Harmonic Current Reduction Control Based on Model Predictive Direct Current Control of IPMSM and Input Grid Circuit

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This paper proposes a new inverter control method to reduce the input current harmonics and switching frequency of electrolytic capacitorless inverters for interior permanent magnet synchronous motor (IPMSM) drives. In an electrolytic capacitorless inverter system, the input current oscillates due to a resonance between the grid inductance and DC-link capacitor. Due to the oscillation in the input current caused by the LC resonance, the regulations for input current harmonics are sometimes not satisfied. The proposed method, which adopts a model predictive direct current control (MPDCC) method, reduces the switching frequency and satisfies the input current harmonics regulations IEC61000-3-2. MPDCC predicts the motor and input currents and evaluates them using a cost function to select a voltage vector. The cost function reduces the number of switching times and results in a current response that follows the current reference. The prediction model for input current is constructed using the state equations of the input LC resonance circuit and IPMSM. The effectiveness of the proposed method is confirmed experimentally. Further, the experimental results confirm that the proposed method satisfies the input current harmonics regulation IEC61000-3-2.

Keywords: IPMSM, electrolytic capacitorless inverter, input current harmonics

1. Introduction

In recent years, inverter-driven AC motor systems are being used in many fields, ranging from home appliances to industrial applications in order to address global environmental problems (14–16). In general, single-phase to three-phase inverters are used for the variable-speed operation of motors; in these inverters, improving the grid input power factor is required. Therefore, a general single-phase to three-phase inverter includes a power factor correction (PFC) circuit with a diode rectifier (4). The purpose of the PFC circuit is to control the input current and hold the output voltage (DC-link voltage) constant; to achieve the latter objective, a high-capacitance electrolytic capacitor is used. However, such capacitors increase the size and shorten the life of inverter systems in compressor motors.

Such problems also occur in three-phase to three-phase inverter systems and hence methods to control capacitor current by predictive control have been proposed (15–16).

In a single-phase to three-phase inverter system, it has been proposed that the electrolytic capacitor in a motor drive system be replaced with a film capacitor (9). In such a system, because the DC-link part does not have a large energy buffer, power ripples are absorbed by the kinetic energy accumulated in the moment of inertia of the motor. However, as input current oscillates due to resonance between the grid inductance and DC-link capacitor, harmonics are generated in the input current. The amplitude of each input current harmonics is governed by IEC 61000-3-2 class A regulations (10), however, the input current harmonics in this system do not satisfy these regulations.

To improve these problems, instantaneous voltage control (IVC) has been proposed to reduce harmonics in input current by directly controlling the DC-link current (11–15). In the IVC method, the DC-link current control modifies the output voltage vector on the DC-link current line. In another approach, a method has been proposed to actively dampen vibrations occurring in the DC-link voltage due to LC resonance (16–18). In the active damping control of electrolytic capacitorless inverters, the output voltage reference is calculated by adding the voltage vector reference used for suppressing oscillations in the DC-link voltage to the voltage vector reference used for controlling the motor.

The voltage vector is outputted by pulse-width modulation (PWM) for suppression control over input current harmonics in electrolytic capacitorless inverters. Therefore, to reduce input current harmonics, it is necessary to set a high carrier frequency; at the same time, increasing the carrier frequency leads to an increase in the switching loss of the inverter. To reduce switching frequency, the authors propose a model predictive direct current control (MPDCC) method (19).
for an electrolytic capacitorless inverter. The method controls motor current and the DC-link current. However, the input current ripple due to the resonance between input filter $L$ and capacitor $C_{dc}$ can not be suppressed and the input current harmonics do not satisfy the regulations because the input current is not predicted.

This paper proposes a control method to reduce switching frequency and satisfy input current harmonics regulations using a MPDCC method. To control input current, in this paper, the state equation is derived for an electrolytic capacitorless system consisting of an IPMSM and a LC resonance circuit and a prediction model for input current is constructed. Furthermore, a suppression control protocol for input current harmonics and switching frequency is built by model predictive control according to the proposed prediction model. Finally, the effectiveness of the proposed method is confirmed experimentally.

