)^{(2)–(4)}. RFVC-DTC can have a fine and smooth torque response without torque ripple. In RFVC-DTC, the torque error is converted into the torque angle information by PI torque controller. Its voltage reference is determined by the torque angle information. Generally, the standard DTC and RFVC-DTC do not use the current regulated and do not regulate the motor current directly. Since IPMSM is driven by motor current, DTC should be regulated by the motor current. Therefore, the sensorless control method having the PI current controller has been realized^{(5)–(6)}, whose name is stator flux vector control (SFVC). In this case, its current controller is located on the rotating reference frame.

Standard DTC, RFVC-DTC and SFVC have the fine torque tracking performance. However, since the standard DTC and RFVC-DTC need not have current controller having a fine current response of primary frequency, the standard DTC and RFVC-DTC sometimes can not have a fine harmonic current disturbance suppression performance.

On condition of primary frequency, its d-q current information in rotating reference frame is equal to its α - β current information in stationary reference frame⁽⁷⁾. On the other hand, in case of harmonic frequency, the d-q current control does not often have the desired performance.

Moreover, the current control performance of IPMSMs using the new sinusoidal tracking current controller is evaluated by the numerical simulation results and experimental results.

The current control in the stationary reference frame can suppress the harmonic current better than current control in the rotating reference frame. Thus, robust performance against the harmonic current is obtained by control of the stationary frame.

In order to obtain robust performance, this paper proposes that the proposed stationary reference frame position sensorless control is accomplished with a new current control system in the α - β stationary frame.

This paper confirms the validity of the proposed method by using the experimental results.

2. Sinusoidal Current Tracking Control

2.1 Nomenclature Definition diagram of coordinates is shown in Fig. 1. Here, superscripts and symbol and abbreviations are shown as follow;

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Fig. 3. Disturbance suppression characteristics using d-q PI current control

$$a_2 = \frac{R_a}{L_d} + \frac{f_3}{L_d} \quad (12)$$

$$b_0 = \frac{f_2}{L_d} \quad (13)$$

$$b_1 = \frac{f_1}{L_d} \quad (14)$$

The current control system is designed by using (9). On the basis of the poles placement method, the poles of the sinusoidal current tracking controller are designed so as to have the desired roots. The feedback gains of the sinusoidal current tracking controller are given by (15)–(17). Using these gains, the poles of the current control system are arranged in multiple roots in the band. The zero also has a stable placement. Therefore, the current control system has stable operation using the feedback gains of (15)–(17).

$$f_1 = L_d(\omega_{re}^2 - 3\omega_c^2) \quad (15)$$

$$f_2 = L_d(3\omega_c\omega_{re}^2 - \omega_c^3) \quad (16)$$

$$f_3 = 3L_d\omega_c - R_a \quad (17)$$

2.4 Evaluation based on Extended Transformation of the Disturbance Suppression Characteristics Using the extended transfer function $(9)^{(10)}$, the both disturbance suppression characteristics of d-q PI current control and α - β sinusoidal current control are evaluated in this paper by using the extended transfer function, which is introduced in Appendix of this paper. The extended transfer function can evaluate the both frequency characteristics of d-q PI regulator and α - β sinusoidal regulator on the positive and negative harmonics phase components. Here, j is the complex coefficient. The transfer function from i_{α}^{dis} to i_{α}^{res} in d-q PI regulator is given by (18).

