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Hybrid SVPWM Scheme to Minimize the Common-Mode Voltage Frequency and Amplitude in Voltage Source Inverter Drives

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Abstract—In this paper, a hybrid space vector pulsewidth modulation synthesis is proposed to lower the frequency and amplitude of the common-mode voltage (CMV). The conventional space vector pulsewidth modulation (SVPWM) techniques for inverters generate a high-frequency CMV. The CMV causes significant bearing current flowing through the path to the ground and leads to premature bearing failure, especially in high switching frequency drive systems. Furthermore, previously proposed CMV reduction studies focus on the reduction of CMV amplitude, leaving a high-frequency CMV. The proposed control algorithm divides each sector in the space vector hexagon into three segments based on the modulation index and the angle of the reference voltage vector. In each segment, the space vectors that produce minimal CMV are selected to synthesize the reference voltage vector. A fair comparison with the conventional CMV reduction methods is made by comparing the CMV characteristics, harmonic distortion factor, dc current ripple, and the switching losses for each method. Both simulation and experimental results have validated the effectiveness of the proposed approach.

Index Terms—Common-mode voltage (CMV), dc–ac adjustable frequency drive systems, space vector pulsewidth modulation (SVPWM).

I. INTRODUCTION

V OLTAGE source inverters (VSIs) are widely used in motor drive applications. The use of VSI ranges from high power (e.g., wind turbines) to medium power (e.g., electric and hybrid electric vehicles). Ever since the invention of the insulated gate bipolar transistor (IGBT), increasing the switching frequency of the power converter has been a trend due to the reduction of the footprint and the increase of power density [1], [2]. Recent development of wide bandgap devices has equipped the power switches with even faster switching speed [3]. However, the increase in the switching frequency has generated several unwanted consequences, one of which is the high-frequency common-mode voltage (CMV) [4], [6]. The high-frequency CMV can cause electromagnetic interference that adversely affects the other components of the system [7]. The induced overvoltage stresses the winding insulation of the drives and can also increase the shaft current [8]. In particular, the shaft current is the result of the fluctuation of the CMV, which is strictly related to the sequence of the inverter switching states [9]. Several studies presented in the literature discussed the reduction of CMV effects. They can be classified into hardware solutions and algorithmic solutions. The hardware ones require either increasing the number of switches [10]–[15], passive components [16]–[25], or both [26], [27]. The study in this paper is mostly related to the algorithmic approaches that require no additional hardware.

The early attempt to reduce the CMV proposed in [9] uses only the odd or only the even voltage vectors. This approach results in dc CMV. However, the linear region in this approach is reduced to \( \text{MI} < \frac{1}{\sqrt{3}} \). Later, a number of CMV reduction methods were proposed such as the active zero-state PWM (AZSPWM1-2 and AZSPWM3), remote state PWM (RSPWM1, RSPWM2, and RSPWM3), and near-state PWM (NSPWM) [28]–[31]. In each one, several active voltage vectors are utilized to meet the voltage second requirements and avoid using the zero vectors. The nonzero switching vectors generate lower magnitude CMV but the conventional space vector PWM uses zero vectors as it helps in reducing the THD [32], [35]. The latter methods utilize all of the linear region (except for NSPWM), which leaves the high frequency of the CMV remaining an issue. The high common-mode \( \frac{dv}{dt} \) causes electric discharge machining (EDM) across the bearing races. Furthermore, the CMV reduction methods assume ideal switches and fail when nonlinearities such as the deadtime and line–line voltage reversal are considered [34].

On the other hand, to further improve the performance of CMV reduction methods, combined approaches are used in different operating conditions. For instance, to make full use of NSPWM, another CMV reduction method is activated at low modulation index such as AZSPWM1 [36]. The modified SVPWM presented in [37] combines both RSPWM and regular SVPWM at low- and high-modulation index, respectively. This method makes a tradeoff between dc-link voltage utilization and common-mode current (CMC) reduction. However, the deadtime effect causes CMV amplitude to reach half of dc-link voltage [38]–[40].

In this paper, a hybrid approach for minimizing the amplitude and the frequency of the CMV is proposed. The control concept...
is based on dividing the space vector hexagon into various segments and synthesizing the reference voltage vector by using vectors that correspond to a minimal CMV. Full treatment of the deadtime effect is presented. The proposed method avoids any CMV peaks or unwanted variations during the deadtime intervals.

The rest of the paper is organized as follows. In Section II, the regular SVPWM is briefly revisited. The proposed control method is presented in Section III. Deadtime effect is presented in Section IV. A complete solution to the deadtime effect is presented in Section V. Performance analyses including CMV characteristics, harmonic distortion factor (HDF), dc current ripple, and the switching losses are presented in Section VI. Hardware results and conclusion are presented in Sections VII and VIII, respectively.

II. CONVENTIONAL SVPWM

For the two-level, three-phase VSI shown in Fig. 1, the reference value of the inverter output voltage can be represented by the following space vector:

$$V_{n r e f} = \frac{2}{3} \left( v_a e^{j \frac{2\pi}{3}} + v_b e^{-j \frac{2\pi}{3}} + v_c e^{-j \frac{2\pi}{3}} \right)$$

where $v_a$, $v_b$, and $v_c$ are the three phase voltages. Since the dc-link should not be shorted, each phase voltage can only attain either $V_{dc}$ or $-V_{dc}$. This also restricts the feasible switching states to only eight states. The eight converter states as well as the corresponding differential-mode voltage (DMV) and CMV are shown in Table I. The linear combination of the possible eight vectors span a hexagonal area, as shown in Fig. 2.

The reference vector $V_{ref}$ is synthesized at each sampling period $T_s$, using the two adjacent active vectors and the zero vectors. This leads to six equal sectors. Because the circular trajectory of $V_{ref}$ in the complex plane corresponds to a sinusoidal three-phase voltage, the maximum achievable sinusoidal output voltage amplitude is $\leq \frac{2}{\sqrt{3}} V_{dc}$. Therefore, the modulation index MI can be defined as the per unit output voltage vector using $V_{dc}$ as the base voltage $MI \in \left[0, \frac{2}{\sqrt{3}} \right]$.

