Control Method of Electrolytic Capacitor-less Dual Inverter Fed IPMSM for Reducing Torque Ripple under Grid Disturbance

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This paper proposes a control method for torque ripple using an open-end winding IPMSM driven by an electrolytic capacitor-less dual inverter under grid disturbances. Under grid disturbances, a conventional motor speed range extension method causes torque ripple due to voltage saturation of the main inverter. The proposed control method achieves both an extension of the motor speed range and a reduction of torque ripple under grid disturbances. The proposed method cancels the harmonic voltage generated due to grid disturbances by the output voltage of the compensating inverter. The effectiveness of the proposed method is verified by experimental results using an electrolytic capacitor-less dual inverter and an open-end winding IPMSM under the input voltage distorted by the seventh harmonic voltage. Compared with the conventional method, the proposed method reduces the harmonic currents and the torque ripple by up to 85.6 and 91.0%, respectively, under the distorted input voltage (seventh harmonic, THD 25%). Therefore, the motor drive system using the proposed method can effectively extend the motor speed range, and reducing the torque ripple under grid disturbances.

Keywords: dual inverter, electrolytic capacitor-less inverter, grid disturbance, open-end winding IPMSM, torque ripple

1. Introduction

In recent years, industrial applications cannot ignore grid disturbances. Large-scale distributed generations (DGs) such as photovoltaic generation and wind power generation, have been implemented in the power grid to conserve energy resources(1)(2). However, DGs cannot provide stable power supplies because of varying weather conditions, thereby resulting in poor power quality of the power grid. In addition, the typical nonlinear loads produce large harmonic currents and cause grid voltage distortions(3). Therefore, the need for countermeasures under grid disturbances is increasing in industrial applications.

In general, grid-connected motor drive systems driven by the dual inverter include a three-phase diode rectifier with a simple circuit configuration, and a large-capacity electrolytic capacitor for energy buffer. However, electrolytic capacitors reduce the reliability and the maintainability of the system as they are the most fragile components in power converters(4)(5). The electrolytic capacitor-less dual inverter using film capacitors improves the reliability and the maintainability because the failure rate of film capacitors is lower than that of electrolytic capacitors(6)(7). The electrolytic capacitor-less dual inverter eliminates the need for the large-capacity energy buffer in the power converter, consequently, problems that occur in the power grid directly affect the motor drive system. As a grid voltage disturbance, there are grid voltage distortions, unbalanced grid voltages, phase jumps, and frequency fluctuations. For example, the average dc-link voltage decreases and the dc-link voltage ripple increases under a distorted grid voltage. In the load condition with high modulation index, the controller output is saturated due to voltage saturation in Inv. 1 under grid disturbances. Consequently, harmonic voltages are superimposed on the output voltage of the main inverter. Thus, torque ripple is generated due to the harmonic currents. Furthermore, the peak current increased by the harmonic current may exceed the rated current and the demagnetization resistance, which reduces the reliability of the system. The authors propose control methods for the dual inverter to reduce the harmonic voltages(8)(9). The improved motor winding voltage using the dual inverter achieves higher motor efficiency, reduced torque ripple and lower weight of the system. However, these control methods cannot extend the motor speed range.

This paper proposes a control method for an open-end winding IPMSM driven by an electrolytic capacitor-less dual inverter with a floating capacitor to improve reliability and reduce torque ripple under grid disturbances. The proposed control method is based on a novel anti-voltage saturation treatment. The proposed method reduces the harmonic voltage superimposed on the motor winding voltage under a distorted grid voltage while extending the motor speed range, and the capacitors design method of the system using the proposed control is also discussed. The effectiveness is demonstrated experimentally by reducing the harmonics superimposed on the motor current and torque under grid disturbances.

2. Analysis of Conventional Control Method

2.1 Electrolytic Capacitor-less Dual Inverter

Fig. 1 shows a motor drive system of an open-end winding IPMSM driven by an electrolytic capacitor-less dual inverter...
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**Fig. 1.** Electrolytic capacitor-less dual inverter consisting of a three-phase diode rectifier, two small film capacitors ($C_{dc1}$ and $C_{dc2}$), two inverters (Inv.1, Inv.2), and an open-end winding IPMSM. The Inv. 2 cannot supply the constant active power because the power supply of Inv. 2 is the compensating capacitor. Thus, the motor speed range of an open-end winding IPMSM driven by a dual inverter is widest when Inv. 1 supplies only active power and Inv. 2 supplies only reactive power. The Inv.1 operates as the main inverter that supplies the active power required to drive the open-end winding IPMSM. The Inv.2 operates as the compensating inverter that supplies the reactive power required to drive the open-end winding IPMSM. The motor speed range is extended by increasing the active power output by the Inv.1 because the Inv.1 does not supply the reactive power required to drive the IPMSM. References (8) and (9) report that dual inverter drives extend the motor speed range by at least 10% compared to single inverter drives. The two film capacitors are connected to the dc side of each inverter. By controlling the charging and discharging of the compensating capacitor $C_{dc2}$, the Inv. 2 supplies the reactive power. Therefore, the dual inverter has the advantages of reducing the volume and weight of the motor drive system because it extends the motor speed range without adding a boost reactor.

