Abstract—This paper examines the control performance of a permanent magnet synchronous motor drive system based on direct torque control (DTC) using a SiC-MOSFET inverter (SiC inverter). The dead-time of the inverter affects the performance of the motor control. In high-speed drives, the dead-time effect increases because the control period becomes shorter. By using a SiC inverter for which the switching speed is faster than that of the Si-IGBT inverter, the dead-time can be shortened, and performance improvement of the control is expected. This paper also proposes a sensorless starting method using DTC. It is difficult to start the motor in a drive system without a position sensor. In the conventional ultra-high-speed drive using DTC, the motor is started by open-loop control. Experiments reveal the effectiveness of using a SiC inverter and shortening the dead-time. In addition, the experimental results reveal the applicability of the proposed starting method.

Keywords—Direct torque control (DTC), permanent-magnet synchronous motor (PMSM), ultra-high-speed drive, SiC inverter

I. INTRODUCTION

Due to advances in power electronics and high-performance microprocessors and the development of low-iron-loss magnetic materials, permanent magnet synchronous motors (PMSMs) have recently been driven at ultra-high speeds. Ultra-high-speed PMSMs are used in a number of applications, including compressors, superchargers, and air blowers [1]-[3].

Current control or direct torque control (DTC) is used for high-efficiency driving with wide ranges and variable speeds. Current control often works on the d-q frame based on the rotor magnet flux, whereas DTC is operated in a stationary reference frame based on the stator winding. DTC was first proposed for induction machine drives but can also be applied to PMSM drives [4]. In high-speed drives, since the fundamental frequency becomes high, a short sampling period is required for stable control, and so the speed sensor becomes unreliable. DTC is therefore suitable for ultra-high-speed PMSM drives because of its simple structure and sensorless feature.

The dead-time is required in order to avoid cross-conduction current of switching devices of the inverter used for the motor drive, but the dead-time causes problems such as output current distortion and voltage error [5]. In particular, in high-speed drives, the voltage error, which can be expressed in the form of the dead-time divided by the control period, increases, thus degrading the control performance.

The sensorless drive system for an ultra-high-speed PMSM based on DTC has been proposed in [6]. This paper examines the control performance of the ultra-high-speed PMSM drive system based on DTC by shortening the dead-time using the SiC inverter. A performance improvement is expected by shortening the dead-time, because the voltage error is proportional to the dead-time.

This paper also proposes a sensorless starting method using DTC. Start the PMSM in a position sensorless drive system is difficult. In the conventional ultra-high-speed drive based on DTC, the motor is started by open-loop control [6]. The proposed method achieves sensorless starting by changing the reference flux depending on the rotor speed.

In the proposed system, the reference flux vector calculator (RFVC) DTC [7] is applied for ultra-high speeds (rated speed: 42,000 min⁻¹). The experimental results demonstrate the effectiveness of the SiC inverter. In addition, the proposed sensorless starting method is experimentally demonstrated to be valid.

II. DIRECT-TORQUE-CONTROL-BASED MOTOR DRIVE SYSTEM

A. Definition of Coordinate Axes

DTC operates in the α-β frame, which is a stationary reference frame. Fig. 1 shows the relationship between the d-q and α-β frames. The stator flux linkage position $\psi_s$ is defined as the angle between the stator flux linkage vector $\psi_s$ and the α-axis, and $\delta$ is the torque angle.
B. Reference Flux Vector Calculator DTC

Fig. 2 shows a block diagram of the PMSM drive system using RFVC DTC. The stator flux linkage is estimated in the \( \alpha-\beta \) frame. Therefore, the DTC system is suitable for drives without position sensors. The estimated stator flux linkage (\( \hat{\psi}_a, \hat{\psi}_b \)), the estimated torque (\( \hat{T}_e \)), the estimated position of the stator flux linkage (\( \hat{\theta}_s \)), and the reference voltages (\( v^*_a \), \( v^*_b \)) are calculated as follows:

