Controlling Torque Ripple Vibration in a Single-Phase PFC
Using the Neutral Point of an IPMSM

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(Manuscript received Jan. 00, 20XX, revised May 00, 20XX)

This paper proposes a method to reduce the torque ripple vibration of an integrated power factor correction (PFC) converter by using the zero-phase inductance of an interior permanent magnet synchronous motor (IPMSM) when the motor is not running. When an integrated PFC converter is applied to electric vehicle (EV) chargers, it is necessary to reduce the torque oscillation of the motor caused by the charging current. In an integrated PFC converter using an IPMSM, the zero-phase current and magnetomotive force harmonics of the permanent magnet cause torque vibration. In the proposed control method, the generated torque is estimated using the back EMF table, and the phase current is controlled to cancel it. The proposed method can be used to suppress torque vibration irrespective of the mechanical angle of the rotor. The effectiveness of the proposed method is experimentally verified. Using the proposed method, the torque vibration can be suppressed by up to 91.3% as compared to that in the case of the conventional control method. Furthermore, a power factor of 99.9% is achieved under a load condition of 1 kW.

Keywords: Integrated battery charger, interior permanent magnet synchronous motor (IPMSM), torque ripple

1. Introduction

In recent years, the use of EVs has become widespread because of concerns regarding environmental problems such as air pollution caused by fossil fuels. Furthermore, the energy stored in a vehicle's battery can also be used to realize smart V2G (vehicle to grid) such as grid peak shaving and frequency adjustment(1), (2). Therefore, research on battery electric vehicles (BEVs) is being conducted actively.

There are two main types of chargers for BEVs: off-board battery chargers installed outside the vehicle and on-board battery chargers (OBCs) that are built into the vehicle(3). Off-board battery chargers can be designed to be suitable for fast charging. However, users need to go to the charging station. In addition, infrastructure needs to be developed to set up stations. Meanwhile, OBCs are built into vehicles. Therefore, no infrastructure is required. Moreover, these can be connected directly to a single-phase or three-phase power grid for charging. OBCs can be used to charge BEVs with the existing grid infrastructure. However, the charging capacity is limited by the cost, volume, and weight of the vehicle(4)–(6).

To overcome these limitations, integrated OBCs (IOCs) have been studied. These integrate the common elements of OBCs and motor drive systems. Because an OBC is connected directly to the grid, a high power-factor operation that satisfies the current harmonics regulation is required. Furthermore, a PFC circuit is necessary. In the IOC configuration, the motor winding is used as a filter to increase the inductance on the grid side.

A method wherein an additional rectifier is used is applied to convert the grid voltage to DC. Then, the motor winding is used as a boost inductor in an interleaved PFC circuit(7), (8). In these methods, the synthetic magnetomotive force (MMF) generated by the DC current in the motor windings is fixed at a certain rotor position. The torque is zero when the angle between the d-axes of the rotor is 0 deg or 180 deg. However, torque vibration occurs in most other positions in these methods. Therefore, these methods are unsuitable for recifier circuits of V2G chargers. To suppress torque oscillation, control methods using multiple inverters have been investigated for motor structures with multiphase windings(9), (10). However, multiphase motors with complex windings are uncommon among EV systems.

Another method is to use an inverter to develop a full-bridge PFC circuit with a pair of inverter legs and the remaining leg(11). In this case, the grid is inserted in the path between the central point of one leg and the motor winding. This circumvents the need for additional components for the system. However, this method requires the re-connection of the grid by the rotor position. In addition, the power is limited by different phase currents.

A few researchers have proposed accessing the neutral point of the motor and using the zero-phase current to perform PFC operation(12). In (12), the motor was considered separately equivalent circuit: the positive-phase sequence and zero-phase sequence. A positive-phase equivalent circuit is used to drive the motor, and a zero-phase equivalent circuit is used as the PFC circuit. In the PFC circuit, the leakage inductance of the motor appears as a zero-phase inductance. This is utilized as a grid filter. The alternating current flows through the motor winding in the PFC. The synthetic MMF using the zero-phase current is zero. Thereby, no torque is generated. An identical topology is considered in the IOC of EV systems(13), (14). However, considering the influence of the magnetomotive force distribution and spatial harmonics in IPMSMs, torque is generated by the zero-phase currents as well. Zero-phase currents generate torque oscillations.