**2.2 Issues Caused by Distorted Grid Voltages** In this study, grid disturbances are focused on distorted grid voltages. The distorted grid voltage input to the diode rectifier significantly affects the dc-link voltage. Generally, the harmonic components superimposed on the distorted grid voltage are dominated by the fifth and seventh harmonics. Fig. 2 shows the simulation results of the dc-link voltage under the distorted grid voltages. Fig. 2(a) shows the dc-link voltage waveform under a normal grid voltage ($200\,\text{Vrms}, \, 50\,\text{Hz}$). Fig. 2(b) shows the dc-link voltage waveform under the grid voltage ($200\,\text{Vrms}, \, 50\,\text{Hz}$ and $250\,\text{Hz}$) with superimposed fifth harmonic voltage (THD, total harmonic distortion = 25%). Fig. 2(c) shows the dc-link voltage waveform under the grid voltage ($200\,\text{Vrms}, \, 50\,\text{Hz}$ and $350\,\text{Hz}$) with superimposed seventh harmonic voltage (THD = 25%). Fig. 2(d) shows the dc-link voltage waveform under the grid voltage ($200\,\text{Vrms}, \, 50\,\text{Hz}$ and $250\,\text{Hz}, \, 350\,\text{Hz}$) with superimposed fifth and seventh harmonic voltage (THD = 25%, Fifth harmonic voltage and the seventh harmonic voltage have the same voltage amplitude.). The dc-link voltage under a normal grid voltage causes the voltage ripple at 300 Hz, which is six times the frequency of the grid frequency (50 Hz) because a large-capacity electrolytic capacitor is not employed in the dc-link. Fig. 2 shows that the dc-link voltage ripple is larger and the minimum dc-link voltage is smaller under the condition of the seventh harmonic are superimposed on the grid voltage compared with the other conditions. Consequently, the countermeasure is especially necessary for the grid voltage dominated by the seventh harmonic. The dc-link voltage ripple under the grid voltage superimposed with seventh harmonic at 25% increases by 335% compared with the dc-link voltage ripple under the normal grid voltage. The harmonic voltage superimposed on the output voltage of Inv.1 under grid disturbances is analyzed in Ref. (7). The frequencies of the harmonic voltages superimposed on the output voltage of Inv.1 are frequencies of the sum and difference of the dc-link ripple frequency and the electric angular frequency. The larger the dc-link voltage ripple and the higher the modulation index, the larger is the harmonic voltage contained in the output voltage of Inv.1. The Inv.1 causes saturation of the controller output due to voltage saturation during the drive with high modulation index under the distorted grid voltage. The harmonic components superimposed on the output voltage of the Inv.1 are input to the motor winding voltage, and the torque ripple is caused by the harmonic current superimposed on the motor current. The harmonic voltage increases at high motor speed and high load torque, which are the advantages of the dual inverter because this load condition requires the high modulation index. Therefore, the motor drive system driven by the electrolytic capacitor-less dual inverter require countermeasures under grid disturbances.

**2.3 Conventional Control Method**

- The motor speed range is extended by increasing the voltage input to the main inverter. References (8) and (9) report that dual inverter drives extend the motor speed range by at least 10% compared to single inverter drives.
the motor winding voltage. The $d$- and $q$-axis motor winding voltage references are expressed as follows:

$$v_d^{ref} = v_{d1}^{ref} - v_{d2}^{ref}, \quad v_q^{ref} = v_{q1}^{ref} - v_{q2}^{ref} \quad (1)$$

Here, $v_{d1}^{ref}$ and $v_{q1}^{ref}$ are the $d$- and $q$-axis voltage references of Inv.1, and $v_{d2}^{ref}$ and $v_{q2}^{ref}$ are the $d$- and $q$-axis voltage references of Inv.2, respectively. The $d$- and $q$-axis motor winding voltages are the voltages of the difference between the $d$- and $q$-axis voltages of each inverter, respectively.