\[
\hat{\theta}_s = \tan^{-1} \frac{\hat{\psi}_a}{\hat{\psi}_b} \\
\hat{T}_e = P_n (\hat{\psi}_b \dot{\theta}_s - \hat{\psi}_a \dot{\theta}_s) \\
v^*_a = \frac{\hat{\psi}_a - \dot{\hat{\psi}}_a}{t_s} + R_a i_a \\
v^*_b = \frac{\hat{\psi}_b - \dot{\hat{\psi}}_b}{t_s} + R_a i_b
\]

where \( R_a \) is the armature resistance, \( i_a \) and \( i_b \) are the \( \alpha \)- and \( \beta \)-axis components of the armature currents, \( P_n \) is the number of pole pairs, \( \dot{\hat{\theta}}_s \) is the estimated position of the stator flux linkage in the \( \alpha-\beta \) frame, and \( t_s \) is the sampling period.

The dashed line shown in Fig. 2 indicates the RFVC block, which calculates the reference flux vector using a proportional and integral (PI) controller.

C. Voltage Error due to the Inverter

In this study, considering the voltage error of the inverter, voltage error compensation is applied to this system. The voltage error of the dead-time is assumed to be much larger than the voltage drop of the switching device due to the short sampling period. The relationship between the reference voltages and the inverter output voltages is expressed as follows [8]:

\[
\begin{bmatrix}
  v^*_a \\
  v^*_b \\
  v^*_w
\end{bmatrix} =
\begin{bmatrix}
  v_a^* \\
  v_b^* \\
  v_w^*
\end{bmatrix} - \Delta V
\begin{bmatrix}
  \text{sgn} \hat{\psi}_a \\
  \text{sgn} \hat{\psi}_b \\
  \text{sgn} \hat{\psi}_w
\end{bmatrix}
\]

\( \Delta V = \frac{I_d}{I_s} V_{DC} \)

where \( v_a^* \), \( v_b^* \), and \( v_w^* \) represent the three-phase inverter output voltages, \( v_a^{**} \), \( v_b^{**} \), and \( v_w^{**} \) represent the compensated three-phase reference voltages, \( \Delta V \) is the voltage error due to the dead time of the inverter, \( i_a \), \( i_b \), and \( i_w \) are the three-phase currents, \( \text{sgn}() \) is the sign function, \( t_d \) is the dead-time of the inverter, and \( V_{DC} \) is the DC link voltage. In this paper, \( t_s \) is equal to the inverse of the PWM carrier frequency.

D. Speed Sensorless Drive

Fig. 3 shows the composition of the speed estimator and the reference flux calculator. The estimated rotor speed (\( \hat{\omega}_m \)) is calculated by taking the difference of the estimated position of the stator flux linkage (\( \hat{\theta}_s \)) and using a low-pass filter (LPF).
because the rotor velocity corresponds to the velocity of the stator flux-linkage vector in a steady state. A PI controller is used for speed control, and the relationship between the torque and the flux is obtained using maximum torque per ampere (MTPA) control for a high-efficiency drive.

III. SENSORLESS STARTING METHOD

The induced voltage $V_a$ and the stator flux linkage $\Psi_s$ are related as follows:

$$V_a = \omega \Psi_s$$  \hspace{1cm} (7)

where $\omega$ is the electrical angular velocity.

In the low-speed region, the induced voltage becomes small. Based on the reference voltage, a sufficient voltage cannot be provided to the motor because the modulation factor of the inverter is close to zero. In addition, the voltage error in the inverter cannot be ignored. The stator flux linkage of the DTC is estimated by integrating the voltage, as shown in (1), so that the stator flux linkage estimation is not performed correctly in the low-speed region.

In this paper, the reference flux is calculated in inverse proportion to the rotor speed in order to ensure that the armature voltage is sufficiently high. In order to apply MTPA control when the rotor speed becomes sufficiently high, the reference flux is calculated as follows:

$$\Psi_s^* = \frac{K_1}{\omega_m} \Psi_s^{*-\text{MTPA}}$$  \hspace{1cm} (8)

where $K_1$ is the constant of proportionality, and $\Psi_s^{*-\text{MTPA}}$ is the reference flux given by MTPA control.

Fig. 4 shows the reference fluxes of the conventional method and the proposed method. Fig. 5 shows the composition of the reference flux calculator of the proposed method.