Fig. 4. Disturbance suppression characteristics using α - β sin current control

$$G_{dq-PI}^{dis} = \frac{i_{\alpha\beta}^{res}}{i_{\alpha\beta}^{dis}} = \frac{-s^2 + n_1s + n_0}{s^2 + m_1s + m_0} \quad (18)$$

$$m_0 = \frac{K_I - j(R_a + K_P)\omega_{re}}{L_d} \quad (19)$$

$$m_1 = \frac{(R_a + K_P - jL_d\omega_{re})}{L_d} \quad (20)$$

$$n_0 = \frac{jR_a\omega_{re}}{L_d} \quad (21)$$

$$n_1 = \frac{-R_a + jL_d\omega_{re}}{L_d} \quad (22)$$

The transfer function from i_{α}^{dis} to i_{α}^{res} in the α - β sinusoidal control is given by (23).

$$G_{\alpha\beta-sin}^{dis} = \frac{i_{\alpha\beta}^{res}}{i_{\alpha\beta}^{dis}} = \frac{-s^3 - q_2s^2 - q_1s - q_0}{s^3 + a_2s^2 + a_1s + a_0} \quad (23)$$

$$q_0 = \frac{R_a}{L_d}\omega_{re}^2 \quad (24)$$

$$q_1 = \omega_{re}^2 \quad (25)$$

$$q_2 = \frac{R_a}{L_d} \quad (26)$$

Using (18) and (23), each bode diagram of d-q PI regulator and α - β sinusoidal regulator is obtained for the positive-phase disturbance suppression characteristics, as shown in Fig. 3(a) and Fig. 4(a). In Fig. 3(a) and Fig. 4(a), the fundamental frequency performance of d-q PI regulator is equivalent equal to that of α - β sinusoidal regulator. However, the harmonic disturbance suppression performance of α - β sinusoidal regulator is slightly better than that of d-q PI regulator. Comparing the two disturbance suppression characteristics, the

Fig. 5. Experimental system for evaluation on disturbance suppression performance d-q PI current regulator and α - β sinusoidal current regulator

(a) FFT analysis u-phase current responses with d-q PI regulator.

(b) FFT analysis results u-phase responses with α - β sinusoidal regulator.
Fig. 6. Experimental results of u-phase current FFT analysis

harmonic disturbance suppression performance of the α - β sinusoidal regulator is one order higher than that of the d-q PI regulator. Therefore, the harmonic disturbance suppression performance of the α - β sinusoidal regulator is slightly better compared with the d-q PI regulator.

Similarly, Fig. 3(b) and Fig. 4(b) show each bode diagram of the negative phase disturbance suppression characteristics, respectively. The negative-phase harmonic disturbance rejection performance of α - β sinusoidal regulator is better than that of d-q-PI regulator.

As the result, the effectiveness of α - β sinusoidal controller is confirmed through the experimental results of Fig. 6. The rated values and specifications of tested motor are shown in Table 1, respectively. In the experiment, the motor torque current is maintained at $i_q = 1$ A, which is 40% load. The motor speed is maintained at 100 rpm. In order to confirm the disturbance suppression characteristics, the experimental results are shown in Fig. 6(a) and Fig. 6(b). The robust performance of $\alpha\beta - \sin$ is better than dq-PI. 7th and 13th are a positive phase disturbance. Because the order of the transfer

Table 1. Rated values and specifications of tested motor

R_a	2.8 [Ω]
L_d	28 [mH]
L_q	70 [mH]
P	3
J_M	1.152×10^{-3} [kgm ²]
Carrier frequency PWM Inverter	10 [kHz]
Rated current of u-phase	3 [A]
Rated speed	1800 [rpm]
Rated power	750 [W]

Table 2. Specification of tested speed control system

Poles of the speed control system	ω_s	80 [rad/s]
Poles of the current control system	ω_c	2000 [rad/s]

Fig. 7. Speed control system using I-P speed controller

function of $\alpha\beta - \sin$ is high, $\alpha\beta - \sin$ can reduce the positive phase disturbance better than the dq-PI. 5th and 11th is a negative phase disturbance. $\alpha\beta - \sin$ can be reduced more than dq-PI because $\alpha\beta - \sin$ can be suppressed even for negative phase. From the above results, the current control system using the sinusoidal tracking controller can reduce the current harmonics better than the current control system using the PI controller. Therefore, the current control system using the sinusoidal tracking controller is expected to suppress torque ripple and motor noise generated by the structure of the winding and the space harmonic flux. The disturbance suppression characteristics of the current control system using the sinusoidal tracking controller in the modified IPMSM can also be found in the Appendix.