The motor speed range is extended by increasing the voltage of the electric angular basic frequency component included in the motor winding phase voltage. The $v_d^{ref}$ and $v_q^{ref}$ are determined by the automatic current regulator (ACR) output. The $v_{d1}^{ref}$ and $v_{q1}^{ref}$ must be generated considering the power constraints of each inverter to extend the motor speed range. The motor speed range of an open-end winding IPMSM driven by a dual inverter is widest when Inv. 1 supplies only active power and Inv. 2 supplies only reactive power. Fig. 3 shows the power flow of the motor drive system driven by the dual inverter. The Inv. 1 supplies only the active power required to drive the IPMSM, and the reactive power is 0. The Inv. 2 supplies only the reactive power required to drive the IPMSM, and the active power is 0. The instantaneous active power reference of the IPMSM is expressed as follows:

$$p_m^{ref} = p_1^{ref} - p_2^{ref} \quad (3)$$

Here, $p_1^{ref}$ is the instantaneous active power reference of Inv.1, and $p_2^{ref}$ is the instantaneous active power reference of Inv.2. The $p_1^{ref}$ is expressed as follows:

$$p_1^{ref} = \frac{3}{2} \left( v_{d1}^{ref} i_d^{ref} + v_{q1}^{ref} i_q^{ref} \right) \quad (4)$$

Here, $i_d^{ref}$ and $i_q^{ref}$ are the $d$- and $q$-axis motor current references, respectively. The $p_m^{ref}$ is expressed as follows using the motor winding voltages and the motor currents:

$$p_m^{ref} = \frac{3}{2} \left( v_d^{ref} i_d^{ref} + v_q^{ref} i_q^{ref} \right) \quad (5)$$

The Inv.1 does not supply the reactive power required to drive the IPMSM. The instantaneous reactive power of Inv.1 is expressed as follows:

$$q_1^{ref} = \frac{3}{2} \left( v_{q1}^{ref} i_d^{ref} - v_{d1}^{ref} i_q^{ref} \right) = 0 \quad (6)$$

From equations (3), (4) and (6), the $v_{d1}^{ref}$ and $v_{q1}^{ref}$ are expressed as follows:

$$v_{d1}^{ref} = \frac{2}{3} p_1^{ref} i_d^{ref} \quad (7)$$

$$v_{q1}^{ref} = \frac{2}{3} p_1^{ref} i_q^{ref} \quad (8)$$

Here, $I_a$ is the amplitude of the motor phase current. From equations (1) and (2), the $d$- and $q$-axis voltage references of Inv.2 are determined by the difference between the $v_{d1}^{ref}$ and $v_{d2}^{ref}$, and the $v_{q1}^{ref}$ and $v_{q2}^{ref}$, respectively.

For the compensating capacitor connected to the input of Inv.2, the compensating capacitor voltage control is essential from the perspective of the rated voltage of Inv.2. The relationship between the compensating capacitor capacitance $C_{dc2}$ and the compensating capacitor voltage $v_{dc2}$ is expressed as follows:

$$w_{c2} = \int p_2 dt = \frac{1}{2} C_{dc2} v_{dc2}^2 \quad (9)$$

Here, $w_{c2}$ is the compensating energy, $p_2$ is the instantaneous active power of Inv.2. The $v_{dc2}$ depends on the instantaneous active power of Inv.2. Consequently, the control of the compensating capacitor voltage uses the instantaneous active power of Inv.2 as the output of the voltage controller. The average active power of the Inv.2 must be set to 0 because the power supply of Inv. 2 is the compensating capacitor. Thus, the compensating capacitor voltage controls only the average voltage of $v_{dc2}$. The average compensating capacitor voltage is controlled by a proportional-integral (PI) controller to track the capacitor voltage reference. From the above, the dual inverter extends the motor speed range. However, the conventional method considers only the voltage of the electric angular basic frequency component included in the motor winding phase voltage. Consequently, the Inv.1 causes saturation of the controller output due to voltage saturation under grid disturbances. An anti-voltage saturation treatment for the electrolytic capacitor-less dual inverter with motor speed range extension method has not been studied. Consequently, the harmonic voltages superimposed on the output voltage of Inv.1 are directly input to the motor winding voltage under grid disturbances. Therefore, the harmonic current generated by grid disturbances causes torque ripple.

