Online Dead Time Effect Compensation Algorithm of PWM Inverter for Motor Drive Using PR Controller

Chang-Seok Park* and Tae-Uk Jung†

Abstract – This paper proposes the dead time effect compensation algorithm using proportional resonant controller in pulse width modulation inverter of motor drive. To avoid a short circuit in the dc link, the dead time of the switch device is surely required. However, the dead time effect causes the phase current distortions, torque pulsations, and degradations of control performance. To solve these problems, the output current including ripple components on the synchronous reference frame and stationary reference frame are analyzed in detail. As a results, the distorted synchronous d-and q-axis currents contain the 6th, 12th, and the higher harmonic components due to the influence of dead time effect. In this paper, a new dead time effect compensation algorithm using proportional resonant controller is also proposed to reduce the output current harmonics due to the dead time and nonlinear characteristics of the switching devices. The proposed compensation algorithm does not require any additional hardware and the offline experimental measurements. The experimental results are presented to demonstrate the effectiveness of the proposed dead time effect compensation algorithm.

Key words: Dead time, Compensator, PWM inverter, PMSM drive

1. Introduction

Voltage Source Inverter(VSI) vector controlled motor drives have been widely developed. There are several causes to distort output voltage. Some is caused by the dead time that is to avoid shoot-through in the dc link. Others originate from the nonlinear characteristics of the switching devices such as turn-on/off time and voltages drop of switches and diodes. The distortion of the output voltage results in phase current distortions, torque pulsations, and degradations of control performance [1-5]. Since these undesirable ripples bring about bad influences to motor driving system, a compensation algorithm must be needed in the control algorithm of the motor drive. To solve the nonlinear characteristics of the VSI, various solutions have already been suggested [1-13]. The lost voltage average value calculation [2, 3], the pulse width modulation(PWM) pulse based technique [6], voltage feed forward method [7-12], and the support vector regression(SVR) method [13] are the main categories of the conventional dead time compensation methods.

In PWM pulse based technique, the compensation is realized for each PWM pulse. In [8, 9], the methods are dependent on the phase current polarity. In the voltage feed forward method, the methods are based on a feedforward method, where the compensated voltages are fed to the reference voltages in order to generate a modified voltage. The compensation voltages are calculated by using the dead time, switching period, current command, and dc-link voltage. This approach can compensate the fundamental and harmonic components of the voltage error in the d- and q-axis frame. In the calculation of the compensating signals, nonlinear characteristics of the switching devices such as finite switching times and voltage drops are also considered. However, these methods can be implemented only by offline methods because the switching times and voltage drops of switching devices and diodes are varied with operating conditions such as dc-link voltage, phase currents, and motor speed. Thus, it is difficult to accurately compensate for the dead time effects by offline methods [12]. In [13], this method is based on an emerging learning technique SVR. The SVR technique is difficult to implement because SVR model method is required some parameters, online computation, and extra memory to construct the regression function.

In this paper, the output current including ripple components on the synchronous reference frame and stationary reference frame are analyzed in detail. Moreover, a new dead time effect compensation algorithm using proportional resonant(PR) controller is also proposed to reduce the output current harmonics due to the dead time and nonlinear characteristics of the switching devices. The proposed compensation algorithm does not require any additional hardware and the off-line experimental to detect the dead time effect contrast with the existing method. Also, it is a very simple controller by adding feed-forward compensation. The simulation and the experimental results are presented to verify the usefulness of the proposed algorithm.
2. System Modeling

2.1 Dead time effects analysis [4, 15]

The switching device has a finite switching time, a blank-time or normally called dead-time must be inserted into the PWM gating signals in order to avoid the conduction overlap of two switching devices in the same leg. This dead-time causes the phase error, output voltage distortions and fundamental voltage drop, which degrade the control performance.

The commonly used three-phase PWM VSI with a permanent magnet synchronous motor (PMSM) load is shown in Fig. 1.

