Fast Torque Response and Reduced Pulse Width Modulation Switching Frequency Based on Model Predictive Direct Torque Control and Selective Harmonic Elimination

Tenjiro Hiwatari\textsuperscript{a)\textcopyright} Student Member, Kiyoshi Ohishi\textsuperscript{f} Fellow
Yuki Yokokura Member, Haruya Kada\textsuperscript{f} Student Member
Sota Sano\textsuperscript{f} Member

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Optimized pulse patterns, such as the selective harmonic elimination pulse width modulation (PWM) pattern (SHE-PWM) are used to suppress current harmonics at steady-state. However, achieving fast dynamic performance is very difficult in systems operated by SHE-PWM. This paper proposes a method that combines the performance of SHE-PWM at a steady-state with the fast torque response of model predictive direct torque control at the transient-state. Finally, the control performance of the proposed system is evaluated based on the experimental results.

Keywords: induction motor, model predictive control, selective harmonic elimination, direct torque control, disturbance observer

1. Introduction

In various control systems, a response that tracks to the reference value without error at steady-state, and the error is required to reach zero as quickly as possible from a transient-state when there is a reference input or disturbance. In an AC motor, it is ideal for the current harmonics at the given switching frequency to be minimized at the steady state, and the fast torque response is realized at the transient state. However, the current harmonics are proportional to the harmonic losses in the stator winding resistance, whereas the switching frequency relates to the switching losses of the inverter.

Therefore, when attempting to reduce the switching loss of an inverter, the current harmonics and the harmonic loss increase. This issue can be addressed using optimized pulse patterns (such as selective harmonic elimination pulse width modulation (PWM) pattern (SHE-PWM)). This technique provides a desired output voltage pulse of a pulse width modulation (PWM) converter to eliminate low-order harmonics by performing an offline Fourier series expansion over one fundamental period of the PWM voltage. Thus, the transition instants (or switching angles $\alpha_i$) of each switch are optimally obtained to eliminate undesired harmonics while regulating the fundamental component at the required value. As a result, the current harmonics are minimized at a given switching frequency. However, in this method, the current harmonic reduction performance deteriorates, and the response is very slow at the transient state.

Since its introduction in 1985 by Takahashi and Noguchi, direct torque control (DTC) has quickly matured to an industrial standard for drive control. The basic characteristic of DTC is that the inverter’s switch positions are directly set; thus, the modulation techniques have no PWM or space vector modulation (SVM). The main advantages of DTC are the fast torque performance in the transient state, simple implementation, and inherent robustness. Significant work has been performed in the past to improve the look-up table (LUT) with the goal of reducing both the torque ripple and the switching frequency of the inverter.

In recent years, model predictive control (MPC) has received considerable attention in the power electronics and drives community. To improve the performance of DTC by reducing the switching frequency while maintaining a fast torque response, MPC for power electronics have been studied. A disadvantage of using direct MPC is that solving the underlying optimization problem, thereby, deriving the discrete manipulated variable, proves to be computationally challenging. Computational issues become especially important for long prediction horizons because the number of possible switching sequences grows exponentially as the prediction length is increased. As a result, when reference tracking of the stator current or the motor torque is considered, the prediction horizon is usually set to one. The most accurate method of extending the prediction value using the hysteresis bounds is to use the internal model of the drive in an open-loop simulation. This method requires excessive computing power and is therefore impractical. Linear and quadratic extrapolation are proposed to extend the prediction value. The computational burden involved can be handled with the currently available controller hardware. The linear extrapolation method was successfully
implemented on the existing DTC hardware.

Without the hysteresis bounds, the number of possible switching sequences increases exponentially as the length of the prediction interval increases. A solution to this problem is proposed that adopts the notion of sphere decoding and tailors it to the problem at hand\(^{(19)-(21)}\). As a result, the un

is proposed that adopts the notion of sphere decoding and switching sequences increases exponentially as the length of harmonics in real time\(^{(22)}\). This proposed method utilizes a PWM technique has been suggested to eliminate lower-order harmonics at steady state using MPC is not the same as the optimum performance due to the reduction in the current harmonics at steady state using MPC becomes longer owing to the increase in calculation time. As addition, it is necessary to consider at least one cycle of the hysteresis bounds or SHE even when the prediction interval is expanded, the hysteresis bounds are optimized, and the cost function is improved.