2. Electrolytic Capacitorless Inverter for an IPMSM Drive System

2.1 System Configuration of an Electrolytic Capacitorless Inverter

Figure 1 shows the system configuration of an electrolytic capacitorless inverter. The main circuit consists of an input filter reactor, a single-phase diode rectifier, a low-capacity film capacitor, a three-phase inverter, and an IPMSM. The DC-link capacitor is changed from a high-capacity electrolytic capacitor to a low-capacity film capacitor to absorb current ripples due to inverter switching. Because the electrolytic capacitorless inverter includes a small film capacitor, single-phase power on the grid is directly supplied to the IPMSM. The power ripple on the grid is absorbed into the moment of inertia of the IPMSM. Therefore, the IPMSM experiences a torque ripple synchronized with the grid power ripple; the motor speed experiences some ripple and is controlled averagely. Since the inverter causes speed variations, applications are limited to compressor motor drives that do not require a high-precision speed response such as air-conditioners.

2.2 q-Axis Current Reference for High Input Power Factor Control

Figure 2 shows a method to calculate the q-axis current reference for a high input power factor control. The q-axis current reference is calculated by dividing torque reference by the torque constant; the torque reference uses the output of the motor speed controller. In an electrolytic capacitorless inverter system without an energy buffer, the electrical power supplied to the motor has a pulsating component with a frequency twice that of the grid frequency. Therefore, upon multiplying the output of the speed controller by the square of the sine wave of the grid frequency which is synchronous with the grid voltage, the motor torque is controlled to ripple synchronously with the grid. The q-axis current reference is calculated from the torque reference. Using the voltage equation (Eq. (1)) of an IPMSM in the d-q reference frame and considering loss as $p_{loss}$, the inverter output power $p_{inv}$ is represented as in Eq. (2).

Substitution yields electrical power from the grid. The torque reference is divided by the torque constant; the torque reference uses the output of the motor speed controller. In an electrolytic capacitorless inverter system without an energy buffer, the electrical power supplied to the motor has a pulsating component with a frequency twice that of the grid frequency. Therefore, upon multiplying the output of the speed controller by the square of the sine wave of the grid frequency which is synchronous with the grid voltage, the motor torque is controlled to ripple synchronously with the grid. The q-axis current reference is calculated from the torque reference. Using the voltage equation (Eq. (1)) of an IPMSM in the d-q reference frame and considering loss as $p_{loss}$, the inverter output power $p_{inv}$ is represented as in Eq. (2).

Here, $v_d$ and $v_q$ refer to the d-q-axis voltages, $i_d$ and $i_q$ are the d-q-axis currents, $R_s$ is the stator resistances, $L_d$ and $L_q$ are the d-q-axis inductances, $P$ represents pole pairs, $\omega_s$ is the electrical angular speed of the motor, $p_{loss}$ is the copper loss, $p_{mag}$ is the change in magnetic energy due to inductance, and $p_{loss}$ represents iron loss, mechanical loss, and stray load excluding copper loss. In an electrolytic capacitorless inverter, the relationship between input power $p_{in}$, capacitor power $p_c$, and inverter output power $p_{inv}$ is given by Eq. (3).
In addition, the DC-link voltage is increased as electromagnetic force increases with an increase in motor speed. As a result, the diode rectifier becomes non-conductive near the zero grid voltage (0 V), the input power factor decreases, and input current harmonics increase. To resolve this issue, the d-axis current is provided by the flux-weakening control method, while the d-axis current reference is constant. In this method, two-phase space vector modulation (SVM) is used to output the voltage vector, and the switching frequency is about two thirds of the control period.

Figure 3 shows experimental IPMSM control results at a high input power factor under the conditions shown in Tables 1 and 2. In the experimental results shown in Fig. 3, the input power factor is 93.5%. In the conventional method, input current harmonics are dependent on the LC resonance frequency. The 9th order of the FFT results does not satisfy harmonics regulations and is close to the LC resonance frequency.

