CARRIER BASED HYBRID PWM ALGORITHM WITH REDUCED COMMON MODE VOLTAGE FOR THREE PHASE VOLTAGE SOURCE INVERTER FED INDUCTION MOTOR DRIVES

Pradeep B Jyoti¹, Dr. J Amarnath², Dr. Subba Rayadu³,

¹Dept. of Electrical & Electronics Engineering, R.Y.M Engineering College, Bellary, Karnataka, India,
²Dept. of Electrical & Electronics Engineering, JNTUH College of Engineering, Hyderabad, Andhrapradesh, India
³Dept. of Electrical & Electronics Engineering, G. Pulla Reddy Engineering College, Kurnool, Andhrapradesh, India

ABSTRACT

This paper proposes a simple carrier based hybrid PWM algorithm (HPWM) to mitigate the common mode voltage (CMV) variations. Though the space vector PWM (SVPWM) gives, it results in large CMV values. Hence, to reduce the CMV and complexity, which is involved in the SVPWM technique, this paper presents simple scalar based PWM techniques. By adding the zero sequence voltage to the phase voltages, the modulating signals are derived for the active zero state PWM (AZSPWM) techniques. Also, to reduce the harmonic distortion further, a HPWM technique is presented based on the concept of the stator flux ripple. As the proposed AZSPWM and HPWM techniques do not use zero states, the CMV variations can be minimized from \( \pm V_{dc}/2 \) to \( \pm V_{dc}/6 \) when compared with the SVPWM algorithm. To evaluate the proposed algorithm simulation studies have been carried on v/f controlled induction motor drive and the results are presented.

Index Terms: AZSPWM, common mode voltage, SVPWM, vector control
I. INTRODUCTION

The Pulse width Modulated Voltage Source Inverter (PWMVSI) fed Variable Speed Drives are increasingly used in many applications. A detailed survey of newly developed PWM algorithms which are different in concept and performance are described in [1]. Compared with the sinusoidal modulation, the PWM method based on the voltage space vectors result in reduced harmonic distortion and excellent dc bus utilization [2]. In SVPWM, based on the magnitude of the reference voltage the switching times are calculated. Though this algorithm gives out superior waveform quality, it gives more switching losses and harmonic distortion at higher modulation indices. In order to obtain superior waveform quality at higher modulation indices, the Discontinuous PWM (DPWM) techniques are proposed [3-4]. But, at all modulation indices SVPWM and DPWM algorithms give large amount of Common Mode Voltage (CMV) variations between $\pm \frac{V_{dc}}{2}$ due to the presence of zero states.

Moreover, with the increase in CMV, the Common Mode Current (CMV) flows through the winding which cause bearing problems and early mechanical failure of the machine [5]. In the traditional days, the effect of CMV and CMC is reduced using passive and active filters which are explained in detail in [6]. But, an active and passive filter requires power electronic converter, inductor and capacitor which increase the cost and requires additional hardware. In the recent years, many researchers are focusing their interest on the development of various reduced CMV PWM algorithms without the requirement of additional equipment [7-8].

This paper first presents a simple space vector approach for the generation of different PWM techniques such as SVPWM, AZSPWM, NSPWM and DPWM algorithms. Then, by utilizing the concept of stator flux ripple, the rms stator flux ripple characteristics are plotted, from which finally Hybrid PWM algorithm (HPWM) techniques using flux ripples are presented with reduced common mode voltage and lesser harmonic distortion.

