Current Ripple Suppression Control Based on Prediction of Resonance Cancellation Voltage for Electrolytic-Capacitor-Less Inverter

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In an electrolytic-capacitor-less single-phase to three-phase inverter, the resonance between the line inductance on the source side and the DC-link capacitor causes an input current ripple. In this paper, a new approach for suppressing the input current ripple due to this resonance is described. This paper proposes a new control method that does not emulate the action of the damping resistor for passive damping. The proposed method cancels the resonance by using a DC-link capacitor discharge actively. In the proposed method, a cancellation voltage pulse that cancels the resonance is added to one of the three-phase voltage references for the inverter. The cancellation voltage is calculated by feeding back the variables relative to the discharge. The effectiveness of the proposed method is validated by experimental results.

Keywords: electrolytic capacitor less inverter, IPMSM, input current ripple, LC resonance, damping control

1. Introduction

Recently, variable-speed AC motor systems using an inverter have been widely used in many fields to reduce energy consumption. These applications range from home appliances to industrial facilities(1–6). In general, variable-speed AC motor systems use a single-phase to three-phase power converter. The single-phase to three-phase power converter consists of a rectifier, an electrolytic capacitor to compensate for the power ripple, and a three-phase inverter. In order to improve the input power factor, the rectifier generally includes a power factor correction (PFC) circuit. The PFC circuit makes the input current a sinusoidal waveform and the electrolytic capacitor generates a constant DC-link voltage(5)–(18). Figure 1 shows a boost chopper type single-phase to three-phase power converter with a general PFC circuit that consists of a boost chopper. The general PFC circuit requires a reactor, an electrolytic capacitor with large capacitance, and power switching devices. The reactor and the switching devices increase power loss. The reactor and the electrolytic capacitor are also heavy and occupy a large space. As a result, reducing the size of a power converter system is quite difficult. Moreover, the electrolytic capacitor also limits the lifetime of the system.

To solve this problem, the authors have proposed an electrolytic capacitor-less single-phase to three-phase power converter(19)–(24) for a residential compressor drive system. This system consists of a single-phase diode rectifier, a film capacitor with small capacitance at the DC-link, a three-phase inverter, and an interior permanent magnet synchronous motor (IPMSM). The DC-link capacitor is 14 μF and the power converter does not have any energy storage. Hence, the single-phase ripple power is provided to the IPMSM directly. The ripple power is smoothed by the moment of inertia of the IPMSM. The inverter controls both motor speed and the input power factor. In previous reports(19)–(20), high power-factor operation has been achieved under all load conditions. However, the input current ripple at the source side is caused by a resonance with the line inductance and the low capacitance at the DC-link.

Passive damping and damping control are the available methods to suppress the ripple due to resonance(25)–(32). Passive damping that connects a damping resistor is simple to implement but increases power loss. Unlike passive damping, damping control emulates the action of the damping resistor for passive damping and does not cause power loss. However, the damping control cannot be used for the electrolytic capacitor-less inverter because the resonance does not occur in the motor-side that inverter controls and the control system of the inverter does not have an input current control loop.

To suppress the input current ripple due to resonance, a motor control method for the drive without electrolytic capacitor has been researched(27)–(28). This method that does not emulate passive damping focuses on the DC-link voltage ripple due to LC resonance and cancels the resonance by utilizing DC-link
capacitor discharge. To discharge the DC-link capacitor, the method adds a cancellation voltage pulse that cancels the resonance to one of the three-phase voltage references. However, the method requires offline experimental tests to obtain the frequency, amplitude, and applying timing of the cancellation voltage pulse. These parameters change depending on operating condition such as motor speed, load torque, and line inductance. Moreover, when the fundamental frequency of the motor current is not a multiple of the source frequency, the method cannot reduce the input current ripple effectively.