For instance, when the trajectory of the reference voltage $V_{ref}$ is passing through the first sector where $\theta \in \left[0, \frac{\pi}{3} \right]$, the following volt-seconds equality holds:

$$T_s \cdot V_{ref} = t_{v1} \cdot V_{(1,0,0)} + t_{v2} \cdot V_{(1,1,0)} + t_{v0} \cdot V_{(0,0,0)} + t_{v7} \cdot V_{(1,1,1)}$$

$$T_s = t_{v1} + t_{v2} + t_{v0} + t_{v7}.$$  \hspace{1cm} (2)

By equating the real part and the imaginary part in (2), the following dwell times can be obtained:

$$t_{v1} = \frac{3}{4} T_s MI \left[ \cos(\theta) - \sin(\theta) \right]$$ \hspace{1cm} (4)

$$t_{v2} = \frac{\sqrt{3}}{2} T_s MI \sin(\theta)$$ \hspace{1cm} (5)

$$t_{v0} = t_{v7} = T_s - t_{v1} - t_{v2}.$$ \hspace{1cm} (6)

The same rules can be applied for calculating the dwell times of the vectors for sectors 2–6 if the following modified $\theta_k$ is used:

$$\theta_k = \theta - (k - 1) \frac{\pi}{3}$$ \hspace{1cm} (7)

where $k$ is the number of the sector in which the $V_{ref}$ resides.

III. HYBRID SPACE VECTOR PULSEWIDTH MODULATION SYNTHESIS

The method of hybrid space vector pulsewidth modulation synthesis (HSPWM) further subdivides each sector into three segments. In each segment, a group of active voltage vectors is selected to match the output reference volt-seconds and achieve minimal CMV amplitude and frequency. The location of the reference voltage vector can be defined in correspondence to the modulation index MI and its angle $\theta$. 

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**TABLE I**

<table>
<thead>
<tr>
<th>Dwelling time</th>
<th>Voltage vector</th>
<th>$v_a$</th>
<th>$v_b$</th>
<th>$v_c$</th>
<th>$v_{dm}$</th>
<th>$v_{cm}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>$t_{v0}$</td>
<td>$V_{(0,0,0)}$</td>
<td>$-V_a$</td>
<td>$-V_b$</td>
<td>$-V_c$</td>
<td>0</td>
<td>$V_{dc}$</td>
</tr>
<tr>
<td>$t_{v1}$</td>
<td>$V_{(1,0,0)}$</td>
<td>$V_a$</td>
<td>$-V_b$</td>
<td>$-V_c$</td>
<td>$\frac{2}{3} V_{dc}$</td>
<td>$-V_{dc}$</td>
</tr>
<tr>
<td>$t_{v2}$</td>
<td>$V_{(1,1,0)}$</td>
<td>$V_a$</td>
<td>$V_b$</td>
<td>$-V_c$</td>
<td>$\frac{2}{3} V_{dc}$</td>
<td>$V_{dc}$</td>
</tr>
<tr>
<td>$t_{v3}$</td>
<td>$V_{(0,1,0)}$</td>
<td>$-V_a$</td>
<td>$V_b$</td>
<td>$-V_c$</td>
<td>$-\frac{2}{3} V_{dc}$</td>
<td>$-V_{dc}$</td>
</tr>
<tr>
<td>$t_{v4}$</td>
<td>$V_{(0,1,1)}$</td>
<td>$-V_a$</td>
<td>$-V_b$</td>
<td>$V_c$</td>
<td>$\frac{2}{3} V_{dc}$</td>
<td>$V_{dc}$</td>
</tr>
<tr>
<td>$t_{v5}$</td>
<td>$V_{(0,0,1)}$</td>
<td>$V_a$</td>
<td>$-V_b$</td>
<td>$V_c$</td>
<td>$-\frac{2}{3} V_{dc}$</td>
<td>$-V_{dc}$</td>
</tr>
<tr>
<td>$t_{v6}$</td>
<td>$V_{(1,0,1)}$</td>
<td>$V_a$</td>
<td>$V_b$</td>
<td>$V_c$</td>
<td>$\frac{2}{3} V_{dc}$</td>
<td>$V_{dc}$</td>
</tr>
<tr>
<td>$t_{v7}$</td>
<td>$V_{(1,1,1)}$</td>
<td>$V_a$</td>
<td>$-V_b$</td>
<td>$V_c$</td>
<td>$-\frac{2}{3} V_{dc}$</td>
<td>$-V_{dc}$</td>
</tr>
</tbody>
</table>
The total eighteen segments of the space vector hexagon are obtained based on their corresponding CMV. They can be categorized into the following:

1) six odd segments that result in CMV $-\frac{V_{dc}}{3}$;
2) six even segments that result in CMV $\frac{V_{dc}}{3}$;
3) six odd–even segments that result in CMV $\pm \frac{V_{dc}}{3}$.

To understand the navigation of the reference voltage through those segments, the authors thoroughly explain the synthesis of the reference voltage vector via the first sector $\theta \in [0, \frac{\pi}{3})$ with reference to Fig. 3.

The sector includes three segments (triangles). The reference vector in the bottom triangle (odd-triangle) can be synthesized using only the odd vectors $\{V_{(1,0,0)}, V_{(0,1,0)}, V_{(0,0,1)}\}$, as shown in Fig. 3(a). The reference vector in the upper right triangle (odd–even triangle) can be synthesized using the odd and even vectors $\{V_{(1,0,0)}, V_{(1,1,0)}, ..., V_{(1,0,1)}\}$, as shown in Fig. 3(b). The reference vector in the upper left triangle (even triangle) can be synthesized using only the even vectors $\{V_{(1,1,0)}, V_{(0,1,1)}, V_{(1,0,1)}\}$, as shown in Fig. 3(c).

The line dividing the odd triangle from the odd–even triangle can be parameterized as follows:

$$\text{edge} = \frac{2}{\sqrt{3}} \left(1 - \frac{\theta}{\pi/6}\right).$$

(8)

When the MI is larger than the value of edge, the reference voltage vector is in the odd triangle. Similar logic can be followed to identify all other segments.

The synthesis of the reference voltage vector in the odd triangle and even triangle is unique, whereas the synthesis for the reference voltage vector in the odd–even triangle can be realized using different groups of vectors. The groups are classified into four groups with each group producing similar harmonic content and switching losses. The four groups are delineated in the following sections.