To realize steady-state current harmonics with the same performance as those of SHE-PWM, the MPC-based SHE-PWM technique has been suggested to eliminate lower-order harmonics in real time\(^{(22)}\). This proposed method utilizes a sliding discrete Fourier transform to obtain the amplitudes of harmonic components in real time and then uses a predictive sliding discrete Fourier transform to obtain the amplitudes of harmonic components in real time. This method has been implemented in real-time calculations for SHE-PWM and has been confirmed to be effective for a single motor drive.

Furthermore, control systems with power converters using SHE-PWM and MPC have been proposed\(^{(23)-(26)}\). With the conventional proposed control system and cost function, the predictive controller prefers to track the stator current reference at transient state while preserving SHE-PWM voltage pattern at steady state. However, the control performance at the transient state deteriorates because the speed controller has an integrator when combining the two control systems and the controlled variable is the stator current.

This paper describes a control system based on SHE-PWM and model predictive direct torque control (MPDTC) for an induction motor (IM). The proposed system realizes fast torque response at the transient state and a fast rejection of torque disturbances. In addition, a three-phase current spectrum nearly equal to that of SHE-PWM is achieved at steady state. The effectiveness of the proposed system is evaluated based on the experimental results.

2. Characteristic of MPDTC and SHE-PWM

This section explains MPDTC and SHE-PWM. In addition, this section explains the reason for combining SHE-PWM and MPC, and the difference between the conventional control scheme and the proposed control scheme.

2.1 Model Predictive Direct Torque Control

This paper explains MPDTC with the controlled object as the IM. When the stator flux linkage \(\psi_s\) and the rotor flux linkage \(\psi_r\) on the stationary reference frame \((abf\) reference frame) are chosen as state variables, the state-space representation of the IM becomes

\[
\begin{align*}
\begin{bmatrix} \mathbf{v}_{abf} \\ \mathbf{v}_{raf} \end{bmatrix} &= \begin{bmatrix} \mathbf{A}_{11abf} & \mathbf{A}_{12abf} \\ \mathbf{A}_{21abf} & \mathbf{A}_{22abf} \end{bmatrix} \begin{bmatrix} \mathbf{\psi}_{abf} \\ \mathbf{\psi}_{raf} \end{bmatrix} + \begin{bmatrix} \mathbf{B}_{1abf} \\ \mathbf{B}_{2abf} \end{bmatrix} \mathbf{u}_{abf} \\
\mathbf{v}_{raf} &= \begin{bmatrix} \mathbf{v}_{sa} \\ \mathbf{v}_{ja} \end{bmatrix}^T
\end{align*}
\]

where \(\mathbf{v}_{abf}\) is the stator flux linkage on the \(abf\) coordinate \([\text{Wb}]\), \(\mathbf{v}_{raf}\) is the rotor flux linkage on the \(abf\) coordinate \([\text{Wb}]\), \(\mathbf{v}_{raf}\) is the stator voltage on the \(abf\) coordinate \([\text{V}]\), \(R_s\) and \(R_r\) are the stator and rotor resistances, respectively \([\Omega]\), \(L_s\) and \(L_r\) are the stator and rotor inductances, respectively \([\text{H}]\), \(M\) is the magnetizing inductance \([\text{H}]\), \(\omega_r\) is the electric angular frequency \([\text{rad/s}]\), and \(p\) is the derivative operator.

The torque equation of the IM is expressed by (2).

\[
\tau_m = P \frac{M}{L_s L_r - M^2} \left( \mathbf{\psi}_{raf}^T \mathbf{\psi}_{raf} - \mathbf{\psi}_{sa}^T \mathbf{\psi}_{ja} - \mathbf{\psi}_{ja}^T \mathbf{\psi}_{raf} \right)
\]

where \(P\) is the number of pole pairs.