II. PROPOSED AZSPWM ALGORITHMS

A. Principal of Carrier Comparison Approach

Volt-Second balance is the most commonly used principle for all the PWM algorithms. According to this principle, the generation of the required reference voltage has to be done within a sampling time period. The principle of the carrier comparison approach involves the comparison of the triangular (carrier) signal with the modulating signal of a, b, c phases. The intersection point of both the waveforms give the switching instant of the switches. Out of the two switches of each leg of the inverter, at any instant either the top or bottom switch will conduct. The derivation of the fundamental output voltage depends on shape of modulating signal. In a sinusoidal PWM (SPWM) technique, three sinusoidal signals, which are shifted by 120° are compared with a common triangular signal for the generation of gating pulses. This technique is most commonly used for variable speed drive applications. But, the line current in SPWM algorithm gives more harmonic distortion. In a 3-phase VSI with isolated neutral point of the load, the voltage between the neutral point and midpoint of the dc bus ($V_{nm}$), which is also known as CMV as shown in Fig. 1 can be varied freely.
Two-level inverter fed induction motor drive

In order to maintain the constant position of the pulse, a zero sequence voltage can be added to the sinusoidal signals which do not affect the average value of the line voltage. But, due to the change in the pulse position, switching frequency characteristics and the harmonic properties will be affected.

B. Proposed Scalar Approach

The classical space vector approach which is explained in [2] requires the sector and angle information which increases the calculation burden. For simplicity, a simple scalar (carrier) approach is presented in this paper. So, in the proposed approach, the calculation of zero sequence signal is carried out using voltage magnitude test. Then, the modulating signals are generated by adding the zero sequence signal to the reference phase voltage.

The set of instantaneous reference phase voltages can be assumed as

\[ V_{in} = V_{ref} \cos(\theta - 2(r-1)\pi/3) \]

for \( i = a, b, c \) and \( r = 1, 2, 3 \)  

(1)

The calculated unique zero sequence signal \( V_{zs} \) based on the voltage magnitude tests is given in (2).

\[ V_{zs} = \frac{V_{dc}}{2} (2k_o - 1) - k_o V_{max} + (k_o - 1)V_{min} \]

(2)

The maximum and minimum values of \( V_{in} \) are represented as \( V_{max} \) and \( V_{min} \). By adding this zero sequence signal to the \( V_{in} \) the modulating waves \( V^*_{in} \) are derived as given in (3).

\[ V^*_{in} = V_{in} + V_{zs} \]

(3)

The variation of the constant \( k_o \) between 0 and 1, the modulating signals for various PWM techniques can be generated. In this paper, the main demonstration is regarding AZSPWM algorithms due to the absence of zero states in the switching sequence. But the modulating waveforms of SVPWM and AZSPWM are similar due to which the constant \( k_o \) value is fixed as 0.5.

In general carrier approach, the generated modulating waveforms are compared with a common triangular waveform to generate the pulse pattern. But, AZSPWM algorithms require, two triangular signals (\( V_{tri} \) and \( -V_{tri} \)) which are in phase opposition with each other. Then, by changing the polarity of the triangular signals different AZSPWM algorithms such as AZSPWM1, AZSPWM2
and AZSPWM3 can be generated. The following procedure can be used to change the polarity of the carrier signals.

The generation of for AZSPWM1, AZSPWM2 and AZSPWM3 techniques involves the three sets of 3-phase voltages are assumed as given in (4)-(6) respectively. These signals are used as test signals for the comparison with triangular waves.

\[
V_{ix} = V_{ref} \cos(\theta - 2(r - 1)\pi/3) \\
\text{for } i = a, b, c \text{ and } r = 1, 2, 3 \\
\]
\[
V_{iy} = V_{ref} \cos(\theta - 2(r - 1)\pi/3 + 2\pi/3) \\
\text{for } i = a, b, c \text{ and } r = 1, 2, 3 \\
\]
\[
V_{iz} = V_{ref} \cos(\theta - 2(r - 1)\pi/3 - 2\pi/3) \\
\text{for } i = a, b, c \text{ and } r = 1, 2, 3
\]

The change in polarity of carrier signals depends on the region. For all AZSPWM algorithms, the corresponding modulating signal of that phase will be compared with the +Vtri if the slope of the corresponding test signal is positive. Similarly, if the slope of the test signal is negative then the modulating signal of that phase will be compared with the -Vtri. After observing the obtained pulse pattern, the sequences of AZSPWM are shown in table.1. Also, the proposed scalar approach is as shown in Fig. 2

