CHAPTER 3

ROTATING SPACE VECTOR MODULATION TECHNIQUE FOR MATRIX CONVERTERS TO ELIMINATE THE COMMON MODE VOLTAGE IN INDUCTION MACHINES

3.1 INTRODUCTION

The conventional PWM voltage applied to an induction machine terminal results in high frequency Common Mode Voltage (CMV). The CMV applied to the machine by the converters creates common-mode currents. Common-mode currents have the potential to cause physical damage or unwanted tripping of ground fault relays in motor drives and electrical networks. In addition, research has identified damages such as frosting; spark tracks in surface of balls, races and pitting of electric machines caused by bearing currents that flow due to the common mode voltage (Chen et al 1996).

In a typical three-phase AC to AC drive, there exists a substantial common-mode voltage between the load neutral and the ground due to the conventional PWM technique (Erdman et al 1996). As the modulation frequency increases and the zero-sequence impedance of the machine decreases, the CMV causes higher common-mode currents, worsening Electromagnetic Interference (EMI) problems and potentially damaging the network or the machine. Most of the common mode voltage appears across the input transformer insulation leading to a higher transformer insulation requirement (Gupta et al 2010).

Matrix converters have the potential to eliminate the common mode voltages,
which are caused by the conventional PWM techniques, by the application of Zero Common Mode Voltage Vectors (ZCMVV).

In this chapter, a control strategy termed as the Rotating Space Vector Modulation (RSVM) technique that uses ZCMVV is proposed, for a direct AC-to-AC converter fed induction machine. A control procedure that provides input current control in the RSVM technique is also proposed.

3.2 PROBLEM FORMULATION

The aim of this chapter is to design a PWM (RSVM) technique that uses the Zero Common Mode Voltage Vectors (ZCMVV) for elimination of the CMV in a matrix converter. The procedure to use ZCMVV for the input current control is also established. The proposed RSVM technique eliminates CMV content especially at low output voltage ranges. A modified CMC topology termed as the Phase Shifted Dual Source Matrix Converter (PSDSMC) is proposed, to increase the voltage transfer ratio to 0.866, with the modified RSVM technique. The performance of this scheme is evaluated through simulation in MATLAB-Simulink. The approach involves deriving mathematical relations for the proposed techniques. The advantages and disadvantages of this methodology are also highlighted in this chapter.

3.3 EXISTING TECHNIQUES

Different CMV reduction techniques such as PWM based (Lee and Sul 1999 and Videt et al 2007), active filter and passive filter based current injection for cancellation of the common mode (Rendusara and Enjeti 1998 and Ogasawara et al 1998) have been proposed for three-phase PWM inverters. Using the optimal zero vector selection to reduce the maximum peak of the common mode voltage by 34%, in the CMC and the IMC fed drives, have been proposed in Cha and Enjeti (2003), Nguyen and Hong-Hee Lee (2012).
A space vector based scheme to reduce the CMV in the cascaded multilevel inverters, has been proposed in Gupta and Khambadkone (2007). Kanchan et al (2006) have proposed the elimination of the CMV for open-ended winding based induction motor drives fed by a three-level inverter. Gupta et al (2010) have proposed the same for a matrix converter. However, methods to eliminate the common mode voltage in drives fed by a CMC without an open-ended winding are not available in the literature.

3.4 COMMON MODE VOLTAGE EFFECTS IN THE INDUCTION MOTOR

Figure 3.1 shows a power converter connected to an induction machine. Due to the stray impedance $Z_s$ between the motor neutral and the ground, the presence of the common mode voltage $V_{cm}$ at the motor neutral point contributes to a high frequency leakage current $I_{cm}$. Equations (3.1) to (3.3) give the common mode voltage $V_{cm}$ at the motor neutral point.

![Figure 3.1 Leakage current paths in an induction machine](image-url)
\[ V_a - V_{cm} = RI_a + L \frac{dl_a}{dt} \]  
(3.1)

\[ V_b - V_{cm} = RI_b + L \frac{dl_b}{dt} \]  
(3.2)

\[ V_c - V_{cm} = RI_c + L \frac{dl_c}{dt} \]  
(3.3)

Assuming that the stray impedance \( Z_s \) is high results in \( I_{cm} \approx 0 \), which results in \( I_a + I_b + I_c \approx 0 \). Adding Equations (3.1) to (3.3), we get \( V_{cm} \), as given by Equation (3.4).

\[ V_{cm} = (V_a + V_b + V_c) / 3 \]  
(3.4)