Substituting $\Psi_s^*$ in (8) for $\Psi_s$ in (7), the induced voltage $V_a$ is expressed as follows:

$$V_a = \omega \left( \frac{K_1}{\omega_m} + \Psi_s^{*-\text{MTPA}} \right)$$  \hspace{1cm} (9)

$$= P_n K_1 + P_e \dot{\omega}_m \Psi_s^{*-\text{MTPA}}$$

Fig. 6 shows the voltage characteristic of (9). The induced voltage becomes $P_n K_1$ when $\dot{\omega}_m = 0$ (the motor is standstill) using the proposed method.

IV. EXPERIMENTAL RESULTS

A. Experimental Setup

Table I lists the parameters of the PMSM drive system tested in this paper. Using the motor for a vacuum cleaner, the load torque is proportional to the square of the rotor speed. All of the controls are processed through a digital signal processor. The sampling period is 62.5 $\mu$s, which is equal to the inverse of the PWM carrier frequency (16 kHz). An SCH2080KE semiconductor (ROHM Corp., Ltd.) is used for the switching
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TABLE I
PARAMETERS OF PMSM DRIVE SYSTEM

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Rated speed</td>
<td>42000 min⁻¹</td>
</tr>
<tr>
<td>Rated current</td>
<td>16 A</td>
</tr>
<tr>
<td>Armature resistance Ra</td>
<td>0.05 Ω</td>
</tr>
<tr>
<td>Magnet flux Φp</td>
<td>2.7 mWb</td>
</tr>
<tr>
<td>d-axis inductance Ld</td>
<td>0.07 mH</td>
</tr>
<tr>
<td>q-axis inductance Lq</td>
<td>0.1 mH</td>
</tr>
<tr>
<td>Number of pole pairs Pa</td>
<td>1</td>
</tr>
<tr>
<td>Inertia moment Jm</td>
<td>5.0×10⁻⁶ kg·m²</td>
</tr>
<tr>
<td>Load torque Tload</td>
<td>3.7×10⁻⁸xωn² [Nm]</td>
</tr>
<tr>
<td>DC link voltage VDC</td>
<td>40 V</td>
</tr>
</tbody>
</table>

devices of the SiC-MOSFET inverter, and a PS21767 semiconductor (Mitsubishi Electric Corp.) is used as a Si-IGBT inverter.

B. Performance Comparison by Changing the Inverter and the Deadtime

Driving the ultra-high-speed PMSM at the rated speed of 42,000 min⁻¹, we compared the control performance by changing the type of switching device and the dead-time. The dead-time of the Si-IGBT inverter is selected to be 3.5 μs, considering the limitation of the MWINV-5R022 inverter unit (Myway Plus Corp.). Similarly, the dead-time for the SiC-MOSFET inverter MWINV-1044-SiC (Myway Plus Corp.) is chosen to be 0.5 μs.

Fig. 7 shows the u-phase current as measured using a current probe. The current waveform is controlled sinusoidally, and the PMSM is found to be operated at 42,000 min⁻¹ because a single period of the u-phase current is approximately 1.4 ms. The current ripples appear at the zero crossing point because voltage error compensation is applied using the sign of the current. Table II shows the results of frequency analysis of the u-phase current. The ratios shown in Table II are the harmonic components divided by the fundamental component. Comparing the Si-IGBT inverter and the SiC-MOSFET inverter when the dead-time is set to 3.5 μs, the harmonic components are not so different. However, the amplitude of the fundamental component of the SiC-MOSFET inverter is smaller than that of the Si-IGBT inverter. This is because MTPA control performs better as a result of the reduction of the flux estimation error in DTC due to the decrease in the voltage error in the inverter, as shown in Fig. 8. The voltage error is calculated from the deference between the reference armature voltage and the line voltage.

In the case of the SiC-MOSFET inverter, the amplitude of the fundamental component and the ratio of the harmonic component in the u-phase current are reduced by setting the dead-time to 0.5 μs, as shown in Table II. The reason for this is that the voltage error due to the dead-time is reduced and the control is better performed by setting a small dead-time.