A. HSVPWMS I

The voltage reference vector passing through the odd–even triangles can be synthesized by using the two adjacent voltage vectors and the two voltage vectors of the neighboring states to match the output and reference volt-seconds, as shown in Fig. 4(b). For the odd triangle shown in Fig. 4(a), the time intervals in which the inverter states $\{(V_{(1,0,0)}, V_{(0,1,0)}, V_{(0,0,1)}\}$ are applied can be calculated by solving the following algebraic equations:

$$V_{\text{ref}}T_s = V_{(1,0,0)}t_{v1} + V_{(0,1,0)}t_{v3} + V_{(0,0,1)}t_{v5}$$

(9)

$$T_s = t_{v1} + t_{v3} + t_{v5}.$$  

(10)
This leads to the following dwell times:

\[ t_{v1} = \frac{T_s}{3} + \frac{T_s \cos(\theta)}{2} \]  
\[ t_{v3} = \frac{T_s}{3} - \frac{T_s \cos(\theta)}{4} + \frac{\sqrt{3} T_s \sin(\theta)}{4} \]  
\[ t_{v5} = \frac{T_s}{3} - \frac{T_s \cos(\theta)}{4} - \frac{\sqrt{3} T_s \sin(\theta)}{4} \]  

For the odd–even triangle shown in Fig. 4(b), the time intervals in which the inverter states \{\(V_{(1,0,0)}, V_{(1,0,0)}\), \(V_{(0,0,0)}, V_{(0,0,0)}\)\} are applied can be calculated by solving the following algebraic equations:

\[ V_{\text{ref}} T_s = V_{(1,0,0)} t_{v6} + V_{(1,0,0)} t_{v1} + V_{(1,1,0)} t_{v2} + V_{(0,1,0)} t_{v3} \]  
\[ T_s = t_{v6} + t_{v1} + t_{v2} + t_{v3} \]  

To solve the algebraic equations, we assume \(t_{v6} = t_{v3}\). This leads to the following dwell times:

\[ t_{v1} = \frac{3 T_s \cos(\theta)}{4} - \frac{\sqrt{3} T_s \sin(\theta)}{4} \]  
\[ t_{v2} = \frac{\sqrt{3} T_s \sin(\theta)}{2} \]  
\[ t_{v3} = \frac{T_s}{2} - \frac{3 T_s \cos(\theta)}{8} - \frac{\sqrt{3} T_s \sin(\theta)}{8} \]  

For the even triangle shown in Fig. 4(c), the time intervals in which the inverter states \{\(V_{(1,1,0)}, V_{(0,1,1)}, V_{(1,0,1)}\)\} are applied can be calculated by solving the following algebraic equations:

\[ V_{\text{ref}} T_s = V_{(1,0,0)} t_{v2} + V_{(0,1,1)} t_{v4} + V_{(1,0,1)} t_{v6} \]  
\[ T_s = t_{v2} + t_{v4} + t_{v6} \]  

This leads to the following dwell times:

\[ t_{v2} = \frac{T_s}{3} + \frac{T_s \cos(\theta)}{4} + \frac{\sqrt{3} T_s \sin(\theta)}{4} \]  
\[ t_{v4} = \frac{T_s}{3} - \frac{T_s \cos(\theta)}{2} \]  
\[ t_{v6} = \frac{T_s}{3} + \frac{T_s \cos(\theta)}{4} - \frac{\sqrt{3} T_s \sin(\theta)}{4} \]  

The similar calculation can be conducted for the other five sectors by employing the appropriate vectors as listed in Table II.

**Table II**

<table>
<thead>
<tr>
<th>Sector</th>
<th>Odd triangle</th>
<th>Odd-even triangle</th>
<th>Even triangle</th>
</tr>
</thead>
<tbody>
<tr>
<td>I</td>
<td>{(V_{(1,0,0)}, V_{(0,1,0)}), (V_{(0,0,0)})}</td>
<td>{(V_{(1,0,1)}, V_{(1,0,0)}, V_{(1,1,0)}), (V_{(0,0,1)})}</td>
<td>{(V_{(1,1,0)}, V_{(0,1,1)}, V_{(1,0,1)})}</td>
</tr>
<tr>
<td>II</td>
<td>{(V_{(1,0,0)}, V_{(1,0,0)}), (V_{(0,0,0)})}</td>
<td>{(V_{(1,0,1)}, V_{(1,0,1)}, V_{(0,0,0)})}</td>
<td>{(V_{(1,1,0)}, V_{(0,1,1)}, V_{(1,1,0)})}</td>
</tr>
<tr>
<td>III</td>
<td>{(V_{(1,0,0)}, V_{(0,1,0)}), (V_{(0,0,1)})}</td>
<td>{(V_{(1,1,0)}, V_{(0,1,1)}, V_{(1,0,0)})}</td>
<td>{(V_{(1,1,0)}, V_{(0,1,1)}, V_{(1,1,1)})}</td>
</tr>
<tr>
<td>IV</td>
<td>{(V_{(0,1,0)}, V_{(0,1,0)}), (V_{(0,0,0)})}</td>
<td>{(V_{(0,1,0)}, V_{(0,1,0)}, V_{(0,0,0)})}</td>
<td>{(V_{(1,1,0)}, V_{(0,1,1)}, V_{(0,1,1)})}</td>
</tr>
<tr>
<td>V</td>
<td>{(V_{(1,0,0)}, V_{(1,0,0)}), (V_{(0,1,0)})}</td>
<td>{(V_{(1,1,0)}, V_{(0,1,1)}, V_{(1,0,0)})}</td>
<td>{(V_{(0,1,1)}, V_{(0,1,1)}, V_{(0,0,0)})}</td>
</tr>
<tr>
<td>VI</td>
<td>{(V_{(1,0,0)}, V_{(1,0,0)}), (V_{(0,0,0)})}</td>
<td>{(V_{(1,0,1)}, V_{(1,0,0)}, V_{(1,0,0)})}</td>
<td>{(V_{(1,1,0)}, V_{(0,1,1)}, V_{(1,0,1)})}</td>
</tr>
</tbody>
</table>

This leads to the following dwell times:

\[ t_{v1} = \frac{T_s}{3} + \frac{T_s \cos(\theta)}{2} \]  
\[ t_{v3} = \frac{T_s}{3} - \frac{T_s \cos(\theta)}{4} + \frac{\sqrt{3} T_s \sin(\theta)}{4} \]  
\[ t_{v5} = \frac{T_s}{3} - \frac{T_s \cos(\theta)}{4} - \frac{\sqrt{3} T_s \sin(\theta)}{4} \]  

B. **HSVPWMS II**

The voltage reference vector passing through the odd–even triangles can be synthesized by using the two adjacent voltage vectors and one voltage vector of either neighboring state to match the output and reference volt-seconds, as shown in Fig. 5. Table III shows the vector combinations used in the three segments of each sector for HSVPWMS II.