This paper explains the prediction calculation formula of the motor torque and the stator flux linkage for MPDTC. It is assumed that the electric angular frequency \(\omega_r\) is a constant within the prediction interval. Equations (3) and (4) are obtained by discretizing equation (1) using a zero-order hold sampling time \(T_s\).

\[
\begin{align*}
\mathbf{\psi}_{raf}(k + 1) &= e^{A_{21abf}T_s} \mathbf{\psi}_{raf}(k) \\
&+ \int_0^{T_s} e^{A_{11abf}T \tau} A_{12abf} \mathbf{\psi}_{raf}(k) d\tau \\
&+ \int_0^{T_s} e^{A_{11abf}T \tau} \mathbf{B} \mathbf{u}_{abf}(k) d\tau \quad \cdots \cdots (3) \\
\mathbf{\psi}_{raf}(k + 1) &= e^{A_{21abf}T_s} \mathbf{\psi}_{raf}(k) \\
&+ \int_0^{T_s} e^{A_{11abf}T \tau} A_{22abf} \mathbf{\psi}_{raf}(k) d\tau \quad \cdots \cdots (4)
\end{align*}
\]

The predicted values of motor torque and stator flux linkage are expressed as in (5) and (6) from (3) and (4), respectively.

\[
\begin{align*}
\mathbf{\psi}_s(k + 1) &= \sqrt{\mathbf{\psi}_{sa}^T(k + 1) + \mathbf{\psi}_{ja}^T(k + 1)} \quad \cdots \cdots (5) \\
\tau_m(k + 1) &= P \frac{M}{L_s L_r - M^2} \left( \mathbf{\psi}_{raf}(k + 1) \mathbf{\psi}_{raf}(k + 1) \\
&- \mathbf{\psi}_{sa}^T(k + 1) \mathbf{\psi}_{ja}(k + 1) \right) \quad \cdots \cdots (6)
\end{align*}
\]

A standard MPC operates in discrete time with sampling frequency \(f_s\) and one-step prediction horizon. Thus, at each sampling time \(k\), a measurement of the system state is taken and a cost function is evaluated for each tentative control input element. Generally, the cost function only considers a
positive sum of the tracking errors for each controlled variable. The standard horizon-one cost function using MPDTC is expressed as

$$C_n = \left| \tau^\text{ref}_m (k) - \tau_m (k + 1) \right| + \lambda \left| \psi^\text{ref} (k) - \psi_m (k + 1) \right|$$

where $\lambda$ is a weighting factor. The optimal control input that minimizes the cost function (7) is applied to the inverter. The tested IM is shown in Fig. 1. Figure 2 is a block diagram using MPDTC. Figure 3 shows the torque response at transient state for the torque step reference (experimental results), and Fig. 4 shows the control performance at transient state when the IM has the 50% load torque (experimental results). The parameters of the tested IM and inverter are summarized in Table 1. As shown in Fig. 3, the torque generated by the IM $\tau^\text{cal}_m$ tracks the torque step reference $\tau^\text{ref}_m$ within 2 ms, and MPDTC realizes a fast torque response at the transient state.

2.2 Selective Harmonic Elimination SHE-PWM

parameters are calculated in an offline procedure by computing the optimal switching angles using a Fourier series expansion over one fundamental period of the PWM voltage waveforms for all possible operating points. Thus, the transition instants (or switching angles $\alpha_i$) of each switches are optimally obtained to eliminate undesired harmonics while regulating the fundamental component to the required value. The Fourier coefficients of the $N_h$-th harmonic for a two-level voltage waveform is given in general form by

$$0 = \frac{4}{N_h \pi} \left[ 1 + 2 \sum_{i=1}^{N_s} (-1)^i \cos N_h (\alpha_i) \right]$$

where $N_h$ is the order of the harmonic and $N_s$ is the number of switching angles $\alpha_i$ in the PWM waveform. Further, the modulation rate $m$ of the inverter is expressed as (9) using the switching angles $\alpha_i$.

$$m = \frac{4}{\pi} \left[ 1 + 2 \sum_{i=1}^{N_s} (-1)^i \cos (\alpha_i) \right]$$

All integer harmonics of even order and all triple-harmonics of the PWM voltage waveform are zero because quarter-wave symmetry is generally assumed when calculating SHE-PWM (Fig. 5). Figure 6 shows the block diagram using V/f control and SHE-PWM, where the modulation rate $m$ of the inverter is calculated as (10) (variable-voltage region).
Motor Drive based on MPDTC and SHE-PWM

Fig. 5. Voltage pulse patterns based on selective harmonic elimination pulse width modulation pattern (SHE-PWM)

Fig. 6. Block diagram of SHE-PWM and V/f control

\[ m = \sqrt{\frac{8 \cdot K_{Vf} \cdot \omega_{ref}}{V_{DC}}} \] .............................. (10)

where \( K_{Vf} \) is the V/f ratio, \( V_{DC} \) is the DC link voltage of the inverter [V], and \( \omega_{ref} \) is the primary angular frequency [rad/s].