### 3. Proposed Control Method

#### 3.1 Harmonic Voltage Reduction Method

The proposed control method is based on the conventional method of Ref. (8). Fig. 4 shows the relationship between the output phase voltages of the dual inverter and the motor winding phase voltage in each control method. From equations (1) and (2), the harmonic voltages superimposed on the output voltage of Inv.1 are cancelled by outputting the same harmonic voltages in Inv.2, and the torque ripple is reduced by inputting the ideal motor winding voltage. Thus, Inv.2 outputs the current controller output saturation and the voltage to provide reactive power required to drive the IPMSM. The $p_2^{ref}$ is expressed as follows:

$$p_2^{ref} = \frac{3}{2} \left( v_{dc2}^{ref} i_d^{ref} + v_{q2}^{ref} i_q^{ref} \right) \quad (10)$$

The $i_d^{ref}$ and $i_q^{ref}$ are constant in order to reduce torque ripple. The $v_{dc2}^{ref}$ and $v_{q2}^{ref}$ are pulsating because the Inv.2...
outputs the harmonic voltages. Consequently, the instantaneous active power of Inv.2 pulsates significantly. The instantaneous active power of Inv.2 must satisfy the three requirements. A) The average instantaneous active power of Inv.2 is 0 because the power supply of Inv.2 is the compensating capacitor. B) The average compensating capacitor voltage is controlled to be constant. C) The Inv.2 outputs the harmonic voltages. The requirements of A) and B) are satisfied by the compensating capacitor voltage PI control and equations (7) and (8). The proposed controller must know the harmonic voltage to satisfy the requirement of C). The \(d\)- and \(q\)-axis motor winding harmonic voltages are expressed as follows:

\[
\Delta v_d = v_d - v_d^{ref} \quad \cdots \cdots \cdots (12) \\
\Delta v_q = v_q - v_q^{ref} \quad \cdots \cdots (13)
\]

Here, \(v_d\) and \(v_q\) are the \(d\)- and \(q\)-axis motor winding voltages, respectively. The \(d\)- and \(q\)-axis motor winding harmonic voltages come from the current controller output saturation. They are determined by the difference between the \(v_d^{ref}\) and \(v_d\), and the \(v_q^{ref}\) and \(v_q\), respectively. The \(v_d^{ref}\) and \(v_q^{ref}\) are determined by the ACR output. However, the \(v_d\) and \(v_q\) are the unknown voltages. The reference and the response of each inverter are not the same because the dc voltage of each inverter pulsates. Although the motor winding voltage is measured using a voltage sensor, accuracy is not guaranteed because of the voltage of the switching frequency component. It is necessary to estimate the motor winding voltage. The \(v_d\) and \(v_q\) are estimated by using the relationship of equations (1) and (2). The \(d\)- and \(q\)-axis voltages of each inverter are estimated by passing the \(d\)- and \(q\)-axis voltage references of each inverter through the variable limiter with each detected dc voltage. From equation (11), the instantaneous active power of Inv.2 pulsates significantly because the Inv.2 outputs the harmonic voltage. The pulsating power component is expressed as follows:

\[
\Delta P_m = \frac{3}{2}(\Delta v_d i_d^{ref} + \Delta v_q i_q^{ref}) \cdots \cdots (14)
\]

The compensating capacitor voltage PI control uses the instantaneous active power of Inv.2 as the controller output, and thus it is necessary to add the pulsating power component to the controller output. In the proposed control method, the instantaneous active power reference of Inv.2 is expressed as follows:

\[
p_2^{ref} = p_{I2} + \Delta P_m \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots (15)
\]

Here, \(p_{I2}\) is the compensating capacitor voltage PI controller output. The proposed method reduces the harmonic voltage superimposed on the motor winding voltage caused by the voltage saturation of Inv.1 due to grid disturbances. Therefore, the proposed control method achieves both an extension of the motor speed range and a reduction of torque ripple under grid disturbances.

### 3.2 Proposed Control Block

Fig. 5 shows the proposed control block diagram. The proposed method is based on a motor speed PI control for the open-end winding IPMSM. The \(d\)-axis motor current reference is generated by the maximum torque per ampere (MTPA) control method. The \(d\)- and \(q\)-axis motor winding voltage references are determined by the ACR output, and thereafter decoupling control is performed. The \(d\)- and \(q\)-axis voltage references for Inv.1 and Inv.2 are determined based on the proposed method. The proposed method is effective in reducing harmonic voltages under unpredictable grid disturbances, because it does not use a phase-locked loop (PLL) and the repeating information in the grid voltage disturbances or the torque ripple. The gain of the compensating capacitor voltage controller is determined by a heuristic based on a simulation.

### 4. Proposed Design Method of Electrolytic Capacitor-less Dual Inverter

#### 4.1 Guideline

The design method of capacitors used in motor drive systems with electrolytic capacitor-less dual inverters to extend the motor speed range and reduce motor winding harmonic voltage under grid disturbances has not been studied in the past. The film capacitors used in the electrolytic capacitor-less dual inverter are selected for their optimum capacitance to reduce size and weight. The dc-link capacitor \(C_{dc1}\) is employed to remove the current of the switching frequency component superimposed on the input current of Inv.1. The compensating capacitor \(C_{dc2}\) is employed to use the charging and discharging.