2.5 Experimental Results of Sinusoidal Current Tracking Controller with Position Sensor Using the α - β sinusoidal current controller, the speed control system with position sensor is constructed. The transfer function from ω_{rm}^{ref} to ω_{rm}^{res} is given by (27).

$$G_{refres} = \frac{\omega_{rm}^{res}}{\omega_{rm}^{ref}} = \frac{K_{Is}}{J_m s^2 + K_{Ps} s + K_{Is}} \quad (27)$$

where K_{Ps} and K_{Is} represent the proportional and integral gains of the speed controller. The speed controller is an I-P type controller. The speed control system is shown in Fig. 6. K_{Ps} and K_{Is} are calculated by the use of (27) and the pole placement method. Using (9) and (27), the specifications of the tested current control system and the tested speed control system are designed as shown in Table 2.

Figure 8 shows the experimental results of speed control system based on α - β sinusoidal current controller with position sensor. The proposed speed control system based on α - β sinusoidal current controller has good responses.

3. Position Sensorless Control based on $\alpha\beta$ -sin Current Controller

3.1 Sensorless Control Method of Torque (Method 1)

Here, the principle of proposed position sensorless control is presented in Fig. 9. The motor torque of IPMSM is defined

Fig. 8. Experimental results of speed step response using speed control system based on sinusoidal current tracking controller with position sensor

Fig. 9. Coordinate of position-sensorless control in stationary reference frame

by (28). The stator flux linkage vector is estimated by using (29) and (30) ⁽¹¹⁾⁽¹²⁾.

$$T = P\psi * i^{ref} = P(\psi_\alpha i_\beta - \psi_\beta i_\alpha) \quad (28)$$

$$\hat{\psi}_\alpha = \int_0^t (v_\alpha^* - R_a i_\alpha) dt + \hat{\psi}_\alpha|_{t=0} \quad (29)$$

$$\hat{\psi}_\beta = \int_0^t (v_\beta^* - R_a i_\beta) dt + \hat{\psi}_\beta|_{t=0} \quad (30)$$

Fig. 10. Stator flux estimator using APF based on BPF

The phase θ_ψ in the α -axis, its angular speed ω_ψ , and the absolute value $|\psi|$ of the stator interlinkage magnetic flux vector are calculated using (31)–(33), respectively. The angular frequency of the stator flux is calculated by the quasi-differentiate of the phase of the stator flux. This paper uses ω_{LPF} in Fig. 10 because the quasi-differentiate is needed for the differential operation and LPF. The motor torque current reference i^{ref} is defined in (34). θ_i is the motor current phase of α -axis, which is given by (35). Using (34) and (35), this paper proposes that the $\alpha\beta$ -axis current references are calculated by using (36) and (37). As the results, this paper proposes a new stationary reference frame position sensorless control based on sinusoidal tracking current controller for IPMSM.

$$\hat{\theta}_\psi = \tan^{-1} \frac{\hat{\psi}_\beta}{\hat{\psi}_\alpha} \quad (31)$$

$$\hat{\omega}_\psi = \frac{s\omega_{LPF}}{s + \omega_{LPF}} \hat{\theta}_\psi \quad (32)$$

$$|\hat{\psi}| = \sqrt{\hat{\psi}_\alpha^2 + \hat{\psi}_\beta^2} \quad (33)$$

$$i^{ref} = \frac{T^{ref}}{P|\hat{\psi}|} \quad (34)$$

$$\hat{\theta}_i = \hat{\theta}_\psi + \frac{\pi}{2} \quad (35)$$

$$i_\alpha^{ref} = i^{ref} \cos \hat{\theta}_i \quad (36)$$

$$i_\beta^{ref} = i^{ref} \sin \hat{\theta}_i \quad (37)$$

3.2 Stator Flux Estimation of Using APF based on BPF (29) and (30) can not be calculated in the actual system because (29) and (30) have a complete integration calculation process. In order to overcome this problem, this paper uses the stator flux estimation of using the all pass filter (APF) based on band pass filter (BPF) ⁽¹³⁾⁽¹⁴⁾. Stator flux estimation using APF based on BPF is shown in Fig. 9. The coordinate of stator flux estimator with APF based on BPF is shown in Fig. 10.