Mechanical clutches are usually removed from EV traction machine to reduce weight and cost(15). In addition, the traction motor...
and wheels are coupled by a transmission. In addition, the traction motor and wheels are coupled by a transmission. In consequence, there is concern that torque vibration generated during charging may be transmitted directly to the transmission, causing noise and vibration.

As mentioned above, most research on conventional integrated PFCs has focused on the use of the motor winding as the inductance of the power converter. There has been negligible research on control methods for suppressing torque vibration that focus on the structure of the IPMSM and spatial harmonics. Therefore, this study proposes a method for suppressing the torque oscillation in an integrated PFC using the motor windings of an IPMSM. The focus is on the spatial harmonics when the motor is stopped.

The remainder of this paper is organized as follows. First, an overview of the problem and an analysis of the torque generated during the charging operation are presented. Subsequently, the strategy proposed for calculating the phase current command that can suppress the torque vibration from the back EMF is explained. The effectiveness of the proposed method in suppressing torque vibration is verified experimentally.

2. Integrated On-board Battery Charging System

2.1 Conventional System

Fig. 1 shows a conventional battery charging system. The system is separated into a motor drive system and an on-board battery charger system. As shown in Fig. 1, the on-board charger is typically independent of the motor drive system. This requires a large capacitor for the DC link for decoupling. In the PFC circuit, an inductor is connected in series to the grid as a filter. Passive components such as capacitors and inductors increase the size and weight of the system.

2.2 Integrated On-board Battery Charging System

Fig. 2 shows the configuration of the battery charging system that is the focus of this study. The system consists of an IPMSM as a traction machine, a three-phase inverter, an additional leg, and two mechanical switches that vary the operation mode. In the traction mode shown in Fig. 2 (a), SW1 and SW2 are turned off, and the system operates as a normal three-phase inverter-driven IPMSM system. In the charging mode shown in Fig. 2 (b), SW1 and SW2 are turned on. The PFC consists of a three-phase inverter, additional leg, and the winding of the IPMSM. In the charging mode, the IPMSM is used as a filter in the PFC circuit. The current demanded from the grid during the PFC operation can be expressed as follows:

\[ i_g = \frac{i_u}{\sqrt{2}} \sin(\omega_g t), \tag{1} \]

where \( \omega_g \) denotes the angular frequency, \( t \) denotes the time variable, and \( i_g \) denotes the maximum value. The phase current of the IPMSM \( (i_u, i_v, i_w) \) and the grid current \( (i_u) \) are related as follows:

\[ i_g = i_u + i_v + i_w. \tag{2} \]

The phase current of the IPMSM should satisfy (2). The switching of the additional leg is synchronized with the polarity of the grid voltage. This indicates that the potential of the neutral point is fixed according to the polarity of the grid. Thus, the voltage applied to each winding is controlled by the inverter-side leg.

2.3 Analysis of torque vibration

The torque produced in the rotor by the current for a phase can be expressed as

\[ \tau_x = i_x \frac{\partial \psi_{xM}}{\partial \theta_e} + \frac{1}{2} \frac{\partial L_x}{\partial \theta_e} \quad (x = u, v, w), \tag{3} \]

where \( i_x \) is the phase current, \( \psi_{xM} \) is the interlinkage flux of \( x \)-phase by a permanent magnet, \( L_x \) is the self-inductance of \( x \)-phase, and \( \theta_e \) is the electrical angle of the rotor. It is evident from (3) that the torque has two main components: the part caused by the permanent magnet and that caused by the self-inductance. Assuming charging mode, the rotor of the IPMSM is fixed at a certain angle.

