Comparative Evaluation of Space Vector Based Pulse Width Modulation Techniques in terms of Harmonic Distortion and Switching Losses

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ABSTRACT
Voltage source inverter (VSI) fed induction motors are widely used in variable speed applications. The harmonic distortion in the line currents of the inverter depends on the switching frequency and the pulse width modulation (PWM) technique employed to switch the inverter. The inverter switching loss is also strongly influenced by the PWM technique used. This paper compares the harmonic distortion and inverter switching losses due to different PWM techniques at a given average switching frequency. The paper evaluates a few specially designed PWM techniques based on space vector approach. It is shown that these techniques lead to reduced line current distortion as well as switching losses compared to well-known techniques such as sine-triangle PWM and conventional space vector PWM (CSVPWM).

Keywords
Induction motor drives, voltage source inverter, pulse width modulation, harmonic distortion, space vector, switching losses.

1. INTRODUCTION
A voltage source inverter (VSI) is used to supply an AC motor drive with a voltage of variable magnitude and frequency. The schematic of a two level VSI is shown in Fig.1. With a fixed DC input voltage, the semiconductor switches in the inverter are required to be switched in an appropriate fashion to obtain the desired fundamental voltage on the line side. Various PWM techniques have been reported for generation of gating pulses for the devices. The choice of PWM technique strongly influences line current distortion and switching losses in the inverter [1-2].

Sine-triangle PWM (SPWM) is a simple and popular method of PWM. In this method, three sinusoidal modulating waves, shifted by 120° from each other, are compared with a high frequency triangular carrier to get the switching pattern for the three phases. A sinusoidal modulating wave is shown in Fig.2(a). Suitable common-mode voltages can be added to the 3-phase sinusoidal modulating waves to achieve better harmonic performance and higher line-side voltage for a given DC bus voltage [1-2]. Examples of such modulating waves are shown in Fig.2(b) to Fig.2(d). With these techniques, a phase remains clamped to a given

2. SPACE VECTOR APPROACH TO PWM
Space vector based generation of PWM waveforms is explained in this section.

2.1 Inverter states and the space vector plane
A three phase quantity can be transformed into a space vector having components along two mutually perpendicular axes. The three phase input voltages at the stator terminals of an induction machine can be transformed to a space vector as shown in Eqn.1, where \( \tau_c(t) \) is the stator voltage space vector; \( v_{RN} \), \( v_{YN} \) and \( v_{BN} \) are the line-to-neutral...
Figure 2: Modulating waves for (a) sine-triangle PWM, (b) conventional space vector PWM, (c) 30° clamp PWM and (d) 60° clamp PWM. Peak value of sine wave is $V_m$. $V_p$ is the peak of triangular wave. Modulation index $\frac{V_m}{V_p} = 0.8$.

Figure 3: Space vectors of a two level voltage source inverter

The voltages $v_s(t) = v_{RN}(t) + v_{YN}(t)e^{j\frac{2\pi}{3}} + v_{BN}(t)e^{j\frac{4\pi}{3}}$ (1)

A two level VSI has $2^3$ or 8 combinations of switching states, which are known as the states of an inverter. Among these states, the two states 7 and 0 are known as zero states, as they yield zero output voltage. The remaining six states (1 to 6) give a non-zero output voltage, and are known as active states. The voltage vectors corresponding to the eight states are shown in Fig.3. The space vector plane is divided into six sectors by the vectors corresponding to active states. The sectors are marked as I to VI in Fig.3.

2.2 Dwell times of inverter states

In space vector based PWM, the voltage reference is provided using a revolving voltage reference vector $V_{REF}$ instead of three phase modulating waves. At steady state, the magnitude of this reference vector is proportional to the desired fundamental voltage and its frequency equals the fundamental frequency.

The revolving $V_{REF}$ vector is sampled once in every sub-cycle $T_s$, which gives the average vector to be applied in the given sub-cycle. This average vector can be best realized by the vectors bounding the sector in which it is present using the principle of volt-second balance. The reference and applied volt-seconds are balanced over a sub-cycle duration, i.e.