(I)

$$u_\alpha = v_\alpha - R_a i_\alpha \quad (38)$$

$$u_\beta = v_\beta - R_a i_\beta \quad (39)$$

(II)

This paper uses the APF whose transfer function of is shown in (40).

$$G_{APF} = \frac{s - k_B}{s + k_B} \quad (40)$$

$$v_\alpha^{apf} = G_{APF} u_\alpha \quad (41)$$

$$v_\beta^{apf} = G_{APF} u_\beta \quad (42)$$

where v_α^{apf} and v_β^{apf} are the value after the APF. Here, k_B is

determined as shown in (43).

$$k_B = \omega_B \dots\dots\dots (43)$$

In this paper, using (43), the phase characteristics is shifted to $-\frac{\pi}{2}$ degrees.

(III)

The drift occurs when the stator flux is estimated by (29) and (30). BPF is used to avoid the drift. However, the phase characteristic of the estimated value of the stator flux is changed by the BPF. Therefore, it is necessary to use (44) in order to obtain the phase characteristic that is equal to an ideal integrator.

$$G_{APF_{ideal}} = \frac{s + k_B}{s - k_B} \dots \dots \dots (44)$$

However, (44) has an unstable pole. In order to realize (44), this paper used a vector rotator. The description of a vector rotator is shown below.

$$\theta_v = \tan^{-1} \frac{u_\beta}{u_\alpha} \dots\dots\dots (45)$$

The output vector phase of APF is shown in (46).

$$\theta_{apf_B} = \tan^{-1} \frac{v_{\beta}^{apf}}{v_{\alpha}^{apf}} \dots\dots\dots (46)$$

The vector rotator is shifted to $\varphi = \theta_v - \theta_{apf_B}$. The rotator is described as shown in (47).

$$\begin{bmatrix} v_{\alpha}^{apf'} \\ v_{\beta}^{apf'} \end{bmatrix} = \begin{bmatrix} \cos 2\varphi & -\sin 2\varphi \\ \sin 2\varphi & \cos 2\varphi \end{bmatrix} \begin{bmatrix} v_{\alpha}^{apf} \\ v_{\beta}^{apf} \end{bmatrix} \dots\dots\dots (47)$$

where $v_{\alpha}^{apf'}$ and $v_{\beta}^{apf'}$ are the value after a vector rotator.

(IV)

As the result, using APF based on BPF, the stator flux linkage is estimated without phase error.

$$\hat{\psi}_\alpha = \frac{\omega_B s}{s^2 + 2\omega_B s + \omega_R^2} v_\alpha^{apf'} \dots\dots\dots (48)$$

$$\hat{\psi}_\beta = \frac{\omega_B s}{s^2 + 2\omega_B s + \omega_B^2} v_\beta^{apf'} \dots\dots\dots (49)$$

$$B = \frac{\sqrt{(\omega_{re}^2 - \omega_B^2)^2 + 4\omega_{re}^2\omega_B^2}}{\omega_B\omega_{re}^2} \dots\dots\dots (50)$$

$$|\hat{\psi}| = B \sqrt{\hat{\psi}_\alpha^2 + \hat{\psi}_\beta^2} \dots\dots\dots (51)$$

Fig. 11. Coordinate of stator flux estimator with APF based on BPF in stationary reference frame

Fig. 12. Bode diagram of the stator flux estimator of using APF based on BPF

Fig. 13. Proposed position sensorless speed control system using sinusoidal current tracking controller without flux control loop (Method 1)

$$\frac{\omega_{rm}^{res}(s)}{\omega_{rm}^{ref}(s)} = \frac{\omega_s^2}{s^2 + 2\omega_s s + \omega_s^2} \dots\dots\dots (52)$$

where B is a correction equation of the gain characteristic. As the DC gain of BPF becomes $-\infty$ as shown in Fig. 12, the offset of the current sensor is completely attenuated. In addition, the gain characteristic in Fig. 12 is affected by the angular frequency of the BPF. The phase characteristic in Fig. 12 is not affected by the angular frequency of the BPF. In addition, the phase characteristic is equal to an ideal integrator.