Figure 7 shows the control performance at transient state when the IM has the 50% load torque (experimental results). \( i_{ref,MT} \) is a current on MT reference frame. MT reference frame is the rotational coordinate synchronized with the stator flux linkage. As shown in Fig. 7, when the IM has the 50% load torque, it is confirmed that the IM generates the motor torque corresponding to the 50% load torque, and the primary angular frequency is controlled to be constant. In addition, the average switching frequency of the inverter at steady-state is low in SHE-PWM. The motor torque cannot be controlled because SHE-PWM systems are controlled by the switching angles calculated under the steady-state condition. Therefore, SHE-PWM is applied only to very slow control loops.

This paper proposes a method of combining two control methods, MPDTC and SHE-PWM, by the cost function. The proposed method realizes a torque response speed that is nearly equal to that of MPDTC at the transient state and a reduction in the switching frequency of the inverter that is nearly equal to that of SHE-PWM at steady state.

2.3 Combination of MPC and SHE-PWM

This paper explains the control method that has the advantages of both MPC and SHE-PWM. In SHE-PWM, the performance at the transient state deteriorates, and in MPC, it is difficult to reduce the switching frequency of the inverter to nearly equal to that of SHE-PWM at steady state. Therefore, a method using a LUT for the cost function of MPC has been proposed. The conventional proposed method recalculates SHE-PWM from the phase [rad] of one sample \( \Delta \delta \). This method has high reproducibility of SHE-PWM at the set sampling frequency \( f_s \) and fundamental frequency \( f_0 \).

\[ \Delta \delta = \frac{2\pi}{f_0 f_s} \] .............................. (11)

The LUT of a general SHE-PWM reads each phase-switching signal corresponding to the modulation rate \( m \) of the inverter and phase of SHE-PWM from a table. However, the method recalculates SHE-PWM from the phase [rad] of one sample \( \Delta \delta \). This method has high reproducibility of SHE-PWM at the set sampling frequency \( f_s \) and fundamental frequency \( f_0 \).

This paper proposes a method of discretization without increasing the LUT of SHE-PWM. The proposed method calculates the phase [rad] of one sample \( \Delta \delta \) at the worst-case frequency (rated frequency \( f_0 \)) and calculates SHE-PWM up to the value one digit higher than the phase of \( \Delta \delta \). The reproducibility of SHE-PWM decreases at low frequencies.
because this method uses SHE-PWM taking the worst case into account at all fundamental frequencies.

The proposed method has different switching timing for the switch states of each phase. In this paper, the time at which the phase shifts by $120^\circ$ is set to an integral multiple of the control period

$$T_{120} = \frac{2}{3} \frac{1}{\omega_1}$$

(12)

where $\omega_1 (=\omega_{re} + \omega_s)$ is the primary angular frequency [rad/s] and $\omega_s$ is the slip angular frequency [rad/s]. By setting the speed reference as in (12), the switch state of each phase is equally delayed from the switch state change of SHE-PWM.

### 3.2 Stator Flux Linkage and Slip Angular Frequency

In order to process MPDTC in parallel with SHE-PWM, it is necessary to set the reference value to the stator flux linkage obtained by SHE-PWM at steady state. This paper explains the calculation formula of stator flux linkage and slip angular frequency for MPDTC.

MT reference frame is the rotational coordinate synchronized with the stator flux linkage. When the stator current $i_s$ and the stator flux linkage $\psi_s$ on MT reference frame are chosen as state variables, the state-space representation of the IM becomes

$$ \begin{bmatrix} i_{s,MT} \\ \psi_{s,MT} \end{bmatrix} = \begin{bmatrix} A_{11,MT} & A_{12,MT} \\ A_{21,MT} & A_{22,MT} \end{bmatrix} \begin{bmatrix} i_{s,MT} \\ \psi_{s,MT} \end{bmatrix} + \begin{bmatrix} B_{1,MT} \\ B_{2,MT} \end{bmatrix} u_{MT} $$

(13)

Fig. 8. Block diagram of proposed control system (combination of SHE-PWM and MPDTC)

Fig. 9. Speed controller using I-P speed controller

Fig. 10. Speed controller using P speed controller and disturbance observer (DOB)

where $A_{ij,MT}$ and $B_{ij,MT}$ are the matrices of the state-space representation of the IM.