![Fig. 2 Proposed scalar approach](image)

<table>
<thead>
<tr>
<th>TABLE I. REGION DEPENDENT SWITCHING SEQUENCES FOR VARIOUS AZSPWM ALGORITHMS</th>
</tr>
</thead>
<tbody>
<tr>
<td>Sector</td>
</tr>
<tr>
<td>---</td>
</tr>
<tr>
<td>1</td>
</tr>
<tr>
<td>2</td>
</tr>
<tr>
<td>3</td>
</tr>
<tr>
<td>4</td>
</tr>
<tr>
<td>5</td>
</tr>
<tr>
<td>6</td>
</tr>
</tbody>
</table>
III. PROPOSED HYBRID PWM ALGORITHM

A. Analysis of Stator Flux Ripple

The measurement of the waveform quality generated by a PWM inverter is defined as the total harmonic distortion (THD) of a line current. It is given by

\[ I_{THD} = \frac{I_1}{I_n} \sqrt{\sum_{n=1}^{\infty} I_n^2} \]  

(8)

where, \( I_1 \) and \( I_n \) are the rms values of the fundamental and the nth harmonic components of the no-load current respectively.

All the AZPWM methods maintain volt-seconds balance to generate a voltage space vector, which equals to reference space vector in an average sense. The difference between the applied voltage vector and reference voltage is the ripple voltage vector. The error volt-seconds corresponding to ripple voltage vectors can be developed as given in [13]. For the AZPWM sequences the d-axis stator flux ripple components are given in terms of \( D, P, S \) and \( R \) and q-axis stator flux ripple components are given in terms of \( Q_1, Q_2, Q_3, Q_5, Q_7, Q_{22} \) which are defined as (22)-(25) and (26)-(32) are common for all the sequences.

The ripple voltage vectors and trajectory of the stator flux ripple for AZSPWM1 sequence are shown in Fig. 3. The corresponding d-axis and q-axis components of the stator flux ripple vector are as shown in Fig. 4. Here \( T_1, T_2, T_2 \) are the switching times of the voltage vectors \( V_1, V_2, \) and zero voltage vectors.

Fig. 3 Voltage ripple vectors and trajectory of the flux ripple
The error volt-seconds corresponding to the ripple voltage vectors for AZSPWM-I are as follows:

\[ V_{rip1T1} = \frac{2}{3} V_{dc} \sin \alpha T_1 + j(\frac{2}{3} V_{dc} \cos \alpha - V_{ref})T_1 = \lambda_D + j\lambda_{Q1} \]  \hspace{1cm} (9)

\[ V_{rip2T2} = \frac{2}{3} V_{dc} \sin (60^\circ + \alpha)T_2 + j(\frac{2}{3} V_{dc} \cos (60^\circ + \alpha) - V_{ref})T_2 = -\lambda_D + j\lambda_{Q2} \]  \hspace{1cm} (10)

\[ V_{rip3T3} = -\frac{2}{3} V_{dc} \sin (60^\circ - \alpha)T_3 - j(\frac{2}{3} V_{dc} \cos (60^\circ - \alpha) + V_{ref})T_3 = -\lambda_P - j\lambda_{Q3} \]  \hspace{1cm} (11)

\[ V_{rip4T4} = \frac{2}{3} V_{dc} \sin (60^\circ - \alpha)T_4 + j(-\frac{2}{3} V_{dc} \cos (60^\circ + \alpha) + V_{ref})T_4 = \lambda_P + j\lambda_{Q4} \]  \hspace{1cm} (12)

**Fig. 4** q-axis and d-axis components of the flux ripple vectors for AZSPWM

**Fig. 5** voltage ripple vectors and trajectory of the flux ripple for AZSPWM-II
Fig. 6 q-axis and d-axis components of the flux ripple vectors for AZSPWM-II

Similarly, the error volt-seconds corresponding to ripple voltage vectors of AZPWM2 sequence are given in (13)-(14).