Fig. 2(a) shows the a-phase current $i_{an}$ flow in positive direction and Fig. 2(b) displayed the a-phase current $i_{an}$ flow in negative direction in a-phase leg of the PWM VSI. In Fig. 2(a), the a-phase current $i_{an}$ flows through switching device $S1$ during the on-time of $S1$. Conversely, the a-phase current $i_{an}$ flows through diode of $S2$ during both the off-time of $S1$ and dead time $T_{dead}$. In Fig. 2(b), the a-phase current $i_{an}$ flows through switching device $S2$ during the on-time of $S2$. Conversely, the a-phase current $i_{an}$ flows through diode of $S1$ during both the off-time of $S2$.

The relationship between ideal and real PWM VSI output voltage in case of positive and negative current direction is shown in Fig. 3. Where $S1$ and $S2$ are PWM gate signal, $V_{an}$ is output voltage of the PWM VSI, $ΔV$ is the distorted voltage respectively. As shown in Fig. 3, the real output voltage $V_{an}$ of PWM VSI is directly influenced by the dead time and turn-on/turn-off delay time of switching device.

From Fig. 3, the average distorted voltage $ΔV$ according to the direction of the a-phase current $i_{an}$ can be represented as [14, 15].

$$ΔV = \frac{T_{dead} - t_{on} + t_{off}}{2T_s} V_{dc} \quad (i_{an} > 0)$$

$$ΔV = \frac{T_{dead} + t_{on} - t_{off}}{2T_s} V_{dc} \quad (i_{an} < 0)$$

where $T_s$ is the sampling period of the current regulator, $t_{on}$ is the turn-on delay time of switching device, $t_{off}$ is the turn-on delay time of switching device, respectively.

The absence of a neutral connection in the motor forces the constraint that [7]:

$$i_{as} + i_{bs} + i_{cs} = 0$$

For any balanced load, the line to neutral voltages are constrained such that :

$$V_{an} + V_{bn} + V_{cn} = 0$$

Therefore, the distorted voltage of three-phases are calculated by (5).

$$ΔV_{as} = ΔV \frac{1}{3} \{2\text{sign}(i_{as}) - \text{sign}(i_{bs}) - \text{sign}(i_{cs})\}$$

$$ΔV_{bs} = ΔV \frac{1}{3} \{2\text{sign}(i_{bs}) - \text{sign}(i_{as}) - \text{sign}(i_{cs})\}$$

$$ΔV_{cs} = ΔV \frac{1}{3} \{2\text{sign}(i_{cs}) - \text{sign}(i_{as}) - \text{sign}(i_{bs})\}$$

If the phase current flow in positive direction, $\text{sign}(i_{as,bs,cs})$ is $+1$. Conversely, if the phase current flow in negative direction, $\text{sign}(i_{as,bs,cs})$ is $-1$.

By (5), the distorted phase voltage $ΔV$ and three-phase current $i_{a}$ can be drawn as Fig. 4. Also, Fig. 5. is shown the distorted voltage and the $dq$-axis currents of the stationary reference frame.

In order to analysis the harmonic component by dead time effect, Fourier series is applied. In case of square-
wave voltage, the harmonic component of voltage is calculated as follows:

\[ v_h = \frac{4V_s}{\pi} \sum_{h=1,3,5,7...} \frac{1}{h} \sin(h\omega_b t) \]  

(6)

where \( v_h \) is the harmonic component of voltage, \( V_s \) is the square-wave voltage, \( h \) is harmonic order and \( \omega_b \) is the motor synchronous frequency (rad/s).