#### 4.2 Dc-link Capacitor

The design of the dc-link capacitor \(C_{dc1}\) is based on Ref. (10). The role of \(C_{dc1}\) is a filter. The filter consists of the line impedance and \(C_{dc1}\). This filter is used to remove the switching frequency components superimposed on the input current of Inv.1. The dc-link capacitor capacitance must satisfy the following inequality to stabilize the motor drive system.

\[
C_{dc1} > \frac{L_g P_I}{r_g V_{dc1}} \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots (16)
\]

Here, \(r_g\) and \(L_g\) are the line impedance, \(P_I\) is rated load power, and \(V_{dc1}\) is the average dc-link capacitor voltage. In this study, \(r_g = 0.01 \Omega\), \(L_g = 4 \mu H\), \(P_I = 1.0 \text{ kW}\), \(V_{dc1} = 269 \text{ V}\). From equation (16), the dc-link capacitor capacitance should be set larger than 5.5 \(\mu F\). Consequently, the \(C_{dc1}\) is selected as 10 \(\mu F\).
4.3 Maximum Compensating Power and Energy

From equation (10), the $C_{dc2}$ depends on the compensating energy. Consequently, the $C_{dc2}$ is designed with maximum compensating energy determined by the maximum instantaneous active power of Inv.2 (maximum compensating power) and the compensating frequency. The maximum compensating power is determined by the worst case of grid disturbances. In this study, the worst case for THD of the grid voltage is set to 25% according to the specifications of power supply environmental test equipment. The worst case must be set appropriately. If the grid voltage disturbances are larger than the set worst case, the responses of the torque and the compensating capacitor voltage are unstable due to matching the voltage limit of Inv. 2. The harmonic voltage superimposed on the output voltage of Inv.2 is significantly affected by the dc-link voltage $V_{dc2}$. Fig. 6 shows the simulation analysis results of the relationship between each harmonic order and the minimum dc-link voltage under the grid disturbance (THD 25%). When the seventh or eleventh harmonic voltage is superimposed on the grid voltage, those minimum dc-link voltages are smaller than the minimum dc-link voltages that are superimposed by other harmonic orders. Consequently, the saturated voltage is most likely to occur under grid voltage with superimposed seventh or eleventh harmonic voltages. Fig. 7 shows the MTPA-controllable motor operating characteristics that can be operated without the saturated voltage, considering the minimum dc-link voltage of Fig. 6 based on the motor parameters in Table 1. When the IPMSM outputs 1.0 kW (2.0 Nm, 4766 r/min) of rated power under a normal grid voltage, the maximum machine output power (load condition without saturated voltage, 2.0 Nm, 3115 r/min) is 652 W under the grid disturbance (seventh harmonic) with the same torque conditions. The proposed method is required
The second harmonic voltage. However, the Wc The equation (17), Table 2, and Fig. 8 under grid disturbances. The comparison of maximum compensating energy considering monics superimposed on the grid voltage. Fig. 9 shows the main ripple frequency is lowest at the second or fourth harmonic and dc-link voltage main ripple frequency. The dc-link analysis results of the relationship between each harmonic to compensate for 348 W, which is the difference in the machine output power under a normal grid voltage and the machine output power under the grid disturbance. Fig. 8 shows the relationship between the compensating power and torque when the second or seventh harmonic voltage is superimposed on the grid voltage. The compensating power in Fig. 8 is calculated as the difference between the machine output power under a normal grid voltage and the machine output power under the distorted grid voltage by calculating the motor speed that can be driven with the same torque and without the saturated voltage. The maximum compensating power is 348 W. In equation (10), the maximum compensating energy expressed as follows:

\[ W_{c2} = \int_0^\omega P_2 \sin(\omega t) dt = \frac{2P_2}{\omega} \]  \hspace{1cm} (17)

Here, \( P_2 \) is the maximum compensating power, and \( \omega \) is the compensating angular frequency. The \( W_{c2} \) depends on the \( \omega \). Equation (17) implies that the \( W_{c2} \) is larger with lower \( \omega \) when the \( P_2 \) is the same. Table 2 shows the simulation analysis results of the relationship between each harmonic order and dc-link voltage main ripple frequency. The dc-link main ripple frequency is lowest at the second or fourth harmonics superimposed on the grid voltage. Fig. 9 shows the comparison of maximum compensating energy considering equation (17), Table 2, and Fig. 8 under grid disturbances. The \( P_2 \) is larger with the seventh harmonic voltage than with second harmonic voltage. However, the \( W_{c2} \) is larger with second harmonic voltage than with seventh harmonic voltage. Therefore, the \( C_{dc2} \) must be designed considering the grid disturbance caused by the second harmonic voltage.