To solve this problem, this paper proposes a new approach to suppress the input current ripple due to resonance. The proposed method calculates the cancellation voltage by feeding back variables relative to the discharge and expands the operating point that the current ripple suppression control can be applied. On the other hand, the ripple has high-frequency components because of the DC-link capacitor with small capacitance. Hence, the influence of controller dead time caused by digital control is large and the resonance is not canceled effectively. To lower the resonant frequency, it is necessary to increase the line inductance. In order to compensate the controller dead time and cancel the resonance without increasing the line inductance, the proposed method predicts the ripple by using memory that matches with the periodicity of the input current and the DC-link voltage.

First, the electrolytic capacitor-less inverter and the cause of the input current ripple in the system are explained. Next, a principle of the proposed method is described; and the periodicity of the input current and the DC-link voltage is determined to predict the ripple. Finally, the experimental results confirm the validity of the proposed method.

2. Electrolytic Capacitor-Less Inverter for an IPMSM Drive System

Figure 2 shows the electrolytic capacitor-less inverter. The system has a small energy storage element in the DC-link. The DC-link capacitor is used to absorb the DC-link current ripple caused by the switching of the inverter. As a result, the source-side power ripple is not smoothed by the DC-link capacitor, and the DC-link voltage severely fluctuates. Therefore, the system generates torque ripple and provides the IPMSM with source-side single-phase power ripple directly. The power ripple is smoothed by the moment of inertia of the IPMSM; and the motor speed in the system has some ripples and is controlled averagely. Hence, application of the system is limited to a residential compressor drive system that does not require speed response with high precision such as those used in air conditioners and refrigerators. The noise sound by the torque ripples is not a problem as a product.

In the system, the inverter controls not only the motor but also the input power factor because the system has a small energy storage element at the DC-link. Figure 3 shows the control block diagram of the inverter-output-power controller for achieving high power-factor operation. The inverter output power is controlled to synchronize with twice of the source voltage frequency to obtain a high power factor. The inverter output power reference \( p_{in}^{*} \) is obtained by subtracting the DC-link capacitor power \( p_c \) from the input power reference \( p_{in} \). The input power reference \( p_{in} \) is determined by multiplying the output of the motor speed controller and \( \sin^2 \omega_{m} t \), which is generated by the source voltage \( v_{in} \). The DC-link capacitor power \( p_c \) and the inverter output power \( p_{in} \) are calculated by (1) and (2), respectively.

\[
\begin{align*}
p_c &= V_{dc} i_c \\
&= \frac{1}{2} \omega_{in} C_{dc} V_{in}^2 \sin(2\omega_{in} t) \\
p_{in} &= v_{in} i_{dc} + v_{in} i_{dq} 
\end{align*}
\]

The inverter output power controller consists of the PI and the repetitive controller. The controller outputs the q-axis current reference \( i_{dq} \) and achieves high-power-factor operation. However, the controller does not consider the input current ripple due to LC resonance and cannot suppress the ripple because of the control bandwidth limitation.

3. Proposed Control Method for Suppressing Input Current Ripple

3.1 Input Current Ripple in Electrolytic Capacitor-Less Inverter

Figure 4 illustrates an example of the input current \( i_{in} \) and the DC-link voltage \( V_{dc} \) waveforms in the system. In the system, the ripple due to LC resonance occurs in the input current \( i_{in} \) and the DC-link voltage \( V_{dc} \) when the diode rectifier is turned on as shown in Fig. 4. The resonance is caused by the line inductance including the line filter inductance and the DC-link capacitor. The system uses a small film capacitor at the DC-link. Hence, the input current ripple attains high frequency and increases the input current harmonics. For example, if the line inductance is 0.2 mH and the capacitance of the DC-link capacitor is 14 µF, the resonant frequency is approximately 3 kHz. Therefore, it is difficult to apply dumping control to the current controller since the resonant frequency is usually higher than the bandwidth of the motor current controller.

3.2 Principle of the Proposed Method

Figure 5 shows the principle of the proposed control method. In order to cancel the resonance, the proposed method focuses on the DC-link voltage ripple due to LC resonance and suppresses the ripple on the DC-link voltage \( V_{dc} \) by controlling the discharge from the DC-link capacitor.