3. MPDCC for an Electrolytic Capacitorless Inverter

3.1 Cost Function of MPDCC

This section explains the cost function of MPDCC, which is crucial in selecting the output voltage vector. Figure 5 shows a schematic diagram of the voltage-vector selection method. The voltage vector is the output of an inverter with a one sample delay. $\psi$ in Fig. 5 is calculated at time point $[t-1]$ and is the output of the inverter at time point $[t]$. Each current at time point $[t]$ is calculated using $\psi$ calculated at time point $[t-1]$ from the prediction equation. Table 3 shows the number of samples for the voltage vector in the frame in Fig. 5. The three-phase inverter includes eight switching patterns and outputs seven voltage vectors corresponding to each pattern. In MPDCC, the output voltage vector corresponding to the proposed modulator is selected by predicting a current when each voltage vector is outputted. In the proposed method, the cost function is calculated using a reference frame set for the current reference and it is determined whether or not the predicted current for each voltage vector is within the reference frame. Those voltage vectors deviating from the reference frame are excluded. When the predicted current is within the current reference frame, the trajectory of the current is extended by linear approximation just before exceeding the current reference frame. The number of samples $N_n$ up to the extended point is counted, and the cost function $C_n$ of each voltage vector is calculated using Eq. (9).

$$C_n = \frac{S_n(t)}{N_n}$$

Here, $S_n(t)$ is the number of switching times of the inverter necessary to switch the voltage vector at point $[t]$. Further, the number of samples is calculated for each evaluation cost function. To control the actual current response within the range of the reference frame, the number of samples is defined in terms of the current value the longest number $N_n$ beyond the upper or lower limit of the reference frame. In this paper, the upper or lower limit of the reference frame are set by cut and try to satisfy the harmonics guideline. Using
In this section, a method has been proposed to control input current by evaluating the input current and motor current. Therefore, a capacitorless inverter, the three-phase inverter controls both input current harmonics and the output voltage vector. Equation (10) shows the prediction equation of the DC-link current.

\[ I_{dc}[t + 1] = \frac{i_d[t]v_{dn} + i_q[t]v_{qn}}{2} \]

Here, \( I_{dc}[t + 1] \) is the predicted value of the DC-link current, \( i_q[t + 1] \) is the predicted value of the dq-axis current, and \( v_{dn} \) is the normalized dq-axis voltage reference. The switching frequency is reduced by MPDCC, while satisfying the input current harmonics regulations. However, it is difficult to consider the influence of LC resonance current on input current harmonics, as it does not flow in the DC-link current. From the experimental results shown in the appendix, it has been confirmed that the regulations for input current harmonics are not satisfied, depending on load conditions. Therefore, to control input current accurately, it is necessary to evaluate the input current instead of DC-link current.

### 3.2 Prediction Model for Input Current

This section describes a prediction model for the input current. An electrolytic capacitorless inverter is divided into two main parts: a capacitor input rectifier circuit and a three-phase inverter. Assuming a three-phase inverter and IPMSM as the current source and assuming that the diode of the rectifier is always conducting, the electrolytic capacitorless system is represented shown in Fig. 6. The differential equation that derives input current and DC-link voltage is as follows:

\[
C_{dc} \frac{dV_{dc}}{dt} = i_c = i_{in} - I_{dc} \nonumber
\]

\[
\frac{dV_{dc}}{dt} = \frac{1}{C_{dc}}(i_{in} - I_{dc}) \nonumber
\]

\[= \frac{1}{C_{dc}}(i_{in} - i_dv_{dn} + i_qv_{qn}) \tag{11} \]

\[L_f \frac{di_{in}}{dt} = |v_{i}| - V_{dc} \nonumber \]

\[L_f \frac{di_{in}}{dt} = 1 \frac{|v_{i}| - V_{dc}}{L_f} \tag{12} \]