### 4.4 Compensating Capacitor Voltage Reference Range

The maximum compensating capacitor voltage is determined by the rated voltage of the IGBTs used in Inv. 2. In this study, IGBTs with a rated voltage of 600 V are used, thus the maximum compensating capacitor voltage is set to 400 V. The minimum compensating capacitor voltage depends on the reactive power of Inv.2 and the harmonic voltage output by Inv.2. The instantaneous reactive power reference of Inv.2 is expressed as follows:

\[ q_2^{\text{ref}} = \frac{3}{2}(v_{d2}^{\text{ref}}i_d^{\text{ref}} - v_{q2}^{\text{ref}}i_q^{\text{ref}}) \]  \hspace{1cm} (18)

From equations (11) and (18), considering \( p_2^{\text{ref}} = 0 \) and \( q_2^{\text{ref}} = q_m^{\text{ref}} \), the \( d-\) and \( q-\)axis voltage references of Inv. 2 to supply the reactive power required to drive the IPMSM are expressed as follows:

\[ v_{d2}^{\text{ref}} = -\frac{2}{3}\frac{q_m^{\text{ref}}}{L_d} i_d^{\text{ref}} \]  \hspace{1cm} (19)

\[ v_{q2}^{\text{ref}} = \frac{2}{3}\frac{q_m^{\text{ref}}}{L_q} i_d^{\text{ref}} \]  \hspace{1cm} (20)

The \( q_m^{\text{ref}} \) is expressed as follows from Ref. (11):

\[ q_m^{\text{ref}} = \frac{3}{2}\omega_e(L_d i_d^{\text{ref}} + L_q i_q^{\text{ref}} + \Psi_{d0} i_d^{\text{ref}}) \]  \hspace{1cm} (21)

Here, \( \omega_e \) is the electric angular frequency, \( \Psi_{d0} \) is the linkage flux, \( L_d \) and \( L_q \) are the \( d- \) and \( q- \)axis inductances, respectively. From equations (19), (20), and (21), the \( d- \) and \( q- \)axis voltage references of Inv. 2 to supply the reactive power required to drive the IPMSM are determined from the motor parameters, and rated load conditions (motor speed and load torque). The rated conditions are determined from motor operating characteristics shown in Fig. 7 and Table 1. There is the following relationship between the \( d- \) and \( q- \)axis voltage references of Inv. 2 and the compensating capacitor voltage.

\[ \frac{V_{dc2}}{2} \geq \sqrt{v_{d2}^{\text{ref}}^2 + v_{q2}^{\text{ref}}^2} \]  \hspace{1cm} (22)

From equations (19) to (22), the minimum compensating capacitor voltage is calculated from the peak of the \( d- \) and \( q- \)axis voltage references of Inv. 2 and the reactive power required to drive the IPMSM. Thus, the compensating capacitor voltage must be higher than 116 V to supply the reactive power required to drive the IPMSM. In addition, the difference between the average dc-link voltage of 269 V under a normal grid voltage and the minimum dc-link voltage

### Table 2. Relationship between each harmonic order and dc-link voltage main ripple frequency

<table>
<thead>
<tr>
<th>Ripple frequency</th>
<th>150 Hz</th>
<th>300 Hz</th>
<th>450 Hz</th>
<th>600 Hz</th>
</tr>
</thead>
<tbody>
<tr>
<td>Harmonic order</td>
<td>2, 4</td>
<td>3, 5, 6, 7</td>
<td>9, 12</td>
<td>8, 10, 11, 13</td>
</tr>
</tbody>
</table>

Fig. 8. Relationship between compensating power and torque under the distorted grid voltage

Fig. 9. Comparison of maximum compensating energy under grid disturbances
of 178 V under the grid voltage with superimposed 7th harmonic voltage is 91 V. The voltage saturation under the grid disturbance is caused by this shortage voltage. Consequently, the minimum compensating capacitor voltage \( V_{dc(2\text{min})} \) in the proposed control method is 207 V. In this study, the compensating capacitor is selected as a film capacitor (Panasonic, EZPE series, 800 V). From the datasheet, the maximum compensating capacitor voltage ripple is set to 100 V. Considering the maximum compensating capacitor voltage and the maximum voltage ripple, the maximum ripple rate \( \varepsilon \) is expressed as follows:

\[
\varepsilon < \frac{\Delta V_{dc2}}{V_{dc(2\text{ave})}} = \frac{2\Delta V_{dc2}}{V_{dc(2\text{max})}} = \frac{2\Delta V_{dc2}}{V_{dc(2\text{max})} - \Delta V_{dc2}} \tag{23}
\]