$$ V_{REF}T_s = V_xT_x + V_yT_y + V_zT_z $$ (2)

where $T_s$ is the duration of a sub-cycle; $V_x$ and $V_y$ are the active vectors bounding the corresponding sector; $V_z$ is the zero vector. The vectors $V_x$ and $V_y$ correspond to the active states $x$ and $y$ respectively. The zero vector $V_z$ may correspond to the zero state 0 or the zero state 7. For sector I, $x = 1$ and $y = 2$. $T_x$, $T_y$ and $T_z$ denote the dwell times of the states/vectors. The expressions for $T_x$, $T_y$, and $T_z$ are given by

$$ T_x = \frac{V_{REF}}{V_{DC}} \cdot \frac{\sin(60^\circ - \alpha)}{\sin 60^\circ} \cdot T_s $$ (3a)

$$ T_y = \frac{V_{REF}}{V_{DC}} \cdot \frac{\sin \alpha}{\sin 60^\circ} \cdot T_s $$ (3b)

$$ T_z = T_s - (T_x + T_y) $$ (3c)
3. ANALYTICAL EVALUATION OF SWITCHING SEQUENCES

In the present section, the switching sequences explained in previous section are compared in terms of RMS current ripple and switching energy lost over a sub-cycle.

3.1 RMS current ripple over a sub-cycle

At any given instant, there is an error between the reference voltage vector ($\vec{V}_{REF}$) and the applied voltage vector. The error voltage vectors corresponding to the vectors $\vec{V}_x$, $\vec{V}_y$ and $\vec{V}_z$ are illustrated in Fig.5(a). The integral of error voltage vector is the stator flux ripple vector. The trajectory of the tip of the stator flux ripple vector is parallel to the instantaneous error voltage vector [10-11].

For harmonic voltages, an induction motor can simply be modelled by its total leakage inductance. Therefore, the integral of error voltage, or the stator flux ripple, is proportional to the current ripple (both differ by a factor of machine leakage inductance). Hence, flux ripple is a measure of the ripple in line current [10-11].

The trajectory of the tip of stator flux ripple vector for sequence 0127 is shown in Fig.5(a) for $V_{REF} = 0.7V_{DC}$ and $\alpha = 20^\circ$. The stator flux ripple vector can be resolved along the reference axes of a synchronously revolving reference frame, namely $q$-axis and $d$-axis. The $q$-axis is aligned with $\vec{V}_{REF}$, while the $d$-axis lags behind the $q$-axis by $90^\circ$. The variation of $d$-axis and $q$-axis components of stator flux ripple vector for the sequence 0127 over a sub-cycle is shown in Fig.5(b). The error volt-second quantities $Q_x$, $Q_y$, $Q_z$ and $D$ marked in the plot are defined in Eqn.4.

\[
Q_x = -V_{REF} \cdot T_x \\
Q_y = (\cos \alpha - V_{REF}) \cdot T_x \\
Q_z = (\cos(60^\circ - \alpha) - V_{REF}) \cdot T_y \\
D = \sin \alpha \cdot T_x
\]

For sequences 012, 721, 0121, 7212, 1012 and 2721, the locus of the tip of the stator flux ripple vector and the $d$-axis and $q$-axis components of the vector for the same reference vector ($V_{REF} = 0.7V_{DC}$ and $\alpha = 20^\circ$) are shown in Fig.6. The RMS $d$-axis and $q$-axis flux ripple over a sub-cycle for a sequence $SEQ$ can be calculated using Eqn.5a and Eqn.5b. The total RMS flux ripple over a sub-cycle for a sequence $SEQ$ is given by Eqn.5c.

\[
\bar{\psi}^2_{dSEQ RMS} = \sqrt{\frac{1}{T_s} \int_0^{T_s} \bar{\psi}_{dSEQ}^2 \, dt} \\
\bar{\psi}^2_{qSEQ RMS} = \sqrt{\frac{1}{T_s} \int_0^{T_s} \bar{\psi}_{qSEQ}^2 \, dt} \\
\bar{\psi}^2_{SEQ RMS} = \sqrt{\bar{\psi}^2_{dSEQ RMS} + \bar{\psi}^2_{qSEQ RMS}}
\]

where $SEQ = 0127, 0121, 7212, 1012$ and 2721.

Figure 4: Seven possible sequences in sector $I$: (a) 0127, (b) 012, (c) 721, (d) 0121, (e) 7212, (f) 1012 and (g) 2721.

where $\alpha$ is the angle measured from the starting of a sector to $V_{REF}$ in anticlockwise direction.

2.3 Possible switching sequences

There are numerous possible ways of applying the vectors for corresponding time intervals. The zero vector can be applied using the zero state $+++(7)$ or the zero state $\cdots (0)$. An active state can be applied more than once in a sub-cycle [5-8,10-11]. The various possible sequences in sector $I$ are 0127,012,012,721,272,1012 and 2721. These seven sequences are illustrated in Fig.4 [7-8]. In all these sequences, the number of switchings in a sub-cycle is less than or equal to three. Also, only one phase switches during a transition from one state to the next. The sub-cycle duration for sequences 012 and 721 is taken to be $\frac{1}{2}$ of that for the other sequences because there are only two switchings per sub-cycle. This facilitates comparison of all the sequences at the same average switching frequency.
Figure 5: (a) Error voltage vectors corresponding to applied vectors $V_x$, $V_y$ and $V_z$, and the trajectory of the tip of stator flux ripple vector for 0127 (b) Variation of stator flux ripple along $d$ and $q$ axes over a sub-cycle for 0127.