Therefore, this paper realizes the proposed position sensorless speed control system using the new sinusoidal current tracking controller as shown in Fig. 12, which is Method 1 of this paper. The control system of Method 1 is composed of only a stationary reference frame. Therefore, this control system does not require a coordinate transformation into a rotating coordinate. This is because it converts the output of the speed controller to the current command on the stationary reference frame by (34)–(36). The Method 1 control system does not require coordinate transformation to a rotating coordinate because it does not depend on the phase estimation accuracy.

(51) is obtained by using the angular frequency of the speed control system from (27). (51) is used for the internal model of the sinusoidal tracking current controller.

3.3 Sensorless Control Method of Torque and Flux (Method 2) In high-speed region of operation, it is necessary to carry out a flux-weakening control. To control the stator flux, this paper perform the coordinate conversion from α - β axis to M-T axis⁽¹⁵⁾⁽¹⁶⁾. The proposed magnetic flux controller is shown in Fig. 14. The current references i_M^{ref} and i_T^{ref} in Method 2 is changed into i_α^{ref} and i_β^{ref} as shown in (53). As the result, Fig. 15 shows the proposed position sensorless speed and flux control system using new sinusoidal current tracking controller, which is Method 2 of this paper. On condition of $i_M^{ref} = 0$ and $i_T^{ref} = i^{ref}$, (53) becomes (54). Therefore, on condition of on $i_M^{ref} = 0$ in Method 2, Method 2 becomes equivalent equal to Method 1.

$$\begin{bmatrix} i_\alpha^{ref} \\ i_\beta^{ref} \end{bmatrix} = \begin{bmatrix} \cos \hat{\theta}_\psi & -\sin \hat{\theta}_\psi \\ \sin \hat{\theta}_\psi & \cos \hat{\theta}_\psi \end{bmatrix} \begin{bmatrix} i_M^{ref} \\ i_T^{ref} \end{bmatrix} \\ = \begin{bmatrix} i_M^{ref} \cos \hat{\theta}_\psi - i_T^{ref} \sin \hat{\theta}_\psi \\ i_M^{ref} \sin \hat{\theta}_\psi + i_T^{ref} \cos \hat{\theta}_\psi \end{bmatrix} \\ = \begin{bmatrix} i_M^{ref} \sin \hat{\theta}_i + i_T^{ref} \cos \hat{\theta}_i \\ -i_M^{ref} \cos \hat{\theta}_i + i_T^{ref} \sin \hat{\theta}_i \end{bmatrix} \dots\dots\dots (53)$$

$$\begin{bmatrix} i_\alpha^{ref} \\ i_\beta^{ref} \end{bmatrix} = \begin{bmatrix} i^{ref} \cos \hat{\theta}_i \\ i^{ref} \sin \hat{\theta}_i \end{bmatrix} \dots\dots\dots (54)$$

4. Experimental Results

This paper confirms the validity of the proposed methods by the experimental results. Table 3 shows the angular frequency of the controller and the carrier frequency used in the experiment.