Here, \( \Delta V_{dc2} \) is the peak-to-peak compensating capacitor voltage, \( V_{dc(2\text{ave})} \) is the average compensating capacitor voltage, and \( V_{dc(2\text{max})} \) is the maximum compensating capacitor voltage. From equation (23), \( V_{dc(2\text{max})} = 400 \text{V}, \Delta V_{dc2} = 100 \text{V}, \) and the \( \varepsilon \) must be smaller than 0.286. In this study, the \( \varepsilon \) is set to 0.25. The maximum and minimum compensating capacitor voltages are expressed as follows:

\[
V_{dc(2\text{max})} = V_{dc(2\text{ave})} + \frac{\Delta V_{dc2}}{2} = 400 \tag{24}
\]

\[
V_{dc(2\text{min})} = V_{dc(2\text{ave})} - \frac{\Delta V_{dc2}}{2} = 207 \tag{25}
\]

The compensating capacitor voltage reference must be determined in the range of equations (24) and (25). From equations (23) to (25), the \( V_{dc2\text{ref}} \) (= average compensating capacitor voltage) must satisfy the following inequalities:

\[
V_{dc2\text{ref}} \geq \frac{2V_{dc(2\text{max})}}{2 - \varepsilon} = \frac{2 \cdot 207}{2 - 0.25} = 237 \tag{26}
\]

\[
V_{dc2\text{ref}} \leq \frac{2V_{dc(2\text{max})}}{2 + \varepsilon} = \frac{2 \cdot 400}{2 + 0.25} = 356 \tag{27}
\]

### 4.5 Compensating Capacitor

The relationship between the compensating capacitor voltage and compensating capacitor is expressed as follows:

\[
W_{c2} = \frac{C_{dc2}}{2}(V_{dc(2\text{max})}^2 - V_{dc(2\text{min})}^2) \tag{28}
\]

From equations (17), (23), (24), (25), and (28), the compensating capacitor \( C_{dc2} \) must satisfy the following inequality:

\[
C_{dc2} > \eta \frac{2P_2}{\varepsilon \omega V_{dc(2\text{ave})}^2} \tag{29}
\]

Here, \( \eta \) is the margin. In this study, the \( \eta \) is set to 115% from Ref. (12), and \( P_2 = 239 \text{W}, \varepsilon = 0.25, \omega = 300\pi \text{rad/s}. \) Fig. 10 shows the relationship between the compensating capacitor voltage reference and the compensating capacitor capacitance. The capacitance must be at least 40.5 \( \mu \text{F} \) when the reference is set to 240 V. In this study, the capacitance is installed with 50 \( \mu \text{F}. \)

### 5. Experimental Results

#### 5.1 Experimental Conditions

Experiments are performed to verify the effectiveness of the proposed control method. The control and system parameters are listed in Table 3. The dc-link capacitor and the compensating capacitor are based on the proposed design method. Fig. 11 shows the experimental setup. Grid disturbances are emulated using power supply environmental test equipment (NF 8484). The proposed method is implemented using a Texas Instruments TM320 floating-point digital signal processor. The load torque is supplied by controlling the load motor using a load inverter. The torque transducer uses TP-50KCM. The torque meter uses DPM-952A, DL850 is used to observe the waveform. The results of the proposed method are compared with the conventional motor speed range extension method without an anti-voltage saturation treatment referred to in Ref. (8). The system parameters of the proposed method and the conventional method are the same. Sections 2 and 4 analyzes the impact of harmonic orders of grid disturbances on the motor operating characteristics. The seventh harmonic is the most affected harmonic order for the motor operating characteristics. The experiments are mainly performed to evaluate the motor current and the shaft torque of the
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5.2 Effectiveness of Proposed Method

Fig. 12 shows the experimental results of the electrolytic capacitor-less dual inverter using each control method under the input voltage with superimposed seventh harmonic voltage. The experiments using the proposed control method under the grid voltage with superimposed second harmonic voltage are performed to verify the effectiveness of the proposed design method of electrolytic capacitor-less dual inverter. Fig. 13 shows the experimental results of the electrolytic capacitor-less dual inverter using the proposed control method under the input voltage distorted by second harmonic voltage at 2.0 Nm and 4650 r/min. The proposed method designs the compensating capacitor capacitance and voltage reference under the grid voltage with superimposed second harmonic voltage because the compensating energy is larger than other harmonic orders. Fig. 13 shows that the dc-link voltage ripple is increased by 208% under the input voltage with superimposed second harmonic voltage. The proposed method reduces the shaft torque ripple under grid disturbances by 66.9% compared with the conventional method. Therefore, the proposed method reduces the torque ripple under grid disturbances in the load condition with extended motor speed range. The proposed method simultaneously extends the motor speed range and reduces the torque ripple under grid disturbances.