Table 2. Parameters of control system

<table>
<thead>
<tr>
<th>Weighting Factor</th>
<th>$A_1$</th>
<th>70</th>
</tr>
</thead>
<tbody>
<tr>
<td>Weighting Factor</td>
<td>$A_2$</td>
<td>80</td>
</tr>
<tr>
<td>Speed Control Bandwidth</td>
<td>$\omega_{s,sc}$</td>
<td>100 rad/s</td>
</tr>
<tr>
<td>DOB Bandwidth</td>
<td>$\omega_{dis}$</td>
<td>100 rad/s</td>
</tr>
<tr>
<td>Primary Angular Freq. Ref.</td>
<td>$\omega_m^{ref}$</td>
<td>328.5326 rad/s</td>
</tr>
<tr>
<td>Inverter modulation rate</td>
<td>$m$</td>
<td>0.9</td>
</tr>
<tr>
<td>Sampling Time</td>
<td>$T_s$</td>
<td>25 $\mu$s</td>
</tr>
</tbody>
</table>

$$ \begin{equation} \begin{bmatrix} i_{s,MT} \\ \psi_{s,MT} \end{bmatrix} = \begin{bmatrix} A_{11,MT} & A_{12,MT} \\ A_{21,MT} & A_{22,MT} \end{bmatrix} \begin{bmatrix} i_{s,MT} \\ \psi_{s,MT} \end{bmatrix} + \begin{bmatrix} B_{1,MT} \\ B_{2,MT} \end{bmatrix} u_{MT} \end{equation} $$

(13)
The calculation formula of the stator flux linkage on MT-coordinate is calculated. Focus on rows 3 and 4 of (13); they become (14) and (15) when the derivative term and the T-axis stator flux linkage are zero.

\begin{align*}
v_{sM} &= R_s i_{sM} \quad \cdots \cdots \cdots \cdots \cdots (14) \\
v_{sT} &= R_s i_{sT} + \omega_1 \psi_{sM} \quad \cdots \cdots \cdots \cdots \cdots (15)
\end{align*}

The amplitude of the three-phase voltage at the inverter modulation rate \( m \) is expressed by (16).

\[ V_1 = \sqrt{3} \frac{V_{DC}}{2} \cdot m \quad \cdots \cdots \cdots \cdots \cdots (16) \]

where \( V_{DC} \) is the DC link voltage of the inverter and \( m \) is the inverter modulation rate. Using (14)–(16), the M-axis stator flux linkage at the inverter modulation rate \( m \) is expressed as

\[ \psi_{sM} = -R_s i_{sT} + \sqrt{V_1^2 - R_s^2 i_{sM}^2} \quad \cdots \cdots \cdots \cdots \cdots (17) \]

To calculate the speed reference, the slip angular frequency of the IM is calculated. Focus on rows 2 and 4 of (13); they become (18) when the derivative term and the T-axis stator flux linkage are zero.

\[ 0 = -\omega_s i_{sM} - \frac{L_s R_s}{L_s L_r - M^2} i_{sT} + \frac{\omega_1 L_s}{L_s L_r - M^2} \psi_{sM} \cdots \cdots \cdots \cdots (18) \]

The slip angular frequency of (19) is obtained inducted from (18).

\[ \omega_s = \frac{L_s R_s i_{sT}}{L_s \psi_{sM} - \left( L_s L_r - M^2 \right) i_{sM}} \quad \cdots \cdots \cdots \cdots \cdots (19) \]

### 3.3 Cost Function Formulation

By combining SHE-PWM and MPDTC, the prediction controller tracks the stator flux linkage and the torque at transient state, while preserving SHE-PWM voltage pattern at steady state. In this paper, (20) is used as the cost function to realize the performance above.