Consequently, all the parameters \( i_x \) are constants. If a sinusoidal wave synchronized with the grid is assumed to be the phase current, the torque can be separated as follows:

\[ \begin{align*}
\tau_{xM} &= i_x \sin(\omega_g t) \frac{\partial \psi_{xM}}{\partial \theta_e}, \\
\tau_{xR} &= \frac{i_x}{2} (1 - \cos(2\omega_g t)) \frac{\partial L_x}{\partial \theta_e}.
\end{align*} \tag{4} \]

Where, \( i_x \) is the maximum value of the \( x \)-phase current.

From (4), the frequency of \( \tau_{xM} \) is equal to that of the grid, and the frequency of the reluctance torque is two times that of the grid. The amplitude of the torque depends on the phase current and rotor angle.

Next, the magnet torque \( \tau_M \) when the current is applied to the three phases can be expressed as

\[ \tau_M = i_u \frac{\partial \psi_{uM}}{\partial \theta_e} + i_v \frac{\partial \psi_{vM}}{\partial \theta_e} + i_w \frac{\partial \psi_{wM}}{\partial \theta_e}, \tag{5} \]

where \( \psi_{uM,vM,wM} \) is the interlinkage flux of each phase and is given by
\[
\begin{align*}
\psi_{uM} &= \sum_{k=1}^{n} \Psi_{MKth} \sin(k\theta_e + \phi_{uMKth}) \\
\psi_{vM} &= \sum_{k=1}^{n} \Psi_{MKth} \sin\left(k\left(\theta_e - \frac{2}{3}\pi + \phi_{vMKth}\right)\right) \\
\psi_{wM} &= \sum_{k=1}^{n} \Psi_{MKth} \sin\left(k\left(\theta_e + \frac{2}{3}\pi + \phi_{wMKth}\right)\right)
\end{align*}
\] (6)

Where, \(\Psi_{MKth}\) is the maximum value of the \(k_{th}\) harmonic and \(\phi_{uMKth}\) is the phase difference of the \(k_{th}\) harmonic. When only the fundamental component is considered, the torques generated by the phase currents cancel each other because the windings have a phase difference of \(2\pi/3\) each. For this reason, each leg of the three-phase inverter is given an equally divided input current command. However, spatial harmonic components are included in the magnetic flux and other components of the IPMSM. The magnet torque produced when the harmonics are considered can be expressed as

\[
\tau_{MN} = \frac{i_g}{3} n \sum_{k=1}^{n} 3k\Psi_{MKth} \cos(3k\theta_e) .
\] (7)

According to (7), the amplitude of \(\tau_{MN}\) is zero when the rotor position \(\theta_e\) is \(n\pi/2\) (\(n = 1, 2, 3, \ldots\)).

3. Control Method for Torque Reduction

From (5), the torque generated by the phase currents is obtained by the product of the phase currents and the number of phase cross-fluxes in each phase. Therefore, if \(\partial\psi_{xM}/\partial\theta_e\) in (5) can be estimated, the rotor torque can be reduced to zero by controlling the phase current. The estimation method is described in the next section.

### 3.1 Estimation of Torque Vibration by Back-EMF

For any \(x\)-phase of the IPMSM, the back EMF \(v_x\) can be expressed as

\[
v_x = \frac{d\psi_{xM}}{dt}.
\] (8)

If the angular velocity of the rotor \(\omega_m\) is constant, the relationship between the rotor position \(\theta_e\) and time \(t\) can be expressed as

\[
p\omega_m = \frac{d\theta_e}{dt},
\] (9)

Where, \(p\) is the number of pole pairs.

It is assumed that \(\psi_{xM}\) depends only on \(\theta_e\) and not on the phase current. Using (9), the rotor position \(\theta_e\) derivative of the interlinkage flux \(\psi_{xM}\) in (8) can be expressed as follows:

\[
\frac{v_x}{p\omega_m} = K_x = \frac{d\psi_{xM}}{d\theta_e}.
\] (10)

Consequently, the magnet torque (expressed as (5)) can be estimated as follows:

\[
\tilde{T}_M = \sum_{x=u,v,w} \frac{v_x}{p\omega_m} I_x = \sum_{x=u,v,w} K_x I_x.
\] (11)

Fig. 3 shows the estimated magnet torque amplitude generated by the simulation when the current amplitude of each phase is 1 A for the model shown in Table 1 and Fig. 4.