From the voltage equation of the IPMSM, the state equation of the current on the motor side, which is considered to be a current source, is as follows:

\[
\frac{d}{dt} \begin{bmatrix} i_d \\ i_q \end{bmatrix} = \begin{bmatrix} \frac{-R_i}{2L_d} & \omega_r \frac{L_q}{2L_d} \\ -\omega_r \frac{L_q}{2L_d} & \frac{R_i}{2L_d} \end{bmatrix} \begin{bmatrix} i_d \\ i_q \end{bmatrix} + \begin{bmatrix} \frac{V_{dc}}{2L_f} \\ 0 \end{bmatrix} \begin{bmatrix} v_{dn} \\ v_{qn} \end{bmatrix} \tag{13} \]

The state equation of the electrolytic capacitorless inverter is as shown in Eq. (14); it is derived from Eqs. (11)–(13).
Each element of matrix A of the electrolytic capacitorless system discrete equation is calculated from Eq. (17) as follows:

\[
A = e^{At_{s}} = L^{-1} \begin{bmatrix}
1 & C_{11} \sin C_{2}T_{s} + \cos C_{2}T_{s} \\
C_{12}e^{-C_{2}T_{s}} \sin C_{2}T_{s} \\
C_{21} \cos C_{2}T_{s} \\
A_{33} \cos C_{2}T_{s} \\
A_{34} \sin C_{2}T_{s} \\
A_{33} \cos C_{2}T_{s} \\
A_{43} \sin C_{2}T_{s} \\
A_{44} \cos C_{2}T_{s} \\
A_{13} = A_{14} = A_{23} = A_{24} = 0 \\
A_{31} = A_{32} = A_{41} = A_{42} = 0
\end{bmatrix}
\]

Similarly, each element of matrix b is calculated from Eq. (18) as follows:

\[
b_{11} = \frac{V_{dc}}{2L_{d}C_{1}^{2} + C_{2}^{2}} \begin{bmatrix}
(1 + C_{11}C_{2})(1 - e^{-C_{2}T_{s}} \cos C_{2}T_{s}) \\
C_{12}e^{-C_{2}T_{s}} \sin C_{2}T_{s} \\
C_{21} \cos C_{2}T_{s} \\
C_{12}e^{-C_{2}T_{s}} \sin C_{2}T_{s} \\
(1 - e^{-C_{2}T_{s}} \cos C_{2}T_{s}) \\
C_{12}e^{-C_{2}T_{s}} \sin C_{2}T_{s} \\
C_{21} \cos C_{2}T_{s} \\
C_{21} \cos C_{2}T_{s} \\
A_{13} = A_{14} = A_{23} = A_{24} = 0 \\
A_{31} = A_{32} = A_{41} = A_{42} = 0
\end{bmatrix}
\]

The accurate performance of the proposed prediction model is confirmed experimentally. Figure 7 shows the experimental results corresponding to the proposed model. The operating condition is the same as Fig. 3. Experimental results confirmed that the input current and DC-link voltage are accurately predicted by the proposed model under the rated conditions with large inductance fluctuation.

### 3.3 Input Current Reference

Figure 8 shows the system used to calculate the input current reference \(i_{\text{in}}^*\). It is obtained by converting the input power reference \(P_{\text{in}}^*\) into the power dimension and dividing \(P_{\text{in}}^*\) (in Fig. 8) by the power grid voltage. The absolute value of the input current reference \(i_{\text{in}}^*\) in an electrolytic capacitorless inverter is calculated.
The equation for calculating the input current reference $|i_{in}^*|$ is as follows:

$$|i_{in}^*| = \frac{\omega_{re} \tau_{in}^*}{P v_s}$$  \hspace{1cm} (52)

Here, $|i_{in}^*|$ is the input current reference and $v_s$ is the grid voltage.

4. Experimental Results

Figure 9 shows the system configuration of the proposed method, while Fig. 10 shows the experimental equipment. Tables 1 and 2 show the motor parameters and inverter circuit constants used for conducting the experiments. A power analyzer (PW6001, HIOKI E.E. CORPORATION Co., Ltd.) is used for measuring the input power factor and inverter loss.