5.3 Effectiveness of Proposed Design Method

The harmonic current superimposed on the output voltage of the inverter because the electrolytic capacitor-less dual inverter does not include an energy buffer in the power converter. The harmonic voltages are superimposed on the output voltage in Inv.1 is due to voltage saturation caused by the grid disturbance. Furthermore, the conventional method increases only the voltage amplitude of the motor winding voltage, thus cannot cope with grid disturbances. The increased peak-to-peak motor phase current might exceed the rated current and demagnetization resistance, thereby reducing the reliability of the motor drive system. The shaft torque ripple increases by 170% owing to the grid disturbance. The harmonic current superimposed on the motor current causes torque ripple, which resulted in mechanical vibration.
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Fig. 13. Experimental results of proposed method under the input voltage distorted by second harmonic voltage at 2.0 Nm and 4650 r/min

Fig. 14. FFT analysis results of the motor U-phase current at 2.0 Nm and 4650 r/min

(a) Conventional method.

(b) Proposed method.

Fig. 15. Comparison of torque ripple at 300 Hz

Fig. 16 shows the simulation results of the d- and q-axis motor current under the input voltage distorted by seventh harmonic voltage at 2.0 Nm and 4650 r/min. The d- and q-axis motor current in the conventional method pulsate significantly due to the grid disturbances. The simulation results of the proposed method show that the d- and q-axis motor current ripple are reduced by 58.6 and 80.9%, respectively, compared to the conventional method. Fig. 17 shows the experimental results of the d- and q-axis motor current under the input voltage distorted by 7th harmonic voltage at 2.0 Nm and 4650 r/min. The experimental results of the proposed method show that the d- and q-axis motor current ripple are reduced by 58.7 and 78.4%, respectively, compared to the conventional method. The reduction rates of the simulations and experiments are approximately the same. Therefore, the effectiveness of the proposed control method is confirmed by simulations and experiments.

5.5 dq-axis Motor Current Ripple

The d- and q-axis motor current ripple are reduced by 58.6 and 80.9%, respectively, compared to the conventional method. The simulation results of the proposed method show that the d- and q-axis motor current ripple are reduced by 58.7 and 78.4%, respectively, compared to the conventional method. The reduction rates of the simulations and experiments are approximately the same. Therefore, the effectiveness of the proposed control method is confirmed by simulations and experiments.

5.6 Motor Speed Characteristic

Fig. 18 shows the motor speed characteristics at 2.0 Nm under the input voltage distorted by seventh harmonic voltage. The torque ripple increases rapidly when the motor speed increases above 291 rad/s.
600 r/min in the conventional method. As the motor speed increases, the Inv.1 operates with the high modulation index. The torque ripple occurs due to voltage saturation of Inv.1 under grid disturbances. The proposed method achieves the lower torque ripple under grid disturbances even when the motor speed is above 3600 r/min. The proposed method reduces the torque ripple at 300 Hz by up to 91.0% compared with the conventional method under the input voltage distorted by seventh harmonic voltage. In addition, the proposed method reduces the torque ripple at 300 Hz by up to 91.0% compared with the conventional method. The proposed method reduces the voltage ripple at 300 Hz by up to 91.0% compared with the conventional method. The proposed method reduces the voltage ripple at 300 Hz by up to 91.0% compared with the conventional method.

6. Conclusions

This paper proposes a control method for torque ripple using an open-end winding IPMSM driven by an electrolytic capacitor dual inverter under grid disturbances. The proposed control method is based on a novel anti-voltage saturation treatment. The proposed method achieves both an extension of the motor speed range and a reduction of torque ripple under grid disturbances. The effectiveness of the proposed method based on the proposed design method is demonstrated by experimental results using an electrolytic capacitor-less dual inverter and an open-end winding IPMSM. The proposed method reduces the harmonic motor current that causes torque ripple by up to 85.6% compared with the conventional method under the input voltage distorted by seventh harmonic voltage. In addition, the proposed method reduces the torque ripple at 300 Hz by up to 91.0% compared with the conventional method. The proposed method reduces the torque ripple at 300 Hz by up to 91.0% compared with the conventional method. The proposed method reduces the torque ripple at 300 Hz by up to 91.0% compared with the conventional method. The proposed method reduces the torque ripple at 300 Hz by up to 91.0% compared with the conventional method.

In the future, the anti-voltage saturation control design considering transient response in a dual inverter drive system will be considered.

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