Fig. 9 shows the torque waveform. The estimated torque is calculated by (2). When the dead-time is 0.5 μs, torque ripples are reduced and so the average of the estimated torque is reduced. The reason for this is that the current ripples are

<table>
<thead>
<tr>
<th>Switching device (Dead-time τd)</th>
<th>u-phase current</th>
<th>harmonic component</th>
</tr>
</thead>
<tbody>
<tr>
<td>(3.5 μs)</td>
<td>18.40 A</td>
<td>5th 7th 11th 13th</td>
</tr>
<tr>
<td>Si-IGBT</td>
<td>10.0 A (5.4%)</td>
<td>0.92 A (5.0%)</td>
</tr>
<tr>
<td>Si-IGBT</td>
<td>10.0 A (5.4%)</td>
<td>0.92 A (5.0%)</td>
</tr>
<tr>
<td>SiC-MOSFET (3.5 μs)</td>
<td>17.63 A</td>
<td>0.83 A (4.7%)</td>
</tr>
<tr>
<td>SiC-MOSFET (3.5 μs)</td>
<td>0.83 A (4.7%)</td>
<td>0.89 A (5.4%)</td>
</tr>
<tr>
<td>SiC-MOSFET (0.5 μs)</td>
<td>17.42 A</td>
<td>0.44 A (2.5%)</td>
</tr>
<tr>
<td>SiC-MOSFET (0.5 μs)</td>
<td>0.44 A (2.5%)</td>
<td>0.69 A (4.0%)</td>
</tr>
</tbody>
</table>

Fig. 7. Waveform of u-phase current.

TABLE II
FREQUENCY ANALYSIS OF THE U-PHASE CURRENT

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reduced, as shown in Table II.

Fig. 10 shows the flux waveform. There is an offset between the estimated flux and the reference flux, which originates from the effect of using an LPF for the voltage integration in the flux estimator and the discretization error.

The flux ripples are reduced due to torque ripple reduction when the dead-time is 0.5 μs. In this system, the estimated rotor speed $\hat{\omega}_m$ is calculated based on the rotor position $\hat{\theta}_r$, which is calculated based on the estimated stator flux $\hat{\psi}_s$. Therefore, the ripples of the estimated speed are also reduced, as shown in Fig. 11.

![Flux waveform](image1)

![Flux waveform](image2)

![Flux waveform](image3)
C. Performance of the Sensorless Starting Method

In this paper, \( K_1 \) in (8) is set to 3.5, and the upper limit of \( \psi_s^* \) is set to 18 mWb in order to prevent the reference flux from diverging.

Fig. 12 shows the starting characteristics. In this experiment, the acceleration rate is limited in order to avoid step out. The motor is started at 0.5 s. The PMSM can be started up from standstill because the amplitude of the \( u \)-phase voltage is maintained approximately constant, as shown in Fig. 12(d), and thus a sufficient voltage is provided to the motor in the low-speed region. In addition, the reference flux is reduced according to the rotor speed and approaches the reference flux given by MTPA control.

Fig. 13 shows the acceleration characteristics of the proposed method when the rotor speed is increased from standstill to 42,000 min\(^{-1}\). The acceleration rate is limited in the low-speed region in order to prevent the rotor from step out. The proposed method allows smooth acceleration without switching the control method.

V. CONCLUSIONS

This paper examined the control performance in the case of changing the dead-time for an ultra-high-speed PMSM drive system using a SiC inverter. The experiment revealed the effectiveness of using the SiC inverter. The control performance of the motor drive system is improved by using the SiC inverter because the voltage error is reduced due to the small voltage drop of the switching devices of the SiC inverter. In addition, the control performance is improved by setting a small dead-time because the voltage error due to the dead-time is reduced. A sensorless starting method using DTC was also proposed, in which the reference flux changes according to the rotor speed in order to keep the voltage approximately constant in a low-speed region. Using the proposed method, providing the reference flux to keep the armature voltage constant, sensorless starting is demonstrated experimentally to be possible, and the rotor speed can be increased smoothly without switching the control method.

Fig. 12. Characteristics of startup.
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REFERENCES


Fig. 13. Acceleration characteristics for the proposed method.