Therefore, the distorted voltage \( \Delta v'_{ar} \) and \( \Delta v'_{br} \) can be obtained by (7), (8).

\[ \Delta v'_{ar} = \frac{4AV_s}{\pi} \{ \sin(5\omega_b t) + \frac{1}{5} \sin(3\omega_b t) + \frac{1}{7} \sin(7\omega_b t) + \frac{1}{11} \sin(11\omega_b t) + \frac{1}{13} \sin(13\omega_b t) + \ldots \} \]  

(7)

\[ \Delta v'_{br} = \frac{4AV_s}{\pi} \{ -\cos(5\omega_b t) + \frac{1}{5} \cos(3\omega_b t) + \frac{1}{7} \cos(7\omega_b t) + \frac{1}{11} \cos(11\omega_b t) + \frac{1}{13} \cos(13\omega_b t) + \ldots \} \]  

(8)

Similarly, the distorted current \( i_h \) which corresponds to the distorted voltage can be calculated by (9).

\[ i_h = \frac{4AV_s}{\pi} \sum_{h=1,3,5,7...} \frac{1}{hZ_h} \sin(h\omega_b t - \phi_h) \]  

(9)

where \( Z_h \) is the load impedance, \( \phi_h \) is the load impedance angle respectively.

From (9), the distorted \( dp \)-axis currents of the stationary reference frame can be expressed as (10), (11).

\[ \Delta i'_{ar} = \frac{4AV_s}{Z} \{ \sin(5\omega_b t - \phi) + \frac{1}{5Z} \sin(5\omega_b t - \phi_1) \]  

\[ + \frac{1}{7Z} \sin(7\omega_b t - \phi) + \frac{1}{11Z} \sin(11\omega_b t - \phi_1) \]  

\[ + \frac{1}{13Z} \sin(13\omega_b t - \phi_1) + \ldots \} \]  

(10)

\[ \Delta i'_{br} = \frac{4AV_s}{Z} \{ -\cos(5\omega_b t - \phi) + \frac{1}{5Z} \cos(5\omega_b t - \phi_1) \]  

\[ - \frac{1}{7Z} \cos(7\omega_b t - \phi) + \frac{1}{11Z} \cos(11\omega_b t - \phi_1) \]  

\[ - \frac{1}{13Z} \cos(13\omega_b t - \phi_1) + \ldots \} \]  

(11)

In order to simple analysis, the distorted voltages and the distorted currents of the synchronous reference frame can be rewritten by (12)-(15).

\[ \Delta v'_{dr} = \frac{4AV_s}{\pi} \{ \sin 6\omega_b t + \frac{24}{143} \sin 12\omega_b t + \ldots \} \]  

(12)

\[ \Delta v'_{qr} = \frac{4AV_s}{\pi} \{ -\cos 6\omega_b t + \frac{2}{143} \cos 12\omega_b t + \ldots \} \]  

(13)

\[ \Delta i'_{dr} = \frac{4AV_s}{Z} \{ \sin 6\omega_b t - \phi \} \]  

\[ + \frac{24}{143Z} \sin 12\omega_b t - \phi_1 + \ldots \} \]  

(14)

\[ \Delta i'_{qr} = \frac{4AV_s}{\pi} \{ -\cos 6\omega_b t - \phi \} \]  

\[ + \frac{2}{143Z} \cos 12\omega_b t - \phi_1 + \ldots \} \]  

(15)

As shown in (14) and (15), the distorted synchronous d-and q-axis current contain the 6th, 12th, and the higher harmonic components due to the influence of dead time effect. The 6th harmonic component is dominated by (14) and (15). The dead time effect can be mitigated by the appropriately harmonic component mitigation, either in stationary or in synchronous reference frame.

3. Proposed Dead Time Compensation Algorithm

3.1 Design of PR controller

In general, PR controller is used to remove the certain
order harmonics for the current regulator because the PR controller has zero steady-state error for sinusoidal waveforms having the same synchronous frequency as \( \omega_r \) [16-19].