**Table III**

<table>
<thead>
<tr>
<th>Sector</th>
<th>Odd-even triangle</th>
</tr>
</thead>
<tbody>
<tr>
<td>I</td>
<td>{(V_{(1,0,0)}, V_{(1,1,0)}), (V_{(0,1,0)}), (V_{(1,0,1)})}</td>
</tr>
<tr>
<td>II</td>
<td>{(V_{(1,0,0)}, V_{(1,1,0)}), (V_{(0,1,1)}), (V_{(1,0,0)})}</td>
</tr>
<tr>
<td>III</td>
<td>{(V_{(0,1,0)}, V_{(0,1,1)}), (V_{(1,0,0)}), (V_{(0,1,0)})}</td>
</tr>
<tr>
<td>IV</td>
<td>{(V_{(0,1,1)}, V_{(0,1,0)}), (V_{(1,0,1)}), (V_{(0,1,0)})}</td>
</tr>
<tr>
<td>V</td>
<td>{(V_{(0,0,1)}, V_{(1,0,1)}), (V_{(0,0,0)}), (V_{(0,0,0)})}</td>
</tr>
<tr>
<td>VI</td>
<td>{(V_{(1,0,0)}, V_{(1,0,0)}), (V_{(0,1,0)}), (V_{(1,0,0)})}</td>
</tr>
</tbody>
</table>

C. **HSVPWMS III**

The voltage reference vector passing through the odd–even triangles can be synthesized by using the two adjacent voltage vectors and the two voltage vectors of the far states to match the output and reference volt-seconds, as shown in Fig. 6. Table IV shows the vector combinations used in the three segments of each sector for HSVPWMS III.

D. **HSVPWMS IV**

The voltage reference vector passing through the odd–even triangles can be synthesized by using the two adjacent voltage vectors and one voltage vector of either far state to match the
TABLE IV
VECTOR COMBINATIONS USED IN THE THREE SEGMENTS OF EACH SECTOR USING HSVPWMS III

<table>
<thead>
<tr>
<th>Sector</th>
<th>Odd-even triangle</th>
</tr>
</thead>
<tbody>
<tr>
<td>I</td>
<td>{V_{(0,0,1)}, V_{(1,0,1)}, V_{(1,0,1)}, V_{(0,1,1)}}</td>
</tr>
<tr>
<td>II</td>
<td>{V_{(1,0,0)}, V_{(1,0,1)}, V_{(1,0,1)}, V_{(0,1,1)}}</td>
</tr>
<tr>
<td>III</td>
<td>{V_{(0,0,1)}, V_{(0,1,1)}, V_{(0,1,1)}, V_{(1,0,1)}}</td>
</tr>
<tr>
<td>IV</td>
<td>{V_{(1,1,0)}, V_{(1,1,0)}, V_{(0,0,1)}, V_{(1,0,1)}}</td>
</tr>
<tr>
<td>V</td>
<td>{V_{(0,1,0)}, V_{(0,1,0)}, V_{(0,1,1)}, V_{(1,0,1)}}</td>
</tr>
<tr>
<td>VI</td>
<td>{V_{(0,1,1)}, V_{(0,1,1)}, V_{(0,1,1)}, V_{(1,0,1)}}</td>
</tr>
</tbody>
</table>

IV. DEADTIME EFFECT AND NARROW CMV PULSE AVOIDANCE

The implementation of only odd or only even vectors requires simultaneous switching of two converter legs. In practice, this could be challenging because of the deadtimes and the differences in the gate delay [30]. The deadtime mismatch causes a narrow pulse in the CMV. The narrow pulse frequency increases as the switching frequency increases. To further assess the situation, consider Fig. 8. During the deadtimes, the resulting CMV is no longer a function of the switching commands. Instead it is strictly dependent on the signs of the three-phase load currents. In this case, the CMV \( \in \{V_{dc}, V_{dc}, -V_{dc}, -V_{dc}\} \). Therefore, the effective switching state during the deadtimes must be analyzed for each transition.

Consider the following three examples.

**Example 1:** Suppose the transition from the odd vector \( V_{(1,0,0)} \) to the odd vector \( V_{(0,1,1)} \) is taking place with the three-phase load currents \( i_a > 0, i_b < 0, \) and \( i_c > 0 \). The effective switching state during the deadtime \( t_{d13} \) will be the same as the destination vector state \( (0, 1, 0) \). This is because of the current polarities, as shown in Fig. 9: \( i_a \) is positive and will flow through the freewheeling diode of \( S_1 \) when \( S_1 \) turns OFF; \( i_b \) is negative and will flow through the freewheeling diode of \( S_2 \); and \( i_c \) is positive and will flow through the freewheeling diode...
During $t_{d13}$ and $i_a > 0$, $i_b > 0$, and $i_c < 0$, the CMV is equal to $-V_{dc}$.

During $t_{d13}$ and $i_a < 0$, $i_b < 0$, and $i_c > 0$, the CMV is equal to $V_{dc}$.

There are six regions in which the three-phase load currents do not change sign.

Example 2: If the same transition occurs with the three-phase load currents $i_a > 0$, $i_b > 0$, and $i_c < 0$, the effective switching state during the deadtime $t_{d13}$ will not be the same as the destination vector state $(0, 1, 0)$, and the resulting CMV during $t_{d13}$ will be equal to $-V_{dc}$, as shown in Fig. 10.

Example 3: If the same transition occurs with the three-phase load currents $i_a < 0$, $i_b < 0$, and $i_c > 0$, the effective switching state at the deadtime $t_{d13}$ will not be the same as the destination vector state $(0, 1, 0)$. Instead it will be $(1, 1, 0)$, and the resulting CMV will be $V_{dc}$, as shown in Fig. 11.

By considering all of the possible intervals in which the three-phase load currents do not change sign, six current sectors can be defined, as shown in Fig. 12. The calculations for the effective switching states during the other deadtimes are performed in the same manner as illustrated in the examples and listed in Tables VII and VIII for odd and even states, respectively. It is important to note that the effective switching state is the same as the effective switching state when the transition is reversed ($t_{dxy} = t_{dyx}$). For this reason, the calculations for this type of transition are not included in the tables.

It is evident from Tables VII and VIII that the CMV is not limited to $|V_{dc}|$ even if only odd or only even states are used. Therefore, to avoid the unwanted CMV peaks ($|V_{dc}|$), the control algorithm must be modified to use only odd states in S-II, S-IV, and S-VI and only even states in S-I, S-III, and S-V. The current sectors are shown in Fig. 13(a).