The experimental results of the proposed method at a control period of 32 kHz, operation-rated speed of 3000 rpm, and 50% or 100% load are shown in Figs. 11 to 14. The proposed method reduced the average switching frequency of the inverter to 3.2 kHz. The average switching frequency is calculated by dividing the switching times during twice the source period by twice the source period. To compare the average switching frequency with a conventional triangular carrier comparison based pulse width modulation, the switching times is averaged between three phase and the switching is one cycle with ON and OFF. As shown in Figs. 11 and 13, the input current harmonics are reduced while and the LC resonance current is suppressed using the proposed method.

Figures 12 and 14 show the FFT results of the input current. By applying the proposed method, the harmonics amplitude of the input current are reduced, and the number of harmonics is below the regulation value. Table 4 shows the inverter losses and input power factors of the proposed and conventional methods. In the proposed method, the switching frequency decreases as the inverter loss decreases because the proposed method selects a voltage vector that minimizes the switching times by the cost function of Eq. (9) in Section 3.1. These results demonstrate the effect of reducing inverter loss using the proposed method. In addition, it is confirmed that the numerical simulation results using Psim is similar to the experimental results.
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5. Conclusion

This paper proposes a new prediction model for input current and proposes direct current control by input current prediction. In the experiment, the proposed input current model accurately predicts input current. In the IPMSM driving experiment using the proposed method, the harmonics of input current are found to be reduced in accordance with IEC 61000-3-2 regulations. In addition, the average switching frequency is reduced by the proposed method along with the inverter loss. These results demonstrate the high effectiveness of the proposed method.

Table 4. Measurement results of inverter loss and input power factor

<table>
<thead>
<tr>
<th></th>
<th>Conventional method</th>
<th>Proposed method</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>50% load</td>
<td>100% load</td>
</tr>
<tr>
<td>Switching frequency</td>
<td>10.6 kHz</td>
<td>3.2 kHz</td>
</tr>
<tr>
<td>Inverter loss</td>
<td>40 W</td>
<td>28 W</td>
</tr>
<tr>
<td>Improvement rate</td>
<td>84.5 %</td>
<td>30.9 %</td>
</tr>
<tr>
<td>Input power factor</td>
<td>96.6 %</td>
<td>96.1 %</td>
</tr>
</tbody>
</table>

References


The equation obtained by discretizing based on the zero-order hold of the sampling time Ts is given below.

\[ x[i+1] = Ax[i] + bu[i] + bd \]  

(A2)

The coefficient matrix A is as follows:

\[ A = \begin{bmatrix} A_{11} & A_{12} \\ A_{21} & A_{22} \end{bmatrix} = e^{A \cdot T_s} = (I - A) \]  

(A3)

Each element of matrix A of the IPMSM discrete equation is calculated from Eq. (A3) as follows:

\[ A_{11} = e^{-c_{11}T_s} \sin C_2T_s + \cos C_2T_s \]  

(A4)

\[ A_{12} = C_{12} e^{-c_{12}T_s} \sin C_2T_s \]  

(A5)

\[ A_{21} = C_{21} e^{-c_{21}T_s} \sin C_2T_s \]  

(A6)

\[ A_{22} = e^{-c_{22}T_s} \sin C_2T_s + \cos C_2T_s \]  

(A7)

Similarly, each element of matrix b is calculated from Eq. (A8) as follows:

\[ b = [b_{11} b_{12}] = \int_0^{T_s} e^{A \cdot T_s} \cdot dT \cdot b_e \]  

(A8)

\[ b_{11} = \frac{1}{L_c} \left( C_{11} C_2 (1 - e^{-c_{11}T_s}) \cos C_2T_s \right) \]  

\[ - C_{11} C_2 e^{-c_{11}T_s} \sin C_2T_s + C_1 (1 - e^{-c_{11}T_s}) \cos C_2T_s \]  

\[ + C_2 e^{-c_{21}T_s} \sin C_2T_s \]  