The ideal PR control, which is based on an internal modern theory, is expressed as (16).

\[
G_s = K_p + \frac{K_i \omega_r s}{s^2 + \omega_r^2}
\]  

(16)

Where \( \omega_r \) is the resonant frequency, \( K_p \) and \( K_i \) represent proportional and resonant gains respectively. For \( K_p \), it is tuned in the same way as for a PI controller, and it basically determines the dynamics of the system in terms of bandwidth, phase and gain margin and \( K_i \) could be tuned for shifting the magnitude response vertically but this does not give rise to a significant variation in bandwidth [20]. The Bode plots of ideal PR control method are shown in Fig. 6.

The ideal PR controller has an infinite gain at the resonant frequency and no phase shift and gain at other frequencies. However, the controller’s infinite gain may cause stability problems and has the practical limitations of signal processing systems. Because of these problems, the non-ideal PR controller transfer function (17) is used instead of ideal PR controller (16).

\[
G_s = K_p + \frac{K_i \omega_r s}{s^2 + 2 \omega_a s + \omega_r^2}
\]

(17)

The Bode plots of non-ideal PR control method are shown in Fig. 7(a) and 7(b).

By applying the bilinear transformation \( S_z \) and substituting in to non-ideal trans function (17), the discrete trans function of the PR controller can be obtained by (18).

\[
S_z = \frac{2}{T_s} \frac{1 - z^{-1}}{1 + z^{-1}}, \quad G_z = \frac{n_0 + n_1 z^{-1} + n_2 z^{-2}}{1 + d_1 z^{-1} + d_2 z^{-2}}
\]

(18)

where \( T_s \) is the sampling time.

\[
\begin{align*}
n_0 &= \frac{(4 + 4T_s \omega_a + \alpha \omega_r^2 T_s^2)K_p + 4K_i T_s \omega_r}{4 + 4T_s \omega_a + \alpha \omega_r^2 T_s^2} \; ; \\
n_1 &= \frac{(-8 + 2 \alpha \omega_r^2 T_s^2)K_p}{4 + 4T_s \omega_a + \alpha \omega_r^2 T_s^2} \; ; \\
n_2 &= \frac{(4 - 4T_s \omega_a + \alpha \omega_r^2 T_s^2)K_p - 4K_i T_s \omega_r}{4 + 4T_s \omega_a + \alpha \omega_r^2 T_s^2} \; ; \\
d_1 &= \frac{-8 + 2 \alpha \omega_r^2 T_s^2}{4 + 4T_s \omega_a + \alpha \omega_r^2 T_s^2} \; ; \\
d_2 &= \frac{4 - 4T_s \omega_a + \alpha \omega_r^2 T_s^2}{4 + 4T_s \omega_a + \alpha \omega_r^2 T_s^2}
\end{align*}
\]

Therefore, the digital equation of non-ideal PR controller can be given by (19).

\[
y(k) = n_0 u(k) + n_1 u(k-1) + n_2 u(k-2) - d_1 y(k-1) - d_2 y(k-2)
\]

(19)

The block diagram of non-ideal PR controller is shown in Fig. 8.
Table 1. The properties of a SPMSM drive parameter

<table>
<thead>
<tr>
<th>Properties SPMSM</th>
<th>Details</th>
</tr>
</thead>
<tbody>
<tr>
<td>Rated voltage</td>
<td>25 [V]</td>
</tr>
<tr>
<td>Rated current</td>
<td>2.8 [A]</td>
</tr>
<tr>
<td>Rated speed</td>
<td>3000 [r/min]</td>
</tr>
<tr>
<td>Maximum speed</td>
<td>3500 [r/min]</td>
</tr>
<tr>
<td>Poles</td>
<td>4</td>
</tr>
<tr>
<td>Stator resistance</td>
<td>0.75 [Ω]</td>
</tr>
<tr>
<td>Stator inductance</td>
<td>0.85 [mH]</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Properties motor drive</th>
<th>Details</th>
</tr>
</thead>
<tbody>
<tr>
<td>Control processor</td>
<td>DSP TMS320F28335</td>
</tr>
<tr>
<td>Switching frequency</td>
<td>10 [kHz]</td>
</tr>
<tr>
<td>Current regulator period</td>
<td>100 [μs]</td>
</tr>
<tr>
<td>Speed regulator period</td>
<td>1000 [μs]</td>
</tr>
<tr>
<td>Dead time</td>
<td>3 [μs]</td>
</tr>
<tr>
<td>Turn on time of switch device</td>
<td>0.1 – 0.4 [μs]</td>
</tr>
<tr>
<td>Turn off time of switch device</td>
<td>0.2 – 1.1 [μs]</td>
</tr>
</tbody>
</table>