3.5 MATHEMATICAL CONCEPTS FOR THE PROPOSED RSVM TECHNIQUE

The CMC topology, shown in Figure 1.2, has been chosen in this work, as it is the only possible converter topology where the CMV can be eliminated. Equation (3.5) gives the expression for the source neutral voltage \( V_N \) that is zero for any three-phase three-wire balanced system.

\[ V_N = (V_A + V_B + V_C) / 3 \]  
(3.5)

From Equations (3.4) and (3.5), the sum of load voltages \( V_a \), \( V_b \) and \( V_c \) should be equal to the sum of input phase voltages \( V_A \), \( V_B \) and \( V_C \) for the load neutral voltage \( V_n \) to be equal to \( V_N \) i.e. zero. Hence, the elimination of the CMV introduces a third constraint in addition to the two constraints explained in Equation (1.2), namely that at any given instant, all the three input phases must be connected to the load. This additional constraint, used for the elimination of the CMV, allows the use of only six switching vectors that connects all the input phases to the output phases, as shown in Figures 3.2(a) and 3.2(b), from the valid 27 vectors of the CMC. The remaining vectors that are not used are the 18 stationary vectors and the three
zero vectors, as shown in Figure 3.2(c). These six vectors used are termed as the rotating space vectors since their positions in the space are not fixed. They are also called as ZCMVV since they produce zero CMV. Among these six rotating space vectors, three space vectors rotate in the direction of the output frame and the remaining three space vectors rotate in the direction opposite to the output reference frame.

![Diagram of rotating vectors](image)

**Figure 3.2** (a) Clockwise rotating vectors, (b) counter-clockwise rotating vectors and (c) stationary vectors and zero vectors

Equation (3.6) gives the switching patterns for the counter-clockwise rotating vectors.

\[
\begin{bmatrix}
S_{Aa} & S_{Ab} & S_{Ac} \\
S_{Ba} & S_{Bb} & S_{Bc} \\
S_{Ca} & S_{Cb} & S_{Cc}
\end{bmatrix}
= \begin{bmatrix}
1 & 0 & 0 \\
0 & 1 & 0 \\
0 & 0 & 1
\end{bmatrix}
\text{ or }
\begin{bmatrix}
0 & 0 & 1 \\
1 & 0 & 0 \\
0 & 1 & 0
\end{bmatrix}
\text{ or }
\begin{bmatrix}
0 & 1 & 0 \\
0 & 0 & 1 \\
1 & 0 & 0
\end{bmatrix}
\] (3.6)
Figure 3.3(a) shows the output space vectors, when any of the three possible switching patterns, given in Equation (3.6), are employed.

Switching space vectors rotate at input frequency $\omega_s$ rad/s while the output reference space vector rotates at $\omega_o$ rad/s. Determination of the duty cycles for the switching space vectors that rotate is tedious. Hence, the switching vectors rotating at $\omega_o$ are taken as the reference. The relative angle $\theta_v$ between the output reference voltage vector angle $\theta_o$ and the actual switching voltage space vector angle $\theta_s$ is computed and used for duty cycle calculations. Figure 3.3(b) illustrates the same. Equations (3.7) to (3.10) give the positions of the active switching voltage vectors and the reference output voltage vector in space

$$v_{abc}^+ = \frac{3}{2} V_i e^{j\omega_s t}$$  
(3.7)

$$v_{cab}^+ = \frac{3}{2} V_i e^{j(\omega_s t + \frac{2\pi}{3})}$$  
(3.8)

$$v_{bca}^+ = \frac{3}{2} V_i e^{j(\omega_s t - \frac{2\pi}{3})}$$  
(3.9)

$$v_o = \frac{3}{2} V_o e^{j\omega_o t}$$  
(3.10)

where, $v_{abc}^+, v_{cab}^+, v_{bca}^+$ are the active vectors with magnitude $V_i$ and $v_o$ is the reference output vector with magnitude $V_o$.