It is important to note the connection between the current sectors {$S-I$–$S-VI$} and the power factor (PF). In machine drive applications the current space vector ($i$) is lagging. This causes the current sectors to rotate clockwise and changes the feasible regions of using only odd or only even states. Fig. 13(b) shows the rotated current space vector when the PF = 0.866 lagging.

Modifying the control algorithm to avoid the peak CMV is not difficult as long as the load current signals are available in the loop. Fig. 14 shows the modified control block diagram.

The current sectors are determined by measuring the load currents and checking the sign of each phase current. Although this modification allows one to avoid the maximum CMV peaks, high $\frac{dv_{cm}}{dt}$ is still possible during some deadtimes. Furthermore, the availability of only odd or only even vectors is linked to the PF. When the PF = 1, the current sectors are completely out of sync. Whereas if the PF = 0.866, the current sectors are completely in sync. To avoid this limitation, the authors propose a new commutation method in the next section. The new commutation method allows a complete decoupling of the effect of PF, and it avoids the narrow CMV pulse during deadtime intervals.
V. PROPOSED COMMUTATION ALGORITHM

As shown in the previous section, by including the load current signals in the control loop, one can determine when it is safe to use only odd or only even vectors without causing peak CMV. However, there are unwanted transitions in the CMV from \( \frac{V_{dc}}{3} \) to \( -\frac{V_{dc}}{3} \) or vice versa during some deadtime intervals. As Tables VII and VIII demonstrate, in each current sector there are two deadtime intervals in which no change in the CMV occurs. These two deadtimes can be utilized to make a proper commutation sequence among the odd vectors or even vectors. Fig. 15 shows the commutation sequences that achieve no change in the CMV during the deadtime intervals. Furthermore, the rotation of the current sectors due to the PF will not affect the availability of only odd or only even vectors because in each current sector there are two safe sequences: one for only odd states and one for only even states. The safe sequences also allow us to equalize switching losses per fundamental cycle among the IGBTs.
Therefore, there is no unbalanced temperature distribution in inverter switches.

VI. PERFORMANCE ANALYSIS

HSVPWMS methods utilize different voltage vectors with different possible sequences. Each synthesis and combination corresponds to a different number of commutations, output current quality, CMV characteristics, dc-link current harmonics, and switching losses. Therefore, to understand the full potential of each method, a fair comparison with the CMV reduction PWM methods in terms of CMV characteristics, HDF, dc-link current, and switching losses is presented next.

A. CMV Characteristics

Common-mode behavior for induction machines is mostly capacitive [41]. Therefore, it does not require an extensive high-frequency machine modeling to determine the relationship between CMV characteristics and the resulting CMC. Early studies showed that the parasitic machine capacitances can be measured [42] or calculated by finite element methods [43], [44]. A machine model for high frequencies is developed in [45] and shows that the CMC is directly proportional to the frequency of the CMV.

Fig. 16 shows the CMV and its spectrum for the conventional SVPWM. The amplitude of the CMV is \( V_{dc} \) and is high due to the use of zero vectors. Another fundamental component of the CMV at the PWM carrier frequency \( f_c \).

All of the other HSVPWMS methods have similar CMV waveforms for \( MI \leq \frac{4}{3\sqrt{3}} \), where the main harmonic components are at \( 3f_1 \) and \( 6f_1 \), where \( f_1 \) is the system fundamental frequency (for instance 60 Hz). These two frequency components result in a very small CMC because the impedance is very high prior to the resonance region [41]. HSVPWMS I and HSVPWMS III have harmonic components around the carrier frequency \( f_c \) with an amplitude lower than that of the conventional SVPWM method. The \( f_c \) component increases as the MI increases, as shown in Figs. 17 and 19. HSVPWMS II and HSVPWMS IV have high-frequency CMV components around \( 2f_c \) that increase as the MI increases. Most of the CMV harmonic components are concentrated in the fundamental component \( 3f_1 \) and baseband harmonic \( 6f_1 \), as shown in Figs. 18 and 20. Generally, HSVPWMS methods have a maximum CMV amplitude equal to \( \frac{V_{dc}}{3} \), which is three times less than the regular SVPWM. More importantly, the amplitudes of CMV at high frequencies are reduced tremendously, especially at \( MI \leq M_{I_{\text{max}}} \). Therefore, the resulting CMC in the HSVPWMS method is significantly reduced as compared against the other methods.

Fig. 21 gives a clear representation of how the fundamental component of the CMV is behaving versus MI within the linear modulation region. The simulation parameters are: dc-link voltage \( 60 \) V, carrier frequency \( f_c = 5 \) kHz, and system fundamental frequency \( f_1 = 60 \) Hz. It has been observed through multiple simulations that the characteristic of each harmonic is relatively constant to those parameters. Therefore, the plots are given in terms of the dc-link voltage, fundamental frequency, and the carrier frequency. The high-frequency CMV of the regular
SVPWM has a very high amplitude when the MI is low because of the high duty ratio of the zero vector. The suitable HSVPWMS can be selected based on the level of the MI and the characteristics of the machine common-mode impedance.

B. Harmonic Distortion Factor

In the motor winding $f_{sw}/f \geq 20$, the harmonic current is directly proportional to the harmonic flux. Therefore, the current quality can be studied by using only the information of the applied voltage vector independent of the load ratings. The concept of HDF introduced in [46] is further expanded to account for the transitions between different syntheses. The harmonic flux for the $N$th carrier cycle can be defined as [29]

$$\lambda_h(MI, \theta, V_{(\ldots)}) = \int_{NT_s}^{(N+1)T_s} (V_{(\ldots)} - V_{ref})dt \quad (24)$$

where $V_{(\ldots)}$ is the applied voltage vector during the sampling time $NT_s$. Normalizing the harmonic flux results in the har-
monic flux voltage vector

\[ \lambda_{hn} = \frac{\pi}{V_{dc}I_s} \lambda_h. \]  

(25)

Each voltage utilization method results in a unique harmonic voltage vector trajectory. The normalized harmonic flux rms value over a PWM (duty cycle \( \delta \) of 0 to 1) is calculated as follows:

\[ \lambda_{hn-rms}(\text{MI}, \theta) = K_f^2 \sqrt{\int_0^1 \lambda_{hn}^2 d\delta} \]  