\[
\begin{align*}
V_{rip1} \frac{T}{2} &= \frac{2}{3}V_{dc} \sin \alpha \frac{T}{2} + j(\frac{2}{3}V_{dc} \cos \alpha - V_{ref}) \frac{T}{2} = \lambda_S + j\lambda_{Q1} \\
V_{rip2} \frac{T}{2} &= -\frac{2}{3}V_{dc} \sin \alpha \frac{T}{2} - j(\frac{2}{3}V_{dc} \cos \alpha - V_{ref}) \frac{T}{2} = -\lambda_S - j\lambda_{Q1}
\end{align*}
\]

(13)\hspace{1cm}(14)

Fig. 7 voltage ripple vectors and trajectory of the flux ripple for AZSPWM-III
Fig. 8 q- and d-axis components of the flux ripple vectors for ABCPWM-II

The error volt-seconds corresponding to ripple voltage vectors of AZPWM3 sequence are given in (15)-(16).

\[
V_{vq22} = -\frac{2}{3} V_{dc} \sin(60^\circ - \alpha) + j\frac{2}{3} V_{dc} \cos(60^\circ - \alpha) - V_{ref} \frac{T_Z}{2} = -\lambda_R + j\lambda_{Q22}
\]  
\[
V_{vq5} = \frac{2}{3} V_{dc} \sin(60^\circ - \alpha) + j\frac{2}{3} V_{dc} \cos(60^\circ - \alpha) + V_{ref} \frac{T_Z}{2} = \lambda_R + j\lambda_{Q5}
\]

From the active state switching times of classical SVPWM algorithm, the following expressions can be derived

\[
\sin(60^\circ - \alpha) = \frac{\pi * T_1}{2 \sqrt{3} * M * T_s}
\]  
\[
\cos(\alpha) = \frac{\pi * (T_1 + 0.5 * T_2)}{3 * M * T_s}
\]  
\[
\cos(60^\circ - \alpha) = \frac{\pi * (T_2 + 0.5 * T_1)}{3 * M * T_s}
\]  
\[
\sin(60^\circ + \alpha) = \frac{\pi * (T_1 + T_2)}{2 \sqrt{3} * M * T_s}
\]  
\[
\cos(60^\circ + \alpha) = \frac{\pi * (T_1 - T_2)}{2 \sqrt{3} * M * T_s}
\]  
\[
\lambda_D = \frac{V_{dc} * \pi * T_2 * T_1}{3 \sqrt{3} * M * T_s}
\]  
\[
\lambda_R = \frac{V_{dc} * \pi * (T_1) * \left(\frac{T_s - T_1 - T_2}{2}\right)}{3 \sqrt{3} * M * T_s}
\]
The rms stator flux ripple for different sequences are derived as shown in (33)-(35)

\[
\lambda_p = \frac{V_{dc} \cdot \pi \cdot (T_1 + T_2) \cdot (T_s - T_1 - T_2)}{3 \cdot \sqrt{3} \cdot M \cdot T_s + 2} 
\]

\[
\lambda_s = \frac{V_{dc} \cdot \pi \cdot T_2 \cdot (T_s - T_1 - T_2)}{3 \cdot \sqrt{3} \cdot M \cdot T_s} 
\]

\[
\lambda_{q1} = \left( \frac{2}{3} V_{dc} \left( \frac{\pi \cdot (T_1 + 0.5 \cdot T_2)}{3 \cdot M \cdot T_s} \right) - V_{ref} \right) * T_1 
\]

\[
\lambda_{q2} = \left( \frac{2}{3} V_{dc} \left( \frac{\pi \cdot (T_2 + 0.5 \cdot T_1)}{3 \cdot M \cdot T_s} \right) - V_{ref} \right) * T_2 
\]

\[
\lambda_{q3} = \left( \frac{1}{3} V_{dc} \left( \frac{\pi \cdot (T_1 - T_2)}{3 \cdot M \cdot T_s} \right) + V_{ref} \right) * \left( \frac{T_s - T_1 - T_2}{2} \right) 
\]