4.1 Experimental Results of Speed Control Figures 16 and 17 show the experimental results of sensorless speed control IPMSM, whose flux current condition is zero. Figure 16 shows the experimental results of Method 1. Figure 17 shows the experimental results of Method 2 on condition that $i_M^{ref} = 0$. The poles of tested current control system of Method 1 and Method 2 are designed to 2,000 rad/s⁽¹³⁾⁽¹⁴⁾. Figures 16 and 17 confirm that Method 1 and Method 2 of the proposed sensorless speed control systems have stable speed and torque responses. Moreover, both speed responses of Method 1 and Method 2 are almost the same as those of Fig. 8, which are the speed responses of a speed control system with a position sensor. However, the ripple of i^{ref} during

Fig. 14. Flux control loop of Method 2

Fig. 15. Proposed position sensorless speed and flux control system using sinusoidal current tracking controller (Method 2)

Table 3. Motor parameters and simulation conditions

Poles of the speed control system	ω_s	80 [rad/s]
Poles of the flux control system	ω_f	80 [rad/s]
Poles of the current control system	ω_c	2000 [rad/s]
Cut-off frequency of low-pass filter	ω_{LPF}	640 [rad/s]
Pole of band-pass filter	ω_B	10 [rad/s]
Carrier frequency	f_s	10 [kHz]

Fig. 16. Experimental results of speed step response in Method 1

the 1.2–1.4 s interval is larger than at any other time. The reason for this is that the phase estimation value of the stator flux linkage by the load torque becomes oscillatory.

Figure 18 shows that the experimental results of step speed response of Method 2 from –700 rpm to 700 rpm. Figure 18 confirms that the proposed sensorless speed control system well regulates the motor speed throughout zero speed area.

4.2 Experimental Results of Flux-weakening Control

When the voltage vector becomes larger than V_{max} , the stator flux command⁽⁴⁾ is determined by (55). V_{max} is the maximum voltage of inverter.

$$\psi^{ref} = \frac{-R_a i_T + \sqrt{V_{max}^2 - (R_a i_M)^2}}{\hat{\omega}_\psi} \quad (55)$$

In this paper, the tested PWM inverter has the triangular wave comparison modulation method. Therefore, the maximum voltage of tested the inverter is given by (56).

Fig. 17. Experimental results of speed step response on condition of $i_M^{ref} = 0$ in Method 2

$$V_{max} = \frac{1}{2} \sqrt{3} V_{dc} \quad (56)$$

To confirm the effectiveness of the flux-weakening control, Fig. 19 shows the experimental results of acceleration experiment for the voltage saturation region. V_{dc} is set to 200 V. The experimental results confirm that the proposed flux weakening control has the fine and stable flux control response as shown in Fig. 19. The proposed sensorless speed control system well regulates the motor speed in the region of voltage saturation. Therefore, the proposed flux-weakening control has the stable speed response in the region of voltage saturation.

5. Conclusion

This paper proposes a new stationary reference frame position sensorless control based on stator flux linkage and new sinusoidal current tracking controller for IPMSM.

At first, this paper proposes a new sinusoidal current tracking controller and carries out its performance analysis based

Fig. 18. Experimental results of speed step response from -700rpm to 700rpm on condition of $i_M^{\text{ref}} = 0$ in Method 2

on the extended transformation. This analysis points out the harmonic disturbance suppression performance of α - β -sinusoidal current controller. The numerical simulation results and experimental results confirm that the disturbance suppression characteristics of new α - β -sinusoidal current tracking controller is higher than that of d-q PI current controller.

Next, this paper newly proposes the two types of position sensorless control systems based on the proposed sinusoidal current tracking controller. Method 1 is the position sensorless control method. Method 2 is the position sensorless speed control of both stator flux and torque. The experimental results confirm that Method 1 and Method 2 have stable speed and torque responses. Moreover, the flux-weakening control of Method 2 is well achieved. Since the proposed flux-weakening control in Method 2 has no information of inductance parameters, it has a fine stator flux control in the region of voltage saturation.