(26)

where \( K_f \) is the ratio of the switching per PWM cycle between the regular SVPWM and the HSVPWMS methods. In order to attain a fair comparison between HSVPWMS methods and the regular SVPWM method, all HSVPWMS methods are set to operate at the same switching frequency by dividing the carrier frequency of each method by \( K_f \). Taking the average value of the rms harmonic flux vector over a fundamental cycle results in the HDF, which is a measure of the ac current ripple

\[ \text{HDF} = \frac{288}{\pi^2} \frac{1}{2\pi} \int_0^{2\pi} \lambda_{hn-rms} d\theta. \]  

(27)

The hybrid harmonic distortion factor HDF\(_H\) is obtained by averaging the HDFs of each sequence based on the value of the modulation index as follows:

\[ \text{HDF}_H = \begin{cases} \text{HDF}_{(O \cup E)} & \text{MI} \leq \frac{1}{3\sqrt{3}} \\ M_p \text{HDF}_{(O \cup E)} + (1 - M_p) \text{HDF}_{(E \cup O)} & \text{MI} > \frac{1}{3\sqrt{3}} \end{cases} \]  

(28)

where \( M_p \) is the modulation parametric factor defined as

\[ M_p = \frac{\text{MI}_{\text{max}} - \text{MI}}{\frac{2}{3\sqrt{3}}}. \]  

(29)

\( \text{HDF}_{(O \cup E)} \) is HDF calculated when only the odd vectors \( \{V_{(1,0,0)}, V_{(0,1,0)}, V_{(0,0,1)}\} \) are applied or only the even vectors \( \{V_{(1,0,0)}, V_{(0,1,1)}, V_{(1,1,0)}\} \) are applied, depending on the location of the segments as described previously for each HSVPWMS. \( \text{HDF}_{(O \cup E)} \) is HDF calculated when both odd and even vectors are applied \( \{V_{(1,0,0)}, V_{(1,1,0)}, V_{(0,1,0)}, V_{(0,1,1)}, V_{(0,0,1)}, V_{(1,0,1)}\} \), depending on the selected group of vectors. When the MI is less than or equal to \( \frac{1}{3\sqrt{3}} \), the HDF\(_H\) is identical to the HDF resulting from only odd vectors or only even vectors. When the MI is greater than \( \frac{1}{3\sqrt{3}} \), the HDF\(_H\) is averaged based on the definition in (28). Fig. 22 shows the numerical results of HDF\(_H\) for all HSVPWMS along with the regular SVPWM.

All the HSVPWMS methods present similar HDF in the region where \( \text{MI} \leq \frac{1}{3\sqrt{3}} \). Although the HDF for HSVPWMS II is lower than the other hybrid methods, it presents higher CMV frequency as will be seen later in the results.

### C. DC-Link Current Harmonics

The rms current on the dc side is important factor in determining the dc-link capacitor sizing [47]. To compare the behavior of the PWM methods, \( K_{\text{dc}} \) is defined in [29] as

\[ K_{\text{dc}} = \frac{I_{m-h-rms}^2}{I_{1-rms}^2}, \]  

(30)

where \( I_{m-h-rms} \) is the rms value of the dc-link current and \( I_{1-rms} \) is the fundamental component of the output rms current. The calculations for \( K_{\text{dc}} \) are performed numerically for different power factors. Fig. 23(a)–(c) show \( K_{\text{dc}} \) for the HSVPWMS methods compared with the conventional SVPWM and CMV reduction methods. When the PF is high, \( K_{\text{dc}} \) for the hybrid methods is higher than the other methods. However, when operating at low PF, \( K_{\text{dc}} \) for the hybrid methods surpasses \( K_{\text{dc}} \) for all of the other CMV reduction methods. In general, \( K_{\text{dc}} \) values for all the hybrid methods are almost identical and decrease with decreasing the power factor.

### D. Switching Losses

The switching losses for the VSI using HSVPWMS are calculated analytically by taking into account the effect of changing different vector syntheses over the switching frequency. The IGBT current and voltage do not abruptly change their values. Instead there is a transition interval in which both the current and voltage across the IGBT have values larger than zero [48], [50]. The average value of the switching losses can be
calculated as follows:

\[
P_{\text{sw-IGBT}} = f_s(MI) \left[ \int_{t}^{t+\tau_r} i_C(t)v_{CE}(t)dt + \int_{t}^{t+\tau_f} i_C(t)v_{CE}(t)dt \right]
\]

where \( f_s \) is the effective switching frequency that represents the number of switching instants for one IGBT in one fundamental frequency period, unlike the regular SVPWM. \( f_s \) in HSVPWMS is a function of the modulation index, as shown in Table VI. \( i_C \) is the collector current, and \( v_{CE} \) is the voltage across the collector-emitter. \( \tau_r \) and \( \tau_f \) are the rising and falling times, respectively. The integration explained in Appendix A results in the following average switching losses per IGBT:

\[
P_{\text{sw-IGBT}} = f_s(MI) \left[ \frac{1}{3}I_L V_{dc} \tau_r + \frac{9}{50}I_L V_{dc} \tau_f \right].
\]

The total switching losses also include the reverse recovery losses associated with the diode turn-off process

\[
P_{\text{rr-D}} = \frac{1}{2} f_s(MI) I_{\text{rr}} V_{dc} \tau_{rr}
\]

where \( I_{\text{rr}} \) is the maximum value of the reverse recovery current, and \( \tau_{rr} \) is the reverse recovery time.

Table VI

<table>
<thead>
<tr>
<th>Modulation method</th>
<th>( f_s(MI)_{MI&lt;0.5} )</th>
<th>( f_s(MI)_{MI&gt;0.5} )</th>
</tr>
</thead>
<tbody>
<tr>
<td>SVPWM</td>
<td>1f_s</td>
<td>1f_s</td>
</tr>
<tr>
<td>HSVPWM 1</td>
<td>4f_s</td>
<td>1f_s + (1-MI)</td>
</tr>
<tr>
<td>HSVPWM II</td>
<td>4f_s</td>
<td>1f_s + (1-MI)</td>
</tr>
<tr>
<td>HSVPWM III</td>
<td>4f_s</td>
<td>1f_s + (1-MI)</td>
</tr>
<tr>
<td>HSVPWM IV</td>
<td>4f_s</td>
<td>1f_s + (1-MI)</td>
</tr>
</tbody>
</table>

![Fig. 23](image1)

![Fig. 24](image2)

**VII. EXPERIMENTAL RESULTS**

The experimental work has been carried out using the following parameters and operating conditions: dc-link voltage 100 V, carrier frequency \( f_c = 5 \text{ kHz} \) and an RL load with 5 \( \Omega \) and 2 mH. The system consists of three inverter legs with each leg being realized by the IGBT module CM100DY-24A. The control algorithm is implemented using DSpace CP1103.