\[
\lambda_{q4} = \left( -\frac{1}{3} V_{dc} \left( \frac{\pi \cdot (T_1 - T_2)}{3 \cdot M \cdot T_s} \right) - V_{ref} \right) * \left( \frac{T_s - T_1 - T_2}{2} \right) 
\]

\[
\lambda_{q5} = \left( \frac{2}{3} V_{dc} \left( \frac{\pi \cdot (T_2 + 0.5 \cdot T_1)}{3 \cdot M \cdot T_s} \right) - V_{ref} \right) * \left( \frac{T_s - T_1 - T_2}{2} \right) 
\]

\[
\lambda_{p5} = \left( \frac{2}{3} V_{dc} \left( \frac{\pi \cdot (T_1 + 0.5 \cdot T_2)}{3 \cdot M \cdot T_s} \right) - V_{ref} \right) * \left( \frac{T_s - T_1 - T_2}{2} \right) 
\]

The total rms ripple over a subcycle can be calculated as given in

\[
\lambda_{rms}^2 = \frac{1}{T_s} \int_0^{T_s} \lambda_p^2 dt + \frac{1}{T_s} \int_0^{T_s} \lambda_q^2 dt
\]
Thus, the rms stator flux ripple characteristics can be obtained for all the switching sequences. The stator flux ripple characteristics at a frequency of 45 Hz (modulation index of 0.8154) are shown in Fig. 7. From the ripple characteristics it is found that the AZSPWM-II and AZSPWM-III will give the same ripple characteristics. Hence, in the proposed hybrid AZSPWM (HAZSPMW) algorithm AZSPWM-I and AZSPWM-III are considered. Based on the ripple characteristics, the zones of superior performance (which results in reduced ripple) can be found for each sequence. Then, the identified sequence will be applied to the VSI, so that the harmonic distortion can be reduced at all modulation indices.

![RMS stator flux ripple variation of AZSPWM algorithms at a supply frequency of 45 Hz](image)

**Fig. 9** RMS stator flux ripple variation of AZSPWM algorithms at a supply frequency of 45 Hz

### III. SIMULATION RESULTS AND DISCUSSION

To prove the performance of proposed HPWM algorithm, the simulation studies have been carried out on v/f controlled induction motor drive by using the MATLAB. The switching frequency is taken as 5 kHz. The steady state results during the no-load condition at a modulation index of 0.815 (operating supply frequency of 45 Hz) are shown in From Fig. 8 to Fig. 12 along with the harmonic spectra of line current and CMV variations.
Fig. 10 (a) steady state plots for SVPWM algorithm based v/f controlled induction motor drive (b) Harmonic spectra of line current (c) CMV variations
Fig. 11 (a) steady state plots for AZSPWM1 algorithm based v/f controlled induction motor drive (b) Harmonic spectra of line current (c) CMV variations
Fig. 12 (a) steady state plots for AZSPWM2 algorithm based v/f controlled induction motor drive (b) Harmonic spectra of line current (c) CMV variations
Fig. 13 (a) steady state plots for AZSPWM3 algorithm based v/f controlled induction motor drive (b) Harmonic spectra of line current (c) CMV variations
IV. SIMULATION RESULTS AND DISCUSSION

To reduce the complexity of the PWM algorithm, a simplified scalar based approach has been presented in this paper. Moreover, to reduce the harmonic distortion further, a HPWM algorithm is developed by using the stator flux ripple concept. The proposed HPWM technique selects a suitable sequence that results in reduced harmonic distortion. To evaluate the performance of the proposed HPWM algorithm, simulation studies have been carried out and results were presented. From the results it can be observed that the proposed HPWM algorithm gives reduced CMV variations between $\pm V_{dc}/6$ with slightly increased ripple when compared to the SVPWM technique due to the opposite pulses in the line to line voltages.

V. REFERENCES