The back-EMFs \(v_u, v_v, v_w\) are determined by measurement, and the amplitude of each phase current can be controlled arbitrarily. Therefore, the torque vibration can be reduced by providing each phase current command such that \(\tilde{T}_{SM}\) in (11) becomes zero.

### 3.2 Method of Determining Current Command

The conditions that need to be satisfied to determine the current command for each phase are as follows:

i. For PFC operation, equation (2) must be satisfied.
ii. To reduce the magnet torque vibration, $\tilde{T}_{mM} = 0$ must be satisfied in equation (11).

iii. The three-phase current is synchronized with the grid, and there is no phase difference between phases.

Given the above conditions, the only variable to be determined is the amplitude of the three-phase current $I_x$. The constraint is as follows:

$$I_g = I_u + I_v + I_w,$$

(12)

$$0 = I_u K_u + I_v K_v + I_w K_w,$$

(13)

Because there are three controllable variables and two constraints, the current command can be determined uniquely by adding another constraint.

### 3.3 Current Command for Minimizing Copper loss

This section describes the method of calculating the current command that satisfies the aforementioned constraints and minimizes the copper loss of the IPMSM simultaneously.

The copper loss $P_c$ caused by the phase current is expressed as

$$P_c = \frac{R}{2} (I_u^2 + I_v^2 + I_w^2),$$

(14)

where $R$ is the phase resistance. $P_c$ can be expressed as a function of $I_u, I_v$ by using (12) to eliminate $I_w$. The copper loss in the windings is minimized when the phase currents are equal. Fig. 5 shows a graph of the phase currents and copper losses in the windings. From (12), assuming that $I_g$ follows the command, $I_w$ is determined uniquely by $I_u$ and $I_v$. Therefore, the phase currents are considered only in terms of the magnitudes of $I_u$ and $I_v$. Furthermore, from (13), the combination of $I_u$ and $I_v$ that satisfies the condition to suppress the torque vibration is restricted as follows:

$$I_v = \frac{K_u - K_w}{K_v - K_w} I_u - \frac{K_w}{K_v - K_w} I_g,$$

$$= \alpha I_u + \beta.$$

(15)

From (15), the combination of phase currents that can reduce torque vibration is expressed as a linear function of the slope $\alpha$ and the intercept $\beta$ calculated using $K_v$. The combinations of values that reduce the torque vibration are on the color boundary of the plane (see Fig. 5). The phase-current command that minimizes the copper loss is determined as a linear combination of $I_u$ and $I_v$. Table 2 and Table 3 list the system parameters and the IPMSM test parameters, respectively.

### Table 2 System parameters.

| Parameter                         | Value       |
|-----------------------------------|-------------|
| AC power supply voltage          | 100 V rms, 50 Hz |
| DC output voltage reference      | 175 V       |
| DC-link capacitor                | 1300 µF     |
| Carrier frequency                | 50 kHz      |
| Dead time                        | 500 ns      |
| Bandwidth of ACR                 | 2000 Hz     |
| Bandwidth of AVR                 | 4 Hz        |

### Table 3 Parameters of the IPMSM test.

| Parameter                         | Value       |
|-----------------------------------|-------------|
| Stator winding resistance         | 0.788 ohm   |
| Number of poles                   | 4           |
| Number of slots                   | 24          |
| d-axis inductance                 | 10.2 mH     |
| q-axis inductance                 | 20.1 mH     |
| zero-phase inductance             | 1.35 mH     |
| Magnet flux linkage               | 0.137 Wb    |
copper loss can be determined analytically by substituting (12) and (15) into (14) and computing the minimum value.