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Fig. 19. Experimental results of acceleration control with flux weakening control in Method 2

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Appendix

1. Extended Transfer Function

In order to confirm the effectiveness of the α - β sinusoidal current control, it is compared with the current control system on a d - q synchronous frame, whose name is d - q PI current control. The input and output frames of d - q PI current control are different from them of α - β sinusoidal current control. In order to analyze the performance of d - q PI current control, this paper uses the rotating coordinate transformation $e^{j\omega_{re}t}$ and $e^{-j\omega_{re}t}$ (9)(10). The d - q PI current control is equivalently converted on α - β stationary frame.

app. Fig. 1 shows the extended transformation of the equivalent d - q PI current controller on α - β stationary frame. In app. Fig. 1, the voltage reference on α - β stationary frame $v_{\alpha\beta}^{ref}$ is defined in (A5). Furthermore, (A5) is rewritten as (A6).

$$v_{\alpha\beta}^{ref} = v_{\alpha}^{ref} + jv_{\beta}^{ref} \quad (A1)$$

$$i_{\alpha\beta}^{err} = i_{\alpha}^{ref} - i_{\alpha}^{res} = i_{\alpha}^{err} + ji_{\beta}^{err} \quad (A2)$$

$$T^{-1} = \begin{bmatrix} \cos \omega_{re}t & j \sin \omega_{re}t \\ j \sin \omega_{re}t & \cos \omega_{re}t \end{bmatrix} = e^{j\omega_{re}t} \quad (A3)$$

$$T = \begin{bmatrix} \cos \omega_{re}t & -j \sin \omega_{re}t \\ -j \sin \omega_{re}t & \cos \omega_{re}t \end{bmatrix} = e^{-j\omega_{re}t} \quad (A4)$$

$$\begin{bmatrix} v_{\alpha}^{ref} \\ jv_{\beta}^{ref} \end{bmatrix} = T^{-1} \begin{bmatrix} \frac{K_p s + K_i}{s} & 0 \\ 0 & \frac{K_p s + K_i}{s} \end{bmatrix} T \begin{bmatrix} i_{\alpha}^{err} \\ ji_{\beta}^{err} \end{bmatrix} \quad (A5)$$

$$\frac{v_{\alpha\beta}^{ref}}{i_{\alpha\beta}^{err}} = \frac{K_p(s - j\omega_{re}) + K_i}{s - j\omega_{re}} \quad (A6)$$

Where, K_p and K_i are the d - q PI current controller gains. app. Fig. 2 shows extended transformation of the equivalent

app. Fig. 1. Equivalent d - q PI current controller on α - β stationary frame

app. Fig. 2. Extended transformation of equivalent d - q PI current control system on α - β stationary frame

app. Table 1. Different IPMSM parameters

R_a	0.78 [Ω]
L_d	6.43 [mH]
L_q	10.37 [mH]
P	5
J_M	3.3×10^{-4} [kgm ²]

d - q PI current control system on the α - β stationary frame. The transfer function to the current response $i_{\alpha\beta}^{res}$ from the disturbance $i_{\alpha\beta}^{dis}$ is defined as (A7).

$$G_{dq-PI}^{dis} = \frac{i_{\alpha\beta}^{res}}{i_{\alpha\beta}^{dis}} = \frac{-s^2 + n_1 s + n_0}{s^2 + m_1 s + m_0} \quad (A7)$$

$$m_0 = \frac{K_i - j(R_a + K_p)\omega_{re}}{L_d} \quad (A8)$$

$$m_1 = \frac{(R_a + K_p - jL_d\omega_{re})}{L_d} \quad (A9)$$

$$n_0 = \frac{jR_a\omega_{re}}{L_d} \quad (A10)$$

$$n_1 = \frac{-R_a + jL_d\omega_{re}}{L_d} \quad (A11)$$

2. Evaluation based on Extended Transformation of the Disturbance Suppression Characteristics

The current control system using the sinusoidal tracking controller in the modified IPMSM obtained similar disturbance suppression characteristics as shown in Fig. 4. The current control system, in the case of the modified IPMSM shown in app. Table 1, obtained the disturbance suppression characteristics shown in app. Figs. 3 and 4.

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