\[ c_n = \frac{v_{sHE}(k) - v_{sHE}(k)}{\omega_1} + \frac{\psi_{sHE}(k) - \psi_{sHE}(k)}{\omega_1^2} \cdot \left( \tau_{m,n}(k) - \tau_{m,n}(k + 1) \right) \]
where $v_{sHE}^*\alpha$ and $v_{sHE}^*\beta$ are the output voltages of SHE-PWM, $v_{s,n}^*$ and $v_{s,n}^*$ are the output voltage vectors ($n = 0...7$), $\omega_1^{\text{ref}} - \omega_1^{\text{res}}$ is the error of the primary angular frequency, $|\tau_m^{\text{ref}}(k) - \tau_m(k + 1)|$ is the error from the predicted value of torque ($n = 0...7$), $|\psi_s^{\text{ref}}(k) - \psi_s(k + 1)|$ is the error from the predicted value of stator flux linkage ($n = 0...7$), and $\lambda_1$ and $\lambda_2$ are weighting factors. At steady-state, since the errors of torque, stator flux linkage, and primary angular frequency become smaller, the first and second terms become the dominant terms. Therefore, at steady state, the cost function selects the same voltage vector as the output voltage of SHE-PWM. At transient state, the errors of torque, stator flux linkage, and primary angular frequency become larger, such that the third and fourth terms are the dominant terms. Therefore, the cost function selects the voltage vector such that the errors of the torque, stator flux linkage, and primary angular frequency become smaller during the transient state. In (20), $\lambda_1$ is a weighting factor that allows one to adjust a desired closed-loop performance. For a faster dynamic, a larger value for $\lambda_1$ must be chosen. Further, in (20), $\lambda_2$ is a weighting factor to reduce the error between the torque and the stator flux linkage.

### 3.4 Proposed Control System

The reference of the primary angular frequency is calculated so as to satisfy (12). The output voltage of SHE-PWM and the predicted values of torque and stator flux linkage of MPDTC are calculated from the reference of the primary angular frequency as shown in Fig. 8. Finally, the switch state of the inverter for driving the IM is selected by the cost function of (20) and (21). In this paper, $\lambda_1$ and $\lambda_2$ are determined by the cut and try on the...
4. Experimental Results

This paper considers the PWM voltage source inverter connected to the IM shown in Fig. 8. The parameters of the control system shows Table 2. Each speed controller (P speed controller or I-P speed controller) is designed with gain by the pole placement method. Figure 9 is the speed controller using the I-P speed controller in MPDTC. Figure 10 is the speed controller using P speed controller and DOB in MPDTC. The torque response $\tau_{\text{res}}$ is measured with a torque sensor. This paper uses drift avoidance (27) for the estimation calculation of stator flux linkage. The SHE-PWM eliminates the 5th-, 7th-, 11th-, 13th-, 17th-, 19th-, 23rd-, 25th-, and 29th-order harmonics.

4.1 Transient State

Figure 11 and Fig. 12 show the response waveforms of the primary angular frequency, motor torque, MT-axis current, stator flux linkage, average switching frequency of the inverter at transient-state when the IM has a load torque of $-6\text{ N-m}$ (50% load). Figure 13 and Fig. 14 show the response at steady-state when the IM has a load torque of $-6\text{ N-m}$ (50% load). From Fig. 11, it is confirmed that the torque reference has an offset at steady state. Therefore, the error of the primary angular frequency is controlled to increase by the torque reference when the IM has the 50% load torque. From Fig. 12, as the torque reference has no offset, the error of the primary angular frequency has no increase owing to the torque reference offset. From Fig. 13, it is confirmed that the torque reference has an offset because the integral value of the speed controller of MPDTC cannot be adjusted. From Fig. 14, it is confirmed that the torque reference has no offset because MPDTC speed controller has no integrator.

Table 3 shows the result of the control performance of each controller when the IM has the 50% load torque. $\omega_{\text{err}}$ is the error of the primary angular frequency, and $f^{\text{sw}}$ is the average switching frequency of the inverter at transient state. From Table 3, it is confirmed that the proposed method using a P speed controller and DOB performs better than other combining methods and SHE-PWM when the IM has the 50% load torque. It is confirmed that although the suppression
performance of the load torque of the proposed method using a P speed controller and DOB is inferior to MPDTC, the average switching frequency of the inverter at transient state is greatly reduced.

Figure 15 and Fig. 16 show the motor torque response at transient state for the torque step reference. From these figures, in the proposed method using a P speed controller and DOB, the motor torque response is nearly equal to MPDTC at the rated torque step reference.