4. Experimental Results

Experiments were performed based on the proposed control method to verify the effectiveness of torque vibration suppression. To measure the shaft torque, one end of the torque meter was fixed mechanically, and the other end was connected to the output shaft of the motor. The experimental conditions and parameters of PFC are listed in Table 2. The experiments were performed using an
open-end winding IPMSM (distributed winding). Table 3 shows the parameters of the IPMSM test. As described earlier, the zero-phase inductance of the IPMSM is used as a filter in the PFC circuit. The zero-phase inductance is calculated by measuring the current when applying a 50 Hz AC voltage through the three-phase terminals and the neutral point of the IPMSM. At the time of measurement, the three-phase terminals are shorted. A digital signal processor (Texas Instruments TMS320F28379D) was used as controller. A control block diagram of the system is shown in Fig. 6.

The system is based on DC voltage control with an IPMSM as the grid filter. The effective value of the input current command is determined by an automatic voltage regulator (AVR). Furthermore, it is converted to a sine wave synchronized with the grid voltage based on the phase signal of the phase-locked loop (PLL). The additional leg is switched according to the polarity of the grid voltage. A current control system is provided for each phase, and each system controls the phase current independently. In the conventional method, the phase current command is the value obtained by dividing the input current command output from the AVR into three equal parts. The amplitude of each phase current command is given by the equation

\[ I_u^* = I_v^* = I_w^* = \frac{I_g^*}{3}, \]

where \( I_g^* \) is the amplitude of the input current command output from the AVR. In the proposed method, the current command of each phase is determined according to the back EMF of the motor to reduce torque vibration.

### 4.1 Performance of PFC Operation
First, the fundamental PFC operation of both control methods was verified. Fig. 8 shows a comparison of the experimental results of the conventional and proposed methods. Two rotor positions are represented. For each of these, the grid voltage \( v_g \), grid current \( i_g \), motor phase current \( i_u, v, w \), and torque \( \tau \) were measured from the shaft of the motor. In the experiments, the grid voltage RMS was fixed at 100 V, and the DC output power was fixed at 1 kW by the electric load.

The grid current \( i_g \) was maintained in phase with the grid voltage by the controller to achieve unity power factor. It is also verified from Fig. 8 (b), (d) that the phase currents are controlled independently in the proposed method. The DC voltage was maintained at 173 V by AVR. Table 4 shows the measurement results of power factor (PF) and total harmonic distortion (THD), which indicate the performance of PFC. The PF and THD were comparable for both.

Fig. 9 shows the system efficiency, PF, and THD for different rotor positions for each output power. The efficiency of the rotor position at 60 deg showed an overall trend. In the proposed method, each phase current command value is assigned a different amplitude for torque reduction. In addition, because the combination of the phase current command is selected to minimize copper loss, the phase current command differs significantly when the rotor angle is 30 deg, and has an almost equal amplitude when the rotor angle is 60 deg. Therefore, this difference manifests as efficiency.

### 4.2 Performance of Torque Suppression
In the conventional method, a torque peak of approximately 3.46 Nm is verified at 30 deg (where, theoretically, the maximum torque occurs). This causes vibrations and noise during the charging process. However, it was verified that the torque amplitude was suppressed when the proposed method for suppressing torque vibration was applied. The peak-to-peak value of torque vibration was 2.27 Nm when the rotor angle was 15 deg. It was suppressed to 0.39 Nm by the proposed method. At 30 deg (where, theoretically, the torque is the highest, it was reduced from 3.46 Nm to 0.30 Nm. Fig. 10 summarizes the peak-to-peak torques of the conventional and proposed methods. The suppression rates were 82.8% and 91.3% at 15 deg and 30 deg, respectively.
5. Conclusion

This paper proposed a torque oscillation suppression method for an integrated PFC using motor windings. The focus is on the spatial harmonics of an IPMSM. The proposed control method generates current command values for each phase of the three-phase inverter individually based on the waveform of the back EMF obtained offline. Furthermore, the proposed current command values are generated such that the copper loss in the system is minimized. The proposed method was verified experimentally. The torque vibration was suppressed by 91.3% compared with the conventional method at 1 kW load. Furthermore, the power factor was 99.9% under an identical load condition. These results verified the effectiveness of the proposed method in suppressing torque vibrations during PFC operation at all rotor positions.

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