![Figure 3.3](image_url)  
Figure 3.3 Dynamic space vector PWM (a) positive sequence rotating reference frame and (b) fixed reference
Equations (3.11) to (3.14) give the duty cycles of the active vectors and the zero vectors obtained by the sine law of triangles, as shown in Figure 3.4.

\[
\frac{d_\alpha v_\alpha}{\sin(120^\circ - \theta_\alpha)} = \frac{d_\beta v_\beta}{\sin(\theta_\beta)} = \frac{V_{\alpha(REF)}}{\sin(60^\circ)} \tag{3.11}
\]

\[d_\alpha = m_v \sin(120^\circ - \theta_\alpha) \tag{3.12}\]

\[d_\beta = m_v \sin(\theta_\beta) \tag{3.13}\]

\[d_0 = 1 - d_\alpha - d_\beta \tag{3.14}\]

Figure 3.4  Duty cycle calculation (a) sector 1 and (b) weighted combination of the active vectors

The conventional SVM technique utilizes zero vectors for the sinusoidal output but in the proposed RSVM technique, the use of the zero vectors introduces common mode voltage. Equations (3.7) to (3.9) show that \(d_m v_{abc}^+ + d_m v_{cab}^+ + d_m v_{bca}^+ = 0\), where \(d_m\) is some arbitrary duty cycle. This idea is used to implement the zero vectors using the active rotating vectors. During the switching time of the zero vector, the RSVM technique uses all the
three active vectors with equal duty ratios. Equations (3.15) to (3.17) give the modified duty ratios for the RSVM technique.

\[ d_1^+ = d_1' = m_v \sin(120^\circ - \theta_v) + \frac{d_0}{3} \]  
\[ d_2^+ = d_2' = m_v \sin(\theta_v) + \frac{d_0}{3} \]  
\[ d_3^+ = d_0' = \frac{d_0}{3} \]  

Hence, within any given sector, the required output can be synthesized over a sampling period by applying the respective active rotating vector, as given in Table 3.1. Equation (3.18) gives the relation between the output voltage \( v_o \) and the active vectors, whose duty ratios satisfy Equation (3.19) at all times

\[ v_o = d_1^+ v_1 + d_2^+ v_2 + d_3^+ v_3 \]  
\[ d_1^+ + d_2^+ + d_3^+ = 1 \]  

where, \( d_1^+ \), \( d_2^+ \), \( d_3^+ \) are the duty cycles and \( v_1 \), \( v_2 \), \( v_3 \) are the active positive rotating vectors respectively.

### Table 3.1 Positive rotating switching vectors and sector no. for the CMC

<table>
<thead>
<tr>
<th>S. No.</th>
<th>( \theta_v )</th>
<th>Sector No</th>
<th>Active vectors</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td></td>
<td></td>
<td>( v_1 ) ( v_2 ) ( v_3 )</td>
</tr>
<tr>
<td>1</td>
<td>( 0^\circ &lt; \theta_v \leq 120^\circ )</td>
<td>1</td>
<td>( v_{abc}^+ ) ( v_{cab}^+ ) ( v_{bca}^+ )</td>
</tr>
<tr>
<td>2</td>
<td>( 120^\circ &lt; \theta_v \leq 240^\circ )</td>
<td>2</td>
<td>( v_{cab}^+ ) ( v_{bca}^+ ) ( v_{abc}^+ )</td>
</tr>
<tr>
<td>3</td>
<td>( 240^\circ &lt; \theta_v \leq 360^\circ )</td>
<td>3</td>
<td>( v_{bca}^+ ) ( v_{abc}^+ ) ( v_{cab}^+ )</td>
</tr>
</tbody>
</table>

Substituting Equations (3.7) to (3.10) in Equations (3.18) and (3.19), the duty cycle of each switch is obtained, as in Equations (3.20) to (3.22). However, the modulation index \( m_v \) of this method is limited to 0.5.
\[ D_{Aa} = D_{Bb} = D_{Cc} = \frac{1}{3} + \frac{2m}{3} \cos(\omega_a t - \omega_c t) \]  \hspace{1cm} (3.20)

\[ D_{Ba} = D_{Cb} = D_{Ac} = \frac{1}{3} + \frac{2m}{3} \cos((\omega_a t - \omega_c t) + \frac{2\pi}{3}) \]  \hspace{1cm} (3.21)

\[ D_{Ca} = D_{Ab} = D_{Bc} = \frac{1}{3} + \frac{2m}{3} \cos((\omega_a t - \omega_c t) - \frac{2\pi}{3}) \]  \hspace{1cm} (3.22)

With a similar procedure, the same output voltages can be synthesized using three negative rotating active space vectors whose duty ratios are given by Equations (3.15) to (3.17). Within any given sector, the required output can be synthesized by applying the respective active rotating vector, as given in Table 3.2.