Fig. 25(a) shows the resulting CMV with MI = 0.75. The reference voltage vector is synthesized using only odd and only even states. This results in a square wave CMV with a frequency equal to \( 3f_1 \) (three times the fundamental frequency of the system 180 Hz). Fig. 25(b) shows the fast Fourier transform (FFT) for the CMV with MI = 0.75. All of the CMV is concentrated at 180 Hz. The CMV frequency spectrum is similar to the one in Fig. 25(b) at any MI \( \in \{0, 0.766\} \).
Fig. 25. Experimental results. (a) CMV at MI = 0.75 (1—HSVPWMS I, 2—HSVPWMS II, 3—HSVPWMS III, 4—HSVPWMS IV). (b) FFT of the CMV.

Fig. 26. Experimental results. (a) CMV waveforms at MI = 0.85 for 1—SVPWM, 2—HSVPWMS I, 3—HSVPWMS II, 4—HSVPWMS III, 5—HSVPWMS IV, 6—AZSPWM1-2, 7—NSPWM, 8—AZSPWM3, and 9—DPWM1. (b) FFT of the CMV.

Fig. 27. Experimental results. (a) CMV waveforms at MI = 1 for 1—SVPWM, 2—HSVPWMS I, 3—HSVPWMS II, 4—HSVPWMS III, 5—HSVPWMS IV, 6—AZSPWM1-2, 7—NSPWM, 8—AZSPWM3, and 9—DPWM1. (b) FFT of the CMV.

Fig. 26(a) shows the CMV with MI = 0.85 for the proposed hybrid approach and several PWM methods. For the MI > 0.766, the reference voltage vector starts entering the odd–even region, leading to a CMV that fluctuates between the values of $\frac{1}{3}V_{dc}$ and $-\frac{1}{3}V_{dc}$. However, the CMV spectrum is still concentrated at 180 Hz frequency, as shown in Fig. 26(b). The frequency spectrum for the conventional SVPWM and DPWM1 includes high amplitude CMV due to the use of the zero vector. This high amplitude is located at the carrier frequency (normally in kHz). The conventional CMV reduction PWM methods have lower CMV amplitudes, but they are still located at the carrier frequency. The main difference between the proposed method and the conventional CMV reduction methods is that the CMV frequency is decreased. A lower frequency CMV has two main advantages over a high-frequency CMV. First, less CMV fluctuation corresponds to less EDM in the bearings. Second, a low-frequency CMV can help meet the EMC standards because the amplitude limit for low frequencies is much higher than the amplitude limit for high frequencies.
Fig. 28. Experimental results of the load current $i_a(t)$. (a) $MI = 0.85$. (b) $MI = 1$ for 1—SVPWM+20A, 2—HSVPWMS I+15A, 3—HSVPWMS II+10A, 4—HSVPWMS III+5A, 5—HSVPWMS IV, 6—AZSPWM1-2-5A, 7—NSPWM-10A, 8—AZSPWM3-15A, and 9—DPWM1-20A.

Fig. 29. Experimental results. (a) HSVPWMS II without implementing the safe commutation algorithm. During the deadtimes the effective switching state could be one of the zero vectors $V_{(0,0,0)}$ or $V_{(1,1,1)}$ and the CMV will be similar to type 1, or it could be similar to type 2 when the effective switching state result in one of the odd or even vectors. (b) HSVPWMS II with implementing the safe commutation algorithm, both types 1 and 2 unwanted transitions are successfully avoided. $MI = 0.85$.

Fig. 27(a) shows the CMV at $MI = 1$ for the proposed hybrid approach and several PWM methods. As the MI increases higher the implementation of only odd or only even synthesis decreases leading to high-frequency CMV frequency components. However, the high-frequency CMV components are still less than the ones resulting from the conventional CMV reduction methods, as shown in Fig. 27(b).

Fig. 28(a) and (b) show the load currents for the proposed methods along with the conventional PWM methods. Although the load is mostly resistive with $PF = 0.988$, current ripples resulting from the proposed hybrid methods are still comparable with the current ripples resulting from the conventional PWM methods. As the MI increases, the HDF of load currents for the proposed hybrid method improves because the reference voltage vector synthesis shifts from only odd and only even to odd–even.

Fig. 30. Switching ON and OFF events for an IGBT.

Fig. 29(a) shows the CMV waveform without implementing the safe commutation algorithm. There are two types of un-
TABLE VII
EFFECTIVE SWITCHING STATES IN CASE OF ONLY ODD MODULATION

<table>
<thead>
<tr>
<th>Sector</th>
<th>Deadtime</th>
<th>Eff. state (S1,S2,S3)</th>
<th>( V_{cm} )</th>
</tr>
</thead>
<tbody>
<tr>
<td>S-I</td>
<td>( t_{d13} ) (0.0,0)</td>
<td>-( V_{dc} )</td>
<td></td>
</tr>
<tr>
<td></td>
<td>( t_{d35} ) (1.0,0)</td>
<td>-( V_{dc} )</td>
<td></td>
</tr>
<tr>
<td></td>
<td>( t_{d51} ) (0.1)</td>
<td>-( V_{dc} )</td>
<td></td>
</tr>
<tr>
<td>S-II</td>
<td>( t_{d13} ) (0.0,1)</td>
<td>-( V_{dc} )</td>
<td></td>
</tr>
<tr>
<td></td>
<td>( t_{d35} ) (1.0,0)</td>
<td>-( V_{dc} )</td>
<td></td>
</tr>
<tr>
<td></td>
<td>( t_{d51} ) (0.1,1)</td>
<td>-( V_{dc} )</td>
<td></td>
</tr>
<tr>
<td>S-III</td>
<td>( t_{d13} ) (0.0,1)</td>
<td>-( V_{dc} )</td>
<td></td>
</tr>
<tr>
<td></td>
<td>( t_{d35} ) (1.0,0)</td>
<td>-( V_{dc} )</td>
<td></td>
</tr>
<tr>
<td></td>
<td>( t_{d51} ) (0.1,1)</td>
<td>-( V_{dc} )</td>
<td></td>
</tr>
<tr>
<td>S-IV</td>
<td>( t_{d13} ) (1.0,1)</td>
<td>-( V_{dc} )</td>
<td></td>
</tr>
<tr>
<td></td>
<td>( t_{d35} ) (1.0,0)</td>
<td>-( V_{dc} )</td>
<td></td>
</tr>
<tr>
<td></td>
<td>( t_{d51} ) (0.1)</td>
<td>-( V_{dc} )</td>
<td></td>
</tr>
<tr>
<td>S-V</td>
<td>( t_{d13} ) (1.0,0)</td>
<td>-( V_{dc} )</td>
<td></td>
</tr>
<tr>
<td></td>
<td>( t_{d35} ) (1.0,0)</td>
<td>-( V_{dc} )</td>
<td></td>
</tr>
<tr>
<td></td>
<td>( t_{d51} ) (0.0,0)</td>
<td>-( V_{dc} )</td>
<td></td>
</tr>
<tr>
<td>S-VI</td>
<td>( t_{d13} ) (1.0,0)</td>
<td>-( V_{dc} )</td>
<td></td>
</tr>
<tr>
<td></td>
<td>( t_{d35} ) (1.1,0)</td>
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<td></td>
</tr>
<tr>
<td></td>
<td>( t_{d51} ) (0.1,0)</td>
<td>-( V_{dc} )</td>
<td></td>
</tr>
</tbody>
</table>