3.2 Proposed dead time compensation method

The distorted synchronous d-and q-axis currents contain the 6th, 12th, and the higher harmonic components due to the influence of dead time effect. Fig. 9 shows the block diagram of the proposed compensation algorithm.

Firstly, the synchronous d- and q-axis currents are passed through the PR controller to extract the 6th, 12th harmonics components in parallel with PI current regulator. Secondly, the output of PR controller is used to feed-forward compensation of PI current regulator output. Hence, the harmonic components due to the dead time effect can be eliminated.

4. Experimental Results

The proposed dead time effect compensation algorithm was implemented with a SPMSM drive system. The parameters given in Table 1. The drive system for experimental is shown in Fig. 10.

Fig. 11(a), (b), and (c) show the experimental waveforms without the dead time compensation when the motor operates at 3000[r/min] at no-load. The phase current \(i_{as}\) with distorted waveform is shown in Fig. 11(b). As shown in Fig. 11(b), the FFT result of the phase current \(i_{as}\) have the 5th and 7th harmonic components, and the higher harmonic components. However, the FFT result contains very small 11th and 13th harmonic components rather than 5th and 7th harmonic components by (10). It means that the distorted waveform is not significantly affected by the 11th and 13th harmonic components. The d- and q-axis currents of the synchronous reference frame have the 6th harmonic component, as shown in Fig. 11(c).

Fig. 13 shows the experimental results waveforms of the proposed compensator. With the proposed method, the x-y plot of the Fig. 12(a) displays circular waveforms rather than Fig. 11(a) and the phase current \(i_{as}\) has the nearly sinusoidal waveform as shown in Fig. 12(b). The harmonic components of the phase current \(i_{as}\) are considerably reduced by the proposed algorithm. Likewise, the dq-axes currents of the synchronous reference frame and the FFT results show the reduced harmonics, as shown in Fig. 12(c).

Fig. 13(a), (b), and (c) illustrate the experimental results without the dead time compensation when the motor operates at 3000[r/min] in the 30[W] output. Similar to the
no-load condition, the x-y plot of the stationary reference frame current $i_{ds}$ and $i_{qs}$ display the hexagonal waveform. The FFT result of the phase current $i_{a}$ have the 5th and 7th harmonic components, and the higher harmonic components. Also, the FFT result contains small 11th and 13th harmonic components rather than 5th and 7th harmonics.

Considering the proposed dead time compensation algorithm, the x-y plot of the Fig. 14(a) shows circular waveforms rather than Fig. 13(a) and the phase current $i_{a}$ has the pure sinusoidal waveform as shown in Fig. 14(b). The harmonics of the phase current $i_{a}$ are considerably reduced by the dead time compensation. Likewise, the dq-axes currents of the synchronous

![Fig. 11. Current waveforms without dead time compensation under no-load](image1)

![Fig. 12. Current waveforms with dead time compensation under no-load](image2)
reference frame and the FFT results show the reduced harmonics, as shown in Fig. 14(c).

5. Conclusions

The dead time compensation algorithm for the motor drive system is proposed. The proposed algorithm is comprised of the PR controller in parallel with the general PI current regulator. The output of PR controller is used to feed-forward compensation of PI current regulator output. Hence, the harmonic components due to the dead time effect can be eliminated.

This algorithm does not require any external hard ware,
off-line experimental measurements. The proposed method can be easily implemented and applied in the system. The validity of the proposed algorithm is proved by the several experimental tests.

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References

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