(A9)

\[ b_{12} = \frac{1}{L_q} \left( C_{21} (1 - e^{-c_{21}T_s}) \cos C_2T_s \right) \]  

\[ - C_{21} e^{-c_{21}T_s} \sin C_2T_s \]  

(A10)

\[ b_{21} = \frac{1}{L_q} \left( C_{21} \left( 1 - e^{-c_{21}T_s} \cos C_2T_s \right) \right) \]  

\[ - C_{21} e^{-c_{21}T_s} \sin C_2T_s \]  

(A11)

\[ b_{22} = \frac{1}{L_q} \left( C_{22} \left( 1 - e^{-c_{22}T_s} \cos C_2T_s \right) \right) \]  

\[ - C_{22} C_2 e^{-c_{22}T_s} \sin C_2T_s + C_1 (1 - e^{-c_{22}T_s} \cos C_2T_s) \]  

\[ + C_2 e^{-c_{21}T_s} \sin C_2T_s \]  

(A12)

The variables used in these equations are given below.

\[ C_1 = \frac{1}{2} \left( \frac{R_d}{L_d} + \frac{R_q}{L_q} \right) \]  

(A13)

\[ C_2 = \frac{1}{2} \sqrt{R_d^2 \left( \frac{R_q^2}{L_d L_q} + \omega^2 \right) - \left( \frac{R_d}{L_d} + \frac{R_q}{L_q} \right)^2} \]  

(A14)

\[ C_{11} = \frac{\omega}{C_2} \]  

(A15)

\[ C_{12} = \frac{\omega_{ref} L_d}{C_2 L_d} \]  

(A16)

\[ C_{21} = \frac{\omega_{ref} L_d}{C_2 L_q} \]  

(A17)

\[ C_{22} = \frac{\omega}{C_2} \]  

(A18)

The prediction equation of the dq axis current is shown in Appendix.
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(a) Input Current  
(b) DC-Link Current  
(c) d-axis Current  
(d) q-axis Current

Fig. 1. Experimental results of MPDCC about DC-link current prediction on condition of rated speed 3000 rpm, 50% load and 32 kHz control period

Fig. 2. FFT analysis of input current by MPDCC about DC-link current prediction on condition of rated speed 3000 rpm, 50% load and 32 kHz control period

Eq. (A19).

\[
\begin{bmatrix}
id(t + 1) \\
iq(t + 1)
\end{bmatrix} =
\begin{bmatrix}
A_{11}i_d(t) + A_{12}i_q(t) \\
A_{21}i_d(t) + A_{22}i_q(t)
\end{bmatrix} + 
\begin{bmatrix}
b_{11}v_d(t) + b_{12}v_q(t) \\
b_{21}v_d(t) + b_{22}v_q(t)
\end{bmatrix} + 
\begin{bmatrix}
-b_{12}\omega_r\phi_a \\
b_{22}\omega_r\phi_a
\end{bmatrix}
\]

(A19)

The prediction equation of the DC-link current using the predicted value of the dq-axis current value is as follows:

\[
I_{dc}[t + 1] = \frac{i_d[t + 1]v_{dn} + i_q[t + 1]v_{qn}}{2}
\]

(A20)

In the MPDCC method, in terms of DC-link current prediction, control is achieved by predicting future values using the dq-axis current and DC-link current prediction equations. The experimental results of MPDCC on DC-link current prediction, at a control period of 32 kHz, operation speed of 3000 rpm, and 50% or 100% load are shown in Fig. 1.

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Fig. 3. Experimental results of MPDCC about DC-link current prediction on condition of rated speed 3000 rpm, 100% load and 32 kHz control period

Fig. 4. FFT analysis of input current by MPDCC about DC-link current prediction on condition of rated speed 3000 rpm, 100% load and 32 kHz control period

to Fig. 4. With response to DC-link current prediction, MPDCC reduced the average switching frequency of an inverter to 3.8 kHz. Experimental results at 50% load satisfy input current harmonics regulations. However, at 100% load, the regulations are not satisfied. The reason for this observation is that the LC resonance current is not considered, as it does not flow in the DC-link current.
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