The DC-link voltage has the switching ripple due to the
charge and the discharge of the DC-link capacitor as shown in
Fig. 5. The discharge from the DC-link capacitor decreases
the DC-link voltage. The proposed method utilizes the DC-link
 capacitor discharge and cancels the resonance by increasing
and decreasing the discharge amount. The motor drive
system uses a voltage-source inverter, and it can control the
discharge every sampling period by changing the output volt-
age. To control the discharge, the proposed method adds a
cancellation voltage that cancels the resonance to one of the
three-phase voltage references. The cancellation voltage is
obtained by feeding back variables relative to the discharge
and calculating appropriate discharge time. In this study, the
cancellation voltage is added to the middle-voltage phase of
the three phases because there is a case in which the other
phase voltage references are saturated. The calculation pro-
cess of the cancellation voltage is described in Sect.3.4 in
detail.

On the other hand, because the ripple due to the resonance
has high-frequency components, the influence of the con-
troller dead time caused by digital control is large. If the
control system does not compensate the influence, it is not
possible to cancel the resonance effectively and it is neces-
sary to increase the line inductance for lowering the resonant
frequency. In this study, the influence is compensated by pre-
dicting the ripple. In order to predict the ripple, the proposed
method uses a memory that matches with the periodicity of the
input current and the DC-link voltage. The memory stores
the calculated cancellation voltage. By using the cancellation
voltage stored in the memory as the predictive value, the in-
fluence of the controller dead time is compensated. The per-
odicity of the input current and the DC-link voltage and de-
termination of the memory number are described in Sect.3.3.

3.3 Periodicity of Input Current and DC-link Voltage

In the system, the DC-link voltage decreases to near zero
because the power converter uses a film capacitor with small
capacitance at the DC-link. If the DC-link voltage is near
zero, the DC-link voltage is charged by the regeneration from
the motor. The amount of increase of the DC-link voltage due
to the charge varies depending on the magnitude of the motor
current at that moment. Then, the variation of the charging
time changes the magnitude and the generation timing of
the ripple caused by the resonance, because the resonance
occurs when the diode rectifier is turned on. Therefore, the
periodicity of the input current and the DC-link voltage is
changed by the motor current frequency.

First, the periodicity of the motor current is described. The
fundamental frequency of the motor current is expressed as

\[ f_{re} = \frac{P}{120} N_s \]  \hspace{1cm} (3)

where \( P \) is the pole number and \( N_s \) is the motor speed in rpm.

In the system, in order to provide the IPMSM with source-
side single-phase power ripple, the q-axis current has a fre-
quency component of twice of the source voltage frequency;
in order to conduct a flux-weakening-control method, the d-
axis current is held to a negative constant. The dq-axis cur-
cents are expressed as

\[
\begin{bmatrix}
    i_d \\
    i_q
\end{bmatrix}
= \begin{bmatrix}
    A_d \\
    A_q
\end{bmatrix} \sin^2(2 \pi f_{in} t) \]  \hspace{1cm} (4)

where \( A_d \) and \( A_q \) are the amplitude of the dq-axis current and
\( f_{in} \) is the source frequency. The three-phase current is ex-
pressed as

\[
\begin{bmatrix}
    i_a \\
    i_b \\
    i_c
\end{bmatrix}
= \begin{bmatrix}
    C & 0 & 0 \\
    0 & C & 0 \\
    0 & 0 & C
\end{bmatrix}
\begin{bmatrix}
    i_d \\
    i_q
\end{bmatrix} \]  \hspace{1cm} (6)

where

\[
\begin{bmatrix}
    \cos \theta_{re} & \sin \theta_{re} \\
    \cos(\theta_{re} - \frac{\pi}{2}) & -\sin(\theta_{re} - \frac{\pi}{2}) \\
    \cos(\theta_{re} + \frac{\pi}{2}) & -\sin(\theta_{re} + \frac{\pi}{2})
\end{bmatrix} \]  \hspace{1cm} (6)

\[
\omega_{in} = 2 \pi f_{in} \]  \hspace{1cm} (7)

\[
\theta_{re} = \omega_{re} t \]  \hspace{1cm} (8)

Eq. (5) indicates that the three-phase current has frequency
components of \( f_{re} \) and \( 2f_{in} \pm f_{re} \).