Figure 6: Trajectory of the tip of stator flux ripple vector and the variation of flux ripple along $d$ and $q$ axes over a sub-cycle for sequences (a) 012, (b) 721, (c) 0121, (d) 7212, (e) 1012 and (f) 2721.
functions of \( \alpha \) for all the sequences.

As seen from the equations, the RMS flux ripple over a sub-cycle for a given sequence is a function of \( V_{REF} \), \( \alpha \) and \( T_r \). These expressions can be used to compare the RMS current ripple over a sub-cycle for different sequences.

### 3.2 Switching energy lost in a sub-cycle

The switching energy loss in a sub-cycle in an inverter leg is directly proportional to the current in that phase and the current ripple over a sub-cycle for different sequences. If \( n_R \), \( n_Y \) and \( n_B \) are the number of switchings of phases \( R \), \( Y \) and \( B \), respectively, in a given sub-cycle for a sequence \( SEQ \), the normalized switching energy loss in the inverter in that sub-cycle is given by

\[
E_{\text{SUB SEQ}} = E_R \text{ SEQ} + E_Y \text{ SEQ} + E_B \text{ SEQ} = n_R \left[ \frac{|I_R|}{I_m} \right] + n_Y \left[ \frac{|I_Y|}{I_m} \right] + n_B \left[ \frac{|I_B|}{I_m} \right] \tag{7}
\]

where

- \( I_R = I_m \sin(\omega t + \phi) \)
- \( I_Y = I_m \sin(\omega t - 120^\circ + \phi) \)
- \( I_B = I_m \sin(\omega t - 240^\circ + \phi) \)
- \( \omega = 2\pi \cdot \text{fundamental frequency} \)
- \( \phi = \text{power factor angle} \)

The values of \( n_R \), \( n_Y \) and \( n_B \) for various sequences in sector \( I \) are given in Table 2 along with the corresponding duration of sub-cycle. In any sector, \( \omega t \) can be written in terms of \( \alpha \) (for sector \( I \), \( \omega t = 90^\circ + \alpha \)). Therefore, the normalized switching energy loss over a sub-cycle is a function of \( \alpha \) and power factor angle \( \phi \).

### 4. HYBRID PWM TECHNIQUES

The modulating waves corresponding to CSVPWM, 30\(^\circ\) clamp PWM and 60\(^\circ\) clamp PWM were shown in Fig.2(b), 2(c) and 2(d). These techniques are illustrated in the space vector domain in Fig.7. In CSVPWM, sequence 0127-7210 is employed sequences 0127, 0121 and 7212. For 60\(^\circ\) clamp PWM were shown in Fig.2(b), 2(c) and 2(d). These techniques are illustrated in the space vector domain in Fig.7. In CSVPWM, sequence 0127-7210 is employed sequences 0127, 0121 and 7212.

### Table 1: Coefficients \( C_0 \), \( C_1 \) and \( C_2 \) for all the sequences

\( a = \sin(60^\circ + \alpha) \) and \( b = \sin(60^\circ - \alpha) \)