**Table 3.2 Negative rotating switching vectors and sector no. for the CMC**

<table>
<thead>
<tr>
<th>S. No.</th>
<th>( \theta_v )</th>
<th>Sector No</th>
<th>Active vectors</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>( 0^\circ &lt; \theta_v \leq 120^\circ )</td>
<td>1</td>
<td>( v_{acb} ) ( v_{bac} ) ( v_{cba} )</td>
</tr>
<tr>
<td>2</td>
<td>( 120^\circ &lt; \theta_v \leq 240^\circ )</td>
<td>2</td>
<td>( v_{bac} ) ( v_{cba} ) ( v_{acb} )</td>
</tr>
<tr>
<td>3</td>
<td>( 240^\circ &lt; \theta_v \leq 360^\circ )</td>
<td>3</td>
<td>( v_{cba} ) ( v_{acb} ) ( v_{bac} )</td>
</tr>
</tbody>
</table>

### 3.6 INPUT POWER FACTOR CONTROL OF THE RSVM TECHNIQUE

Input power factor control in the direct AC-to-AC converters (Milanovic and Dobaj 2000) is carried out using the principle of the space vector technique applied to the input current but the RSVM technique uses the shared duty ratio control technique for achieving the same. At any instant, based on the applied active voltage space vector at the output, any of the input phase currents (say \( i_a(t) \)) should be equal to any one of the output phase currents (say \( i_A(t) \), \( i_B(t) \) or \( i_C(t) \)). Equations (3.33) to (3.36) give the positions
of the active output current vectors, when using the corresponding positive directional rotating voltage vectors and the input current vectors in space.

\[ i_{abc}^+ = \frac{3}{2} I_o e^{j(\omega t - \phi)} \]  \hspace{1cm} (3.33)

\[ i_{cab}^+ = \frac{3}{2} I_o e^{j(\omega t + \frac{2\pi}{3})} \]  \hspace{1cm} (3.34)

\[ i_{bca}^+ = \frac{3}{2} I_o e^{j(\omega t + \frac{2\pi}{3})} \]  \hspace{1cm} (3.35)

\[ i_s = \frac{3}{2} I_o e^{j(\omega t - \phi)} \]  \hspace{1cm} (3.36)

Equations (3.37) to (3.40) give the positions of the active output current vectors, when using the corresponding negative directional rotating voltage vectors and the input current vectors in space.

\[ i_{acb}^- = \frac{3}{2} I_o e^{-j(\omega t - \phi)} \]  \hspace{1cm} (3.37)

\[ i_{bac}^- = \frac{3}{2} I_o e^{-j(\omega t + \frac{2\pi}{3})} \]  \hspace{1cm} (3.38)

\[ i_{cba}^- = \frac{3}{2} I_o e^{-j(\omega t + \frac{2\pi}{3})} \]  \hspace{1cm} (3.39)

\[ i_s = \frac{3}{2} I_o e^{j(\omega t + \phi)} \]  \hspace{1cm} (3.40)

Equations (3.36) and (3.40) show that the input current lags or leads respectively the input voltage by \( \phi \) degrees when positive or negative directional voltage vectors are applied, where \( \cos \phi \) lagging is the output load power factor, as shown in Figure 3.5(a). Since the output power factor depends upon the load, it is not possible to control the output power factor. Hence, the only way to control the input power factor \( \cos \phi \) is to apply both +ve and -ve rotating voltage space vectors at the output terminals, as shown in Figures 3.5(b) to 3.5(d). Equation (3.50) gives the ratio \( r \) with which the +ve and -ve rotating voltage space vectors are applied to decide the input power
factor. However, the input power factor can be controlled in a limited range between the output power factor and unity i.e. $\cos \rho < \cos \phi < 1$. Equations (3.41) to (3.49) give the duty cycles of each switch.

$$D_{Aa} = \frac{1}{3} + \frac{2r m_v}{3} \cos(\omega_0 t - \omega_d t) + \frac{2(1-r) m_v}{3} \cos(\omega_d t + \omega_d t) \quad (3.41)$$

$$D_{Ba} = \frac{1}{3} + \frac{2r m_v}{3} \cos((\omega_0 t - \omega_d t) + \frac{2\pi}{3}) + \frac{2(1-r) m_v}{3} \cos((\omega_0 t + \omega_d t) - \frac{2\pi}{3}) \quad (3.42)$$

$$D_{Ca} = \frac{1}{3} + \frac{2r m_v}{3} \cos((\omega_0 t - \omega_d t) - \frac{\pi}{3}) + \frac{2(1-r) m_v}{3} \cos((\omega_0 t + \omega_d t) + \frac{\pi}{3}) \quad (3.43)$$