Next, the periodicity of the input current and the DC-link
voltage is described. In the system, the motor speed affects
the periodicity of the input current and the DC-link volt-
age. Therefore, the periodicity of the input current and DC-
link voltage is determined by the source-side frequency and

3 IEEJ Journal IA, Vol.6, No.1, 2017
motor-side frequency. If the frequency of the periodicity is defined as \( f_p \) and (9) to (11) are satisfied, the input current and DC-link voltage can have the periodicity of the frequency \( f_p \).

\[
\begin{align*}
  f_p &= \frac{f_m}{n} \quad (n = 1, 2, 3, \ldots) \quad \cdots \quad (9) \\
  f_{re} &= m f_p \quad (m = 1, 2, 3, \ldots) \quad \cdots \quad (10) \\
  f_s &= l f_p \quad (l = 1, 2, 3, \ldots) \quad \cdots \quad (11)
\end{align*}
\]

where \( n, m, \) and \( l \) are positive integers and \( f_s \) is the sampling frequency. By (9) and (10), frequency components \( f_{re} \) and \( f_s \) which the three-phase current has become integer multiples of \( f_p \), and the motor current is the same in each period of \( 1/f_p \). The outputs of the controller are also the same in each period of \( 1/f_p \) by satisfying (11). Hence, in steady state, the input current and the DC-link voltage have the periodicity of \( f_p \), and the motor current is the same in each period of \( 1/f_p \) by satisfying (11). In other words, the generation timing of the resonant ripple has the periodicity of \( f_p \). For prediction of the resonant ripple and compensation of the influence of the controller dead time, the memory number \( N \) is determined as

\[
N = \frac{f_s}{f_p} - 1. \quad \cdots \quad (12)
\]

If \( f_p \) is a repeating decimal, the fundamental frequency \( f_{re} \) of the motor current and the sampling frequency \( f_s \) cannot be set as a positive integer or a finite decimal to meet (10) and (11), and the memory number \( N \) also cannot be set as a positive integer. In order not to set a fraction at a repeating decimal, it is necessary that the prime factor of the denominator is only 2 and 5. Therefore, \( n = 2^a \cdot 5^b \) should be set as \( n = 2^a \cdot 5^b \) to compensate the influence of the controller dead time effectively, where \( a \) and \( b \) are non-negative integers. Then the prime factor of \( n \) includes only 2 and 5, \( f_p \) is a positive integer or a finite decimal, and \( f_{re} \) and \( f_s \) can be set as a positive integer or a finite decimal. The calculation process of the cancellation voltage with the memory is described in detail in Sect. 3.4.

### 3.4 Proposed Control Block Diagram

Figure 6 shows the main circuit and the control block diagram. In Fig. 6, \( l \) is the line inductance and \( r \) is the line resistance. The conventional control system in the black broken line in Fig. 6 consists of the motor speed controller, the inverter-output-power controller, and the motor current controller. The motor speed \( \omega_m \) is controlled by the PI controller. The output of the speed PI controller corresponds to the peak value of the input power reference of the system. Using the input power reference, the power control system controls the inverter output power to correct the input power factor as described in Sect. 2. When the motor speed increases, the DC-link voltage is increased because the back electromotive force increases. The increasing DC-link voltage reduces the input power factor and increases the input current harmonics. In order to solve this problem, the d-axis current is given by the flux-weakening-control method, and the d-axis current reference \( i_d^* \) is set at a negative constant value to apply flux weakening. In the system, the DC-link voltage fluctuates synchronously with the source voltage. Because of the fluctuating DC-link voltage, the outputs of the PI controllers for the dq-axis currents are saturated. The saturation causes input current harmonic distortions. In order to improve the input current waveform, a PI current control method is used that considers the voltage limitations of the inverter. The voltage references \( v_u^*, v_v^*, \) and \( v_w^* \) are the outputs of the d-q axis current controller. The cancellation voltage \( v_{c\text{con}}^* \) is calculated by the proposed method in the blue part in Fig. 6.
Current Ripple Suppression Control for Electrolytic-Capacitor-Less Inverter

By Kodai Abe et al.