TABLE VIII
EFFECTIVE STATES IN CASE OF ONLY EVEN MODULATION

<table>
<thead>
<tr>
<th>Sector</th>
<th>Deadtime</th>
<th>Eff. state (S1,S2,S3)</th>
<th>( V_{cm} )</th>
</tr>
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<tr>
<td>S-I</td>
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<td></td>
<td>( t_{d46} ) (0.1,0)</td>
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<tr>
<td></td>
<td>( t_{d62} ) (0.1,1)</td>
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<td></td>
</tr>
<tr>
<td>S-II</td>
<td>( t_{d24} ) (1.1,1)</td>
<td>( V_{dc} )</td>
<td></td>
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<tr>
<td></td>
<td>( t_{d46} ) (0.1,1)</td>
<td>( V_{dc} )</td>
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</tr>
<tr>
<td></td>
<td>( t_{d62} ) (0.1,1)</td>
<td>( V_{dc} )</td>
<td></td>
</tr>
<tr>
<td>S-III</td>
<td>( t_{d24} ) (1.0,1)</td>
<td>( V_{dc} )</td>
<td></td>
</tr>
<tr>
<td></td>
<td>( t_{d46} ) (0.1,1)</td>
<td>( V_{dc} )</td>
<td></td>
</tr>
<tr>
<td></td>
<td>( t_{d62} ) (0.0,1)</td>
<td>-( V_{dc} )</td>
<td></td>
</tr>
<tr>
<td>S-IV</td>
<td>( t_{d24} ) (1.0,1)</td>
<td>( V_{dc} )</td>
<td></td>
</tr>
<tr>
<td></td>
<td>( t_{d46} ) (1.1,1)</td>
<td>( V_{dc} )</td>
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<tr>
<td></td>
<td>( t_{d62} ) (1.0,1)</td>
<td>( V_{dc} )</td>
<td></td>
</tr>
<tr>
<td>S-V</td>
<td>( t_{d24} ) (1.0,0)</td>
<td>-( V_{dc} )</td>
<td></td>
</tr>
<tr>
<td></td>
<td>( t_{d46} ) (1.0,0)</td>
<td>( V_{dc} )</td>
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<tr>
<td></td>
<td>( t_{d62} ) (1.0,1)</td>
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<td>( V_{dc} )</td>
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<tr>
<td></td>
<td>( t_{d62} ) (1.1,1)</td>
<td>( V_{dc} )</td>
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</tr>
</tbody>
</table>

VIII. CONCLUSION

In this paper, a set of new HSVPWMS control methods are proposed to minimize the amplitude and frequency of the CMV. A detailed comparison shows the relative performance of the conventional PWM methods and the proposed HSVPWMS methods in terms of CMV characteristics, HDF, switching losses, and dc ripples. The conventional PWM methods have a CMV harmonic component with high frequency equal to the carrier frequency \( f_c \). The new HSVPWMS method proposed in the paper limits the CMV amplitude to \( V_{dc} / 3 \) and concentrates the CMV harmonics to a low frequency equal to three times the fundamental frequency \( 3f_1 \). A new commutation algorithm that avoids nonlinear behavior during deadtime intervals is proposed. The performance of the proposed HSVPWMS algorithms has been verified via simulations and validated by laboratory experiments. The new HSVPWMS method can potentially enable converters to operate at a very high switching frequency using the newly emerging wide bandgap devices while effectively mitigating the adverse CMV effects associated with conventional high-frequency PWM schemes.

APPENDIX A

TABLES

Tables VII–VIII show the effective switching states in case of only odd and only even syntheses deadtimes.

APPENDIX B

AVERAGE LOSSES IN TURN ON AND OFF EVENTS

Fig. 30 shows a typical turn-on and turn-off waveforms characteristic for an IGBT switch.

Equation (31) can be expanded by simple linear approximation of \( i_C \) and \( v_{CE} \), except during turn-off events, in which case \( i_C \) can be approximated as an exponentially decaying function as follows [48]:

\[
i_C(t) = I_L e^{-a(t-t_i)}
\]

where \( t_i \) is the initial value of time in which the turn-off event started and \( a \) is a constant that can be adjusted to achieve the best function approximation. To solve (34), it is assumed that at \( t = t_i + t_f \), 95% of the collector current \( i_C \) vanishes. Therefore, \( a \) can be defined as a function of the falling time \( t_f \) as follows:

\[
a = \frac{3}{t_f}
\]

Therefore, the integral becomes

\[
P_{sw-IGBT} = f_s(MI) \left[ \int_0^{t_f} \frac{0}{t_r} i_C \left( 2V_{dc} - \frac{t}{t_r} 2V_{dc} \right) dt + \int_0^{t_f} I_L e^{-a t} \frac{t}{t_f} 2V_{dc} dt \right]
\]

\[
P_{sw-IGBT} = f_s(MI) \left[ \frac{1}{3} I_L V_{dc} t_r + \frac{2I_L V_{dc}}{t_f} \left( 1 - e^{-at_f} (at_f + 1) \right) \right].
\]
Substituting (35) in (37) leads to

$$P_{sw-IGBT} = f_s(M) \left[ \frac{1}{3} I_L V_{dc} t_r + \frac{9}{50} I_L V_{dc} t_f \right].$$  \hspace{1cm} (38)

REFERENCES


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