$$D_{Bb} = \frac{1}{3} + \frac{2r m_v}{3} \cos(\omega_0 t - \omega_d t) + \frac{2(1-r) m_v}{3} \cos((\omega_0 t + \omega_d t) + \frac{\pi}{3}) \quad (3.44)$$

$$D_{Cb} = \frac{1}{3} + \frac{2r m_v}{3} \cos((\omega_0 t - \omega_d t) - \frac{\pi}{3}) + \frac{2(1-r) m_v}{3} \cos((\omega_0 t + \omega_d t) + \frac{\pi}{3}) \quad (3.45)$$

$$D_{Ab} = \frac{1}{3} + \frac{2r m_v}{3} \cos((\omega_0 t - \omega_d t) + \frac{\pi}{3}) + \frac{2(1-r) m_v}{3} \cos((\omega_0 t + \omega_d t) - \frac{\pi}{3}) \quad (3.46)$$

$$D_{Cc} = \frac{1}{3} + \frac{2r m_v}{3} \cos(\omega_0 t - \omega_d t) + \frac{2(1-r) m_v}{3} \cos((\omega_0 t + \omega_d t) - \frac{\pi}{3}) \quad (3.47)$$

$$D_{Ac} = \frac{1}{3} + \frac{2r m_v}{3} \cos((\omega_0 t - \omega_d t) + \frac{\pi}{3}) + \frac{2(1-r) m_v}{3} \cos((\omega_0 t + \omega_d t) + \frac{\pi}{3}) \quad (3.48)$$

$$D_{Bc} = \frac{1}{3} + \frac{2r m_v}{3} \cos((\omega_0 t - \omega_d t) - \frac{\pi}{3}) + \frac{2(1-r) m_v}{3} \cos((\omega_0 t + \omega_d t)) \quad (3.49)$$
Figure 3.5 Input power factor control (a) input current positions, (b) lagging power factor ($d^+ < d^-$), (c) unity power factor ($d^+ = d^-$) and (d) leading power factor ($d^- < d^+$)

Let $T^+$ be the time for which the positive rotating space vectors are applied and $T^-$ be the time for which the negative rotating space vectors are applied, within a given sampling time $T_s$. The respective duty ratios are $d^+$ and $d^-$ and they satisfy the relation $d^+ + d^- = 1$ at all times. Equation (3.50) gives the input power angle $\theta$.

$$\theta = \tan^{-1}\left(\frac{1-2d^+}{\tan \rho}\right)$$  \hspace{1cm} (3.50)
3.7 RSVM TECHNIQUE FOR THE PSDSMC

The RSVM technique for the CMC uses only three vectors with the modulation index limited to 0.5, which is a major limitation of the RSVM technique. The shortcoming of this technique is overcome by using a 6×3 CMC consisting of two 3×3 CMCs [MC\(_x\), MC\(_y\)] fed by a three-phase center-tapped transformer. The transformer produces 180° shifted space vector pattern by generating a six-phase supply, as shown in Figure 3.7. In the proposed topology, the modulation index is extended to 0.866 with the help of the newly available three space vectors, as shown in Figure 3.6. Equations (3.51) and (3.52) give the six switching patterns for the counterclockwise rotating space vectors, where all the elements of the S\(_{MC,x}\) are equal to 0 if any one switching sequence of the S\(_{MC,y}\) is applied. Similarly, all the elements of the S\(_{MC,y}\) are equal to 0 if any one switching sequence of the S\(_{MC,x}\) is applied.

![Figure 3.6 (Continued)](image-url)
Figure 3.6 Space vector distribution of the PSDSMC (a) +ve sequence vectors and (b) -ve sequence vectors

Figure 3.7 Phase shifted dual source matrix converter
Equations (3.53) to (3.59) give the positions of the active switching voltage space vectors and the reference output voltage space vector

\[
    \mathbf{v}_{abc}^+ = \frac{3}{2} V_i e^{j\alpha t} 
\]

\[
    \mathbf{v}_{cab}^+ = \frac{3}{2} V_i e^{j(\alpha t + \frac{2\pi}{3})} 
\]

\[
    \mathbf{v}_{bca}^+ = \frac{3}{2} V_i e^{j(\alpha t - \frac{2\pi}{3})} 
\]

\[
    \mathbf{v}_{abc}^y = \frac{3}{2} V_i e^{j(\alpha t + \pi)} 
\]

\[
    \mathbf{v}_{cab}^y = \frac{3}{2} V_i e^{j(\alpha t - \frac{2\pi}{3})} 
\]