Fig. 7. Process of proposed method

(a) (b)

(c) (d)

(e) (f)

(g) (h)

IPMSM is driven by the voltage references $v_{ref}^{u}$, $v_{ref}^{v}$, and $v_{ref}^{w}$ obtained by $v_{u}^{*}$, $v_{v}^{*}$, $v_{w}^{*}$, and $v_{can}^{*}$.

Figures 6 and 7 show the control block diagram and the process of the proposed method for calculating the cancellation voltage. As mentioned previously, the proposed method uses the discharge from the DC-link capacitor. The cancellation voltage is obtained by feeding back variables relative to the discharge and calculating the appropriate discharge time. Therefore, the cancellation voltage is calculated online. The process for calculating the cancellation voltage is described below. Each step corresponds to the steps in Figs. 6 and 7.

(a) First, the ripple component $v_{rip}^{*}$ of the DC-link voltage is extracted by subtracting the absolute value of the input voltage $|v_{in}|$ from the DC-link voltage $V_{dc}$.

(b) Near the zero-cross of $v_{in}$, the waveform unrelated to the ripple due to resonance remains by the charge of the DC-link capacitor. During the charge time, the input current $|i_{in}|$ does not flow. Hence, if $|i_{in}| = 0$, $v_{rip}^{*}$ is regarded as zero and converted to $v_{rip}^{*}$.

(c) Moreover, the frequency component such as the sensor’s offset unrelated to the resonance ripple is removed by a second-order high-pass filter. In this study, the bandwidth of the high-pass filter is set at 300 Hz.

(d) The relational equation between the current and voltage of a capacitor is expressed as (13). The discharge of the DC-link capacitor that is connected to the load is also based on (13)

$$v(t) = \frac{1}{C} \int_{t_0}^{t} i(t) dt$$

where $v(t)$, $C$, $i(t)$, and $t$ are regarded as the ripple component $v_{rip}^{*}$, the capacitance of the DC-link capacitor $C_{dc}$, the $x$-phase motor current, and the desired discharge time $t_{can}$, respectively. In this paper, $x$ stands for the phase to which the cancellation voltage is added. Because the discharge current in the DC-link is direct current, the $x$-phase motor current is converted to an absolute value. If the system is discrete and these variables are constant in a sampling period, the required discharge time $t_{can}$ for each sampling period is expressed as

$$v_{rip}^{*} = \frac{1}{C_{dc}} |i_x| t_{can}$$

$$t_{can} = \frac{v_{rip}^{*}}{|i_x|}$$

By (14), the cancellation voltage to obtain the required discharge time is expressed as

$$v_{can}^{*} = 2D_{can} = \frac{2t_{can}}{T_s}$$

where $D_{can}$ and $T_s$ represent the duty ratio and the sampling period, respectively. The conversion using duty ratio makes it possible to add the cancellation voltage to the normalized three-phase voltage references. Then, the unit of the cancellation voltage is p.u. on the basis of the DC-link voltage.

(e) The motor current is alternating current. The current path is changed by the polarity of the motor current as shown in Figs. 8 and 9. Therefore, in order to increase and decrease the discharge time, it is necessary to increase and decrease the on-time of the upper (or lower) arm if the phase current is positive (or negative). Then, the cancellation voltage is rewritten as (16) by taking into account the polarity of the phase current.

$$v_{can}^{*} = \text{sgn}(i_x) \cdot v_{can}^{*}$$

5 IEEJ Journal IA, Vol.6, No.1, 2017
(f) If the cancellation voltage calculated so far is applied, it is not possible to cancel the resonance effectively by the influence of the controller dead time caused by digital control. Because the input current ripple has high-frequency components, the influence is large. In this study, the influence is compensated by using the memory described in Sect. 3.3. The memory is designed to match with the periodicity of the input current and the DC-link voltage. The memory number \( N \) is determined by (12). The influence is compensated by using the cancellation voltage in the memory as the predictive value.