4.2 Steady State

Figure 17–Fig. 20 show the frequency spectrum of the three-phase current at steady state. From Fig. 19 and Fig. 20, it is confirmed that the current spectrum of the proposed system using the P speed controller and DOB is nearly equal to that of SHE-PWM at steady state. Table 4 shows the total harmonic distortion and the fundamental of the three-phase current in the fast Fourier transform results of each controller and the average switching frequency (during 5.0 s) of the inverter. $I_{THD}$ is the total harmonic distortion of the three-phase current, $I_{fund}$ is the fundamental of the three-phase current, and $f_{SW}$ is the average switching frequency of the inverter at steady state. The performance of each method at steady state is compared by the average switching frequency of the inverter when the current harmonic is matched. From Table 4, in the proposed method, the reduction in the average switching frequency of the inverter is 31.3% compared with three-phase PWM. The reduction in the average switching frequency of the inverter is 10.9% compared with two-phase PWM.

5. Conclusion

This paper proposes a control system to realize a reduction in harmonic current and fast torque response by combining SHE-PWM and MPDTC for IM. The method of realizing SHE-PWM of each control cycle and the method of constructing the combined SHE-PWM and MPDTC are described. Through the experimental results, it is confirmed that the reduction in the harmonic current of the proposed method is nearly equal to that of SHE-PWM at steady state, and the torque response of the proposed method is nearly equal to that of MPDTC at the transient state. In the proposed method, the average switching frequency of the inverter is achieved a reduction of 31.3% in comparison with the three-phase PWM.

References

Motor Drive based on MPDTC and SHE-PWM (Tenjiro Hiwatari et al.)


Tenjiro Hiwatari (Student Member) received B.S. degree in Electrical, Electronics and Information Engineering from Nagaoka University of Technology, Japan in 2016. Now he is a candidate of the M.S. degree in Electrical, Electronics and Information Engineering. His research interests include motor control. He is a student member of the Institute of Electrical Engineers of Japan (IEEE).

Kiyoshi Ohishi (Fellow) received the B.S., M.S., and Ph.D. degrees, all in electrical engineering, from Keio University, Yokohama, Japan, in 1981, 1983, and 1986, respectively. From 1986 to 1993, he was an Associate Professor with Osaka Institute of Technology, Osaka, Japan. From 1993 to 2003, he was an Associate Professor with Nagaoka University of Technology, Niigata, Japan. Since August 2003, he has been a Professor at the same university. He is an administration committee member of the IEEE Industrial Electronics Society, the Institute of Electrical Engineers of Japan (IEEE), the Japan Society of Mechanical Engineers (JSME), the Society of Instrument and Control Engineers (SICE), and the Robotics Society of Japan (RSJ).

Yuki Yokokura (Member) received the B.E. and M.E. degrees in electrical engineering from Nagaoka University of Technology, Niigata, Japan, in 2007 and 2009, respectively, and Ph.D. degree in integrated design engineering from Keio University, Yokohama, Japan, in 2011. From 2010 to 2011, he was a JSPS (Japan Society for the Promotion of Science) Research Fellow (DC2 and PD). He was a Visiting Fellow at Keio University, and a Postdoctoral Fellow at Nagaoka University of Technology in 2011. Since 2012, he has been an Assistant Professor with Nagaoka University of Technology. His research interests include motion control, motor drive, power electronics, and real-world haptics.

Haruya Kada (Student Member) received B.S. degree in Electrical, Electronics and Information Engineering from Nagaoka University of Technology, Japan in 2018. Now he is a candidate of the M.S. degree in Electrical, Electronics and Information Engineering. His research interests include motor control. He is a student member of the Institute of Electrical Engineers of Japan (IEEE).

Sota Sano (Non-member) received the B.E., M.E. degree from Osaka University, Japan, in 2012, and 2014, respectively. He joined Advanced Technology R&D Center of Mitsubishi Electric Corporation in 2014. He currently research of motor drive in Advanced Technology R&D Center of Mitsubishi Electric Corp., Amagasaki, Japan.

Akira Satake (Member) received the B.E. degree from The University of Tokyo, Japan, in 1985. He joined Industrial System Laboratory of Mitsubishi Electric Corporation in 1985. His research focused on motor drive technologies, including controls for mechatronics components and power-electronic circuits. He received R&D100 awards with sensor-less servo drive system in 2014. He is currently a group manager for research of motor drive in Advanced Technology R&D Center of Mitsubishi Electric Corp., Amagasaki, Japan.