\[
    \mathbf{v}_{bca}^y = \frac{3}{2} V_i e^{j(\alpha t + \frac{2\pi}{3})} 
\]

\[
    \mathbf{v}_o = \frac{3}{2} V_o e^{j\beta t} 
\]

where, \( \mathbf{v}_{abc}^+, \mathbf{v}_{cab}^+, \mathbf{v}_{bca}^+ \) and \( \mathbf{v}_{abc}^y, \mathbf{v}_{cab}^y, \mathbf{v}_{bca}^y \) are respectively the active positive rotating switching vectors corresponding to \( \mathbf{S}_{MC,x} \) and \( \mathbf{S}_{MC,y} \) respectively. \( V_i \) and \( V_o \) are the magnitudes of the input and the output space vectors respectively. The duty cycles of the active vectors and the zero vector
are computed using the sine law of triangles, as explained in section 2, and given by Equation (3.60).

\[ d_\alpha = m_v \sin(60^\circ - \theta_v), \quad d_{\beta} = m_v \sin(\theta_v), \quad d_0 = 1 - d_\alpha - d_{\beta} \]  

(3.60)

As explained in section 2, this technique does not utilize zero vectors. From Equations (3.53) to (3.59), it can be shown that \( d_m v_{abc,x}^+ + d_m v_{abc,y}^+ = 0 \), where \( d_m \) is some arbitrary duty cycle. To achieve the sinusoidal output and to eliminate the common mode voltage, RSVM technique utilizes two opposite active vectors in equal ratio within the time for zero switching that modifies the duty ratios, as given by Equation (3.61).

\[ d_1^+ = d_\alpha, \quad d_2^+ = d_1 + \frac{d_0}{2}, \quad d_3^+ = \frac{d_0}{2} \]  

(3.61)

Hence, within any given sector, the required output can be synthesized by applying the respective active rotating vectors, as given in Table 3.3. Equation (3.62) gives the relation between the output voltage \( v_o \) and the active vectors.

\[ v_o = d_1^+ v_1 + d_2^+ v_2 + d_3^+ v_3 \]  

(3.62)

<table>
<thead>
<tr>
<th>Table 3.3</th>
<th>Positive rotating switching vectors and sector no. for the PSDSMC</th>
</tr>
</thead>
<tbody>
<tr>
<td>S. No.</td>
<td>( \theta_v )</td>
</tr>
<tr>
<td>1</td>
<td>( 0^\circ &lt; \theta_v \leq 60^\circ )</td>
</tr>
<tr>
<td>2</td>
<td>( 60^\circ &lt; \theta_v \leq 120^\circ )</td>
</tr>
<tr>
<td>3</td>
<td>( 120^\circ &lt; \theta_v \leq 180^\circ )</td>
</tr>
<tr>
<td>4</td>
<td>( 180^\circ &lt; \theta_v \leq 240^\circ )</td>
</tr>
<tr>
<td>5</td>
<td>( 240^\circ &lt; \theta_v \leq 300^\circ )</td>
</tr>
<tr>
<td>6</td>
<td>( 300^\circ &lt; \theta_v \leq 360^\circ )</td>
</tr>
</tbody>
</table>
As described in section 2, the same output voltage can be obtained by using the negative rotating space vectors, as shown in Figure 3.6(b). Within any given sector, the required output can be synthesized by applying the respective active rotating vectors, as given in Table 3.4.

Table 3.4  Negative rotating switching vectors and sector no. for the PSDSMC

<table>
<thead>
<tr>
<th>S. No.</th>
<th>$\theta_v$</th>
<th>Sector No.</th>
<th>Active vectors</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>$0^\circ \leq \theta_v \leq 60^\circ$</td>
<td>1</td>
<td>$v_{acb}^-$ $v_{cba}^-$ $v_{cab}^-$</td>
</tr>
<tr>
<td>2</td>
<td>$60^\circ \leq \theta_v \leq 120^\circ$</td>
<td>2</td>
<td>$v_{cba}^-$ $v_{bac}^-$ $v_{bca}^-$</td>
</tr>
<tr>
<td>3</td>
<td>$120^\circ \leq \theta_v \leq 180^\circ$</td>
<td>3</td>
<td>$v_{bac}^-$ $v_{acb}^-$ $v_{abc}^-$</td>
</tr>
<tr>
<td>4</td>
<td>$180^\circ \leq \theta_v \leq 240^\circ$</td>
<td>4</td>
<td>$v_{abc}^-$ $v_{cba}^-$ $v_{bca}^-$</td>
</tr>
<tr>
<td>5</td>
<td>$240^\circ \leq \theta_v \leq 300^\circ$</td>
<td>5</td>
<td>$v_{cba}^-$ $v_{bac}^-$ $v_{bca}^-$</td>
</tr>
<tr>
<td>6</td>
<td>$300^\circ \leq \theta_v \leq 360^\circ$</td>
<td>6</td>
<td>$v_{bac}^-$ $v_{acb}^-$ $v_{abc}^-$</td>
</tr>
</tbody>
</table>