(g) The cancellation voltage is calculated on the basis of \( v_{rip} \) and consists of multiple pulses. However, the pulses following first pulse generate a resonant ripple newly. Therefore, the pulses following the first pulse are removed by counting the number of zero crossings. Then, the first pulse is determined as the cancellation voltage \( v_{can}^{c} \).

(h) Finally, \( v_{can}^{c} \) is added to the middle-voltage phase that is not saturated. Thereby, the proposed method can be applied in the case of a motor drive condition in which the other phase voltage references are saturated. The proposed method adds the cancellation voltage directly to the three-phase voltage references. Therefore, the proposed method is not limited by the bandwidth of the PI controller for the motor current and controls the discharge every sampling period. In this study, because the bandwidth of the high-pass filter in the process (c) of the proposed method is set at 300 Hz, the effectual frequency range by the proposed method is approximately 1500 Hz to \( f_{p} / 5 \) Hz. Therefore the line inductance including the line filter inductance should be designed so that resonant frequency is lower than one-fifth of the sampling frequency. In this study, the sampling frequency is set at 16 kHz and designed \( l \) including the line filter inductance is 0.2 mH as a typical value so that the influence of the controller dead time is large. The resonant frequency with \( l = 0.2 \text{ mH} \) and \( C_{dc} = 14 \text{ \mu F} \) is about 3 kHz, which is lower than one-fifth of the sampling frequency. The proposed method uses the DC-link capacitor discharge and prevents the ripple. Hence, the proposed method prevents a positive component of the ripple effectively. For a negative component of the ripple, the proposed method weakens the ripple by decreasing the discharge amount.

4. Experimental Results

The experimental setups are tested to verify the effectiveness of the proposed control method. Table 1 and Table 2 list the parameters of the experimental system. The line inductance \( l \) including the line filter inductance is set at 0.2 mH so that the influence of the controller dead time is large. The load condition is set at 1.5 Nm. The motor speed and the frequency \( f_{p} \) of the periodicity of the input current and the DC-link voltage are set at 4500 rpm and 50 Hz, respectively, for one experiment, and 3750 rpm and 25 Hz, respectively for the second experiment.

Figures 10, 11, 12, and 13 show the experimental results with the conventional method, which does not use the proposed method, and the proposed method at 4500 rpm and \( f_{p} = 50 \text{ Hz} \). Figure 14 shows the FFT analysis result of the input current at 4500 rpm and \( f_{p} = 50 \text{ Hz} \). In Fig. 10, the resonant ripple on the input current and the DC-link voltage is observed when the diode rectifier is turned on. Figures 10 and 11 show that the periodicity of the input current, the DC-link voltage, and the three-phase current is the same as 50 Hz set by \( f_{p} \). In Fig. 12, the cancellation voltage cancels the resonance on the DC-link voltage effectively and reduces the input current ripple. The input current distortion remains but the distortion is caused by the spatial harmonics of the IPMSM. Figure 13 shows that the proposed method does not affect the motor current control because the resonant frequency is enough higher than the fundamental frequency of the motor current. In the FFT analysis result shown in Fig. 14, the fundamental frequency of the input current is 50 Hz. The FFT analysis considers inter-order harmonics and groups them on the basis of IEC 61000-4-7. In Fig. 14, the harmonics obtained using the proposed control method are
suppressed more effectively than that obtained using the input current vibration reduction method \((23)\) that requires offline experimental tests and setting the fundamental frequency \(f_{\nu}\) of the motor current at a multiple of the source frequency. The proposed control method reduces the input current harmonics from thirtieth-order to sixty-fourth-order, the effective frequency range \((1500\, \text{Hz to } f_s/5\, \text{Hz})\) by the proposed method, by 40.8%.