In a similar manner, as described in Section 2, the ratio in which the +ve and –ve rotating voltage space vectors are applied decides the input power factor at the primary of the center tapped transformer, as given by Equation (3.50).

It is also observed that the input power factor control reduces the magnitude of the input current, as shown in Figure 3.8, by a factor given in Equation (3.63)

$$\frac{|I_{s,c}|}{|I_s^+|} = \frac{\cos \rho}{\cos(\tan^{-1}[\cos(1-2d^+ \tan \rho)])}$$

(3.63)

where, $|I_{s,c}|$ and $|I_s^+|$ are respectively the peak magnitude of the controlled input current and the peak magnitude of the uncontrolled input current of the CMC.
Figure 3.8 Input current magnitude for unity power factor
(a) instantaneous and (b) vector

From Equation (3.63), it is observed that the modulation index at the output of the matrix converter reduces by the same factor, as described in Equation (3.64).

\[
m_{v,c} = \frac{\cos \rho}{\cos(\tan^{-1}\left(\frac{1-2d^+}{\tan \rho}\right))} \cdot m_v \quad (3.64)
\]

3.8 UNBALANCE AND HARMONIC ANALYSIS OF THE RSVM TECHNIQUE

From Equation (3.5), it can be seen that when unbalanced inputs are applied to the CMC or the PSDSMC, the RSVM technique does not eliminate the CMV. For the inputs containing homopolar harmonics, the RSVM technique fails to eliminate the CMV, as homopolar harmonics introduce a zero sequence component. However, the magnitudes of the CMV introduced under such conditions are very low and that they do not affect the system very seriously. The RSVM technique is unaffected by non-homopolar harmonics since they do not introduce a zero sequence component in the system.
3.9 SIMULATION

Simulation of the CMC and the proposed PSDSMC was carried out using mathematical models, as shown in Figures 3.9 and 3.10, and also verified using ideal switches. The system parameters used in the simulation were: Supply – 220 V, 50 Hz, Load – R=5Ω, L=12 mH, \( \cos \varphi = 0.8 \) at 25 Hz output frequency and switching frequency of 7 kHz. The input filter capacitance and inductance were designed to be \( C_f=10 \mu F \) and \( L_f=1 \) mH with a damping resistor \( r_d = 15\Omega \) for filtering the higher order frequencies very near to the switching frequency.

![Figure 3.9](image)

**Figure 3.9** Mathematical model of the CMC (a) voltage model and (b) current model

The mathematical model of the PSDSMC is implemented using two three-phase AC voltage regulators connected to a single matrix converter, as shown in Figure 3.10. Although this topology reduces the switch count by three, it operates with three additional devices during conduction, which increases the conduction losses by 100%. Figure 3.11 shows the block diagram for implementing the RSVM technique.
Figure 3.10 Mathematical model of the PSDSMC (a) voltage model and (b) current model

Figures 3.12 and 3.13 show the simulation results of the CMC controlled by the RSVM technique without and with current control respectively. Figures 3.14 and 3.15 show the simulation results of the PSDSMC controlled by the RSVM technique without and with current control respectively. Figures 3.3 and 3.6 show that the modulation index of the PSDSMC controlled by the RSVM technique increases by 73.2% as
compared to modulation index of the CMC controlled by the RSVM technique.

Figure 3.11 Block diagram of the proposed current controlled RSVM technique

Figure 3.15(f) shows the increased modulation index of the PSDSMC as compared to the CMC in Figure 3.13(f). The peak magnitude of the load current in PSDSMC increases by approximately 73%.