Figures 15, 16, 17, and 18 show the experimental results of the conventional method and the proposed method at 3750 rpm and \(f_p = 25\, \text{Hz}\). Figure 19 shows the FFT analysis result of the input current at 3750 rpm and \(f_p = 25\, \text{Hz}\). In Fig. 15, the resonant ripple similar to the results at 4500 rpm occurs in the input current and the DC-link voltage. The
periodicity of the input current, the DC-link voltage, and the three-phase current is the same as 25 Hz set by $f_p$. In Fig. 17, the cancellation voltage cancels the resonance effectively. In the FFT analysis result as shown in Fig. 19, the proposed control method reduces the input current harmonics from the thirtieth-order to the sixty-fourth-order by 35.8%. The reason why there are large even-order harmonics in Fig. 19 compared with Fig. 14 is that the input current does not have vertical symmetry at this operating point.

Figures 20 and 21 show the experimental results of the conventional method and the proposed method at 2190 rpm and $f_p = 1$ Hz. Figure 22 shows the FFT analysis result of the input current at 2190 rpm and $f_p = 1$ Hz. The input current ripple is also reduced when the periodicity $f_p$ is low. In the FFT analysis result, the proposed control method reduces the input current harmonics from the thirtieth-order to the sixty-fourth-order by 6.4%.

Table 3 lists the input power factors of the experimental results. The power factor of the proposed method is higher than that of the conventional method. Hence, the proposed method suppresses the resonant ripple without affecting the fundamental frequency.

5. Conclusion

This paper proposes a new control method for suppressing the resonant ripple in an electrolytic capacitor-less inverter for IPMSM drive systems. The system consists of a single-phase diode rectifier, a film capacitor with small capacitance at the DC-link, a three-phase inverter, and an IPMSM. The DC-link capacitor is 14 $\mu$F. In the system, single-phase power ripple is smoothed by the moment of inertia of the IPMSM. A new inverter control method suppresses the input current ripple caused by the resonance with the line inductance and the low-capacitance DC-link capacitor. The proposed method does not emulate the action of the damping resistor for passive damping and cancels the resonance actively by using the DC-link capacitor discharge. In addition, the influence of the controller dead time caused by digital control is compensated by predicting the resonant ripple. In order to predict the resonant ripple, the proposed method uses a memory that matches with the periodicity of the input current and the DC-link voltage. In order to prevent the input current ripple, the proposed method adds a cancellation voltage to the three-phase voltage reference. The proposed method is not limited by the bandwidth of the PI controller for the motor current. The effectiveness of the proposed method is
verified through experiments in comparison with the conventional method.

References

To confirm the effectiveness at $f_{in} = 60$ Hz, experiment is carried out. The sampling frequency $f_s$ is changed 15 kHz so that the number of delay steps in the repetitive control of power controller is a positive integer\(^{(20)}\). app. Figs. 1, 2, 3, and 4 show the experimental results with the conventional method and the proposed method at 3600 rpm, 0.7 Nm and $f_p = 60$ Hz. app. Fig. 5 shows the FFT analysis result of the input current. These figures show that the proposed method cancels the resonance and the periodicity $f_p$ is the same as 60 Hz as well as the trend at $f_{in} = 50$ Hz. In app. Fig. 5, the proposed control method reduces the input current harmonics from twenty-fifth-order to fiftieth-order, the effective frequency range (1500 Hz to $f_s/5$ Hz) by the proposed method, by 35.5%.

app. Fig. 1. Experimental results of conventional control method at 3600 rpm and $f_p = 60$ Hz

app. Fig. 2. Three-phase current waveforms with conventional method at 3600 rpm and $f_p = 60$ Hz

app. Fig. 3. Experimental results of proposed method at 3600 rpm and $f_p = 60$ Hz

app. Fig. 4. Three-phase current waveforms with proposed method at 3600 rpm and $f_p = 60$ Hz

app. Fig. 5. FFT analysis result of input current at 3600 rpm and $f_p = 60$ Hz

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<tr>
<th>Conventional method</th>
<th>Proposed method</th>
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Current Ripple Suppression Control for Electrolytic-Capacitor-Less Inverter (Kodai Abe et al.)

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