(a)

Figure 3.12 (Continued)
Figure 3.12 (Continued)
Figure 3.12 RSVM technique for the CMC without current control (a) input current and voltage respectively, (b) output line voltage, (c) output phase voltage, (d) output line voltage magnified, (e) input currents and (f) output currents

Figure 3.13 (Continued)
Figure 3.13 RSVM technique for the CMC with current control (a) input current and voltage respectively, (b) output line voltage, (c) output phase voltage, (d) output line voltage magnified, (e) input currents and (f) output currents

Figures 3.12(d), 3.13(d), 3.14(d) and 3.15(d) show that the output voltage consists of three levels when the positive rotating vectors are applied while it consists of six levels when both the positive and the negative vectors are applied. This indicates that the number of switching states increases with
the current control technique. Figures 3.12(a), 3.13(a), 3.14(a) and 3.15(a) show that the input peak current reduces approximately by the factor given by the Equation (3.30).

Figure 3.14 (Continued)
Figure 3.14 RSVM technique for the PSDSMC without current control
(a) input current and voltage respectively, (b) output line voltage, (c) output phase voltage, (d) output line voltage
Magnified, (e) input currents and (f) output currents
Figure 3.15 (Continued)
Figure 3.15 RSVM technique for the PSDSMC with current control
(a) input current and voltage respectively, (b) output line voltage, (c) output phase voltage, (d) output line voltage magnified, (e) input currents and (f) output currents

From Figures 3.3 and 3.6, it can be observed that for the CMC and the PSDSMC the space vectors are distributed by 120° and 60° respectively. Due to this, the maximum voltage stresses on the devices during the
commutation for the CMC and the PSDSMC are $\sqrt{3}V_m$ and $V_m$ respectively. Hence, it can be inferred that during commutation the devices in the CMC topology are subjected to higher voltage stresses.

Figure 3.16 (a) shows the CMV induced due to the ISVM technique in the CMC and Figure 3.16 (b) shows the elimination of the CMV by the RSVM technique for both the CMC and the PSDSMC. The peak of the CMV can be as high as the magnitude of the input phase voltage in the ISVM technique. This has been eliminated in the proposed RSVM technique along with the input current control.

![Figure 3.16 (a)](image1)

![Figure 3.16 (b)](image2)

Figure 3.16 Common mode voltage (a) ISVM technique for the CMC and (b) RSVM technique for the CMC (or) the PSDSMC
Figure 3.17(a) shows that an unbalance of 4.3% in the B phase of the input voltage is applied to the CMC and the PSDSMC. Figure 3.17(b) illustrates that a low magnitude, low frequency CMV, proportional to the magnitude of the unbalance, is present in the CMC. However, Fig. 3.17(c) indicates that in the PSDSMC, a low magnitude, high frequency CMV is introduced by the space vectors of the phase shifted matrix converter.

![Figure 3.17](image)

**Figure 3.17** (a) Unbalanced input voltage, (b) CMV of the CMC and (c) CMV of the PSDSMC
Figure 3.18(a) shows a third harmonic (homopolar) component injected at the input from 0.04 s to 0.08 s and a second harmonic (non-homopolar) component injected from 0.08 s to 0.12 s. Figures 3.18(b) and (c) indicate the elimination of the CMV for the non-homopolar harmonics while the homopolar harmonics introduce a CMV due to the zero sequence voltage.

Figure 3.18 (a) Non-sinusoidal input voltage, (b) CMV of the CMC and (c) CMV of the PSDSMC
As the number of branches increases, as in the case of the PSDSMC, the current and voltage stresses are shared among the branches equally, hence reducing the stresses on each device and extending the life of the device. If one of the arms fails to conduct, the PSDSMC can be made to operate as a CMC with reduced CMV by optimal vector selection in the ISVM technique. Considering the economic aspect, this solution for the CMV elimination seems to be costly. However, with the growth of the power electronics technology, the cost of the devices is expected to come down. Although the design looks expensive, it can be used where CMV elimination is essential.

3.10 SUMMARY

The PSDSMC topology has twice the number of switches as compared to the CMC topology. However, it is the only possible configuration to eliminate the common mode voltage with an increased modulation index of 0.866 in the direct AC-to-AC converters fed machines. It is found that in the RSVM technique, current control can be carried out using both the positive and the negative rotating vectors. RSVM technique also eliminates the CMV for the inputs having non-homopolar harmonics. However, it does not eliminate the CMV for unbalanced inputs or inputs with homopolar harmonics.

The input current control range of the PSDSMC and the CMC is limited from unity power factor to output power factor. In the current controlled technique for the PSDSMC and the CMC, the modulation index reduces further. It is also found that the switching stresses of the individual switches reduce in the PSDSMC as compared to the CMC because of the additional states introduced in the PSDSMC topology. Hence, the proposed PSDSMC modulated by the RSVM technique can be used for the elimination of the CMV in electrical machines with a higher modulation index.