<table>
<thead>
<tr>
<th>( SEQ )</th>
<th>( C_0(SEQ) )</th>
<th>( C_1(SEQ) )</th>
<th>( C_2(SEQ) )</th>
</tr>
</thead>
<tbody>
<tr>
<td>0127</td>
<td>( \frac{1}{12} )</td>
<td>( \frac{1}{3\sqrt{3}} (ab^4 + a^3b^2 - a^2b^3) )</td>
<td>( \frac{1}{3} \left( a^3b + 2ab^3 - a^2b^2 \right) )</td>
</tr>
<tr>
<td></td>
<td></td>
<td>( + \frac{2}{3} (-2b^5 - 3ab^4 - a^2b) + b - a )</td>
<td>( + a^2 + 4b^2 - 2ab )</td>
</tr>
<tr>
<td>0121</td>
<td>( \frac{1}{3} )</td>
<td>( \frac{1}{3\sqrt{3}} \left[ -b^5 - ab^2 - 2a^2b - 12a + 3b \right] )</td>
<td>( \frac{1}{3} \left[ a^2b - ab^3 + 2a^3b \right] )</td>
</tr>
<tr>
<td></td>
<td></td>
<td>( + \frac{2}{3} \left( 4a^2 - 2ab + b^2 \right) )</td>
<td></td>
</tr>
<tr>
<td>7212</td>
<td>( \frac{1}{3} )</td>
<td>( C_1(0121)(60^\circ - \alpha) )</td>
<td>( C_2(0121)(60^\circ - \alpha) )</td>
</tr>
<tr>
<td>012</td>
<td>( \frac{1}{3} )</td>
<td>( \frac{1}{3\sqrt{3}} \left[ \frac{2}{3}a^5 + 4ab^2 - 8a^2b - 4b^3 - 6a + 6b \right] )</td>
<td>( \frac{1}{3} \left[ (-15a^3b^2 + 11ab^3 + 14a^3b - 4a^4 - 4b^4) \right] )</td>
</tr>
<tr>
<td></td>
<td></td>
<td>( + \frac{2}{3} \left( 4a^2 - 2ab + b^2 \right) )</td>
<td></td>
</tr>
<tr>
<td>721</td>
<td>( \frac{1}{3} )</td>
<td>( C_1(0121)(60^\circ - \alpha) )</td>
<td>( C_2(0121)(60^\circ - \alpha) )</td>
</tr>
<tr>
<td>1021</td>
<td>( \frac{1}{3} (1 - ab) )</td>
<td>( \frac{1}{3\sqrt{3}} \left[ 16a^5 - 2b^5 + 20ab^4 - 56a^2b + 74a^3b^2 \right] )</td>
<td>( \frac{1}{3} \left[ (-16a^4 - 6b^4 + 38a^3b + 25ab^3 - 41a^2b^2) \right] )</td>
</tr>
<tr>
<td></td>
<td></td>
<td>( - 60a^2b^3 + 24a^2b - 3ab^2 - 9b^3 + 18b - 27a )</td>
<td>( + 12a^2 - 18ab + 7b^2 )</td>
</tr>
<tr>
<td>2721</td>
<td>( C_0(0121)(60^\circ - \alpha) )</td>
<td>( C_1(0121)(60^\circ - \alpha) )</td>
<td>( C_2(0121)(60^\circ - \alpha) )</td>
</tr>
</tbody>
</table>

### Table 2: Number of switchings of the phases in a sub-cycle for different sequences in sector \( I \)

<table>
<thead>
<tr>
<th>( SEQ )</th>
<th>( n_R )</th>
<th>( n_Y )</th>
<th>( n_B )</th>
<th>( T_r )</th>
</tr>
</thead>
<tbody>
<tr>
<td>0127</td>
<td>1</td>
<td>1</td>
<td>1</td>
<td>( T )</td>
</tr>
<tr>
<td>0121</td>
<td>1</td>
<td>2</td>
<td>0</td>
<td>( T )</td>
</tr>
<tr>
<td>7212</td>
<td>0</td>
<td>2</td>
<td>1</td>
<td>( T )</td>
</tr>
<tr>
<td>012</td>
<td>1</td>
<td>1</td>
<td>0</td>
<td>( \frac{2}{3} T )</td>
</tr>
<tr>
<td>721</td>
<td>0</td>
<td>1</td>
<td>1</td>
<td>( \frac{2}{3} T )</td>
</tr>
<tr>
<td>1012</td>
<td>2</td>
<td>1</td>
<td>0</td>
<td>( T )</td>
</tr>
<tr>
<td>2721</td>
<td>0</td>
<td>1</td>
<td>2</td>
<td>( T )</td>
</tr>
</tbody>
</table>
4.1 Comparison of sequences in terms of RMS current ripple

The RMS current ripple over a sub-cycle corresponding to sequences 0127, 0121 and 7212 is compared here. Based on the expressions for RMS flux ripple over a sub-cycle derived in section 3.1, the sequences can be compared with each other and their individual zones of superior performance can be identified.

The RMS flux ripple over a sub-cycle is plotted in Fig.8 for the whole range of \( \alpha \). Here, \( V_{REF} = 0.866 \text{p.u.} \) and \( T_s = 100 \mu s \). It is evident from Fig.8 that for \( V_{REF} = 0.866 \text{p.u.} \), sequences 0121 and 7212 yield lesser ripple than 0127 in the range of \( 5^\circ \leq \alpha \leq 55^\circ \) (approximately). Similarly, for different values of \( V_{REF} \), the range of \( \alpha \) over which these sequences result in less ripple will be different. These sets of \( V_{REF} \) and \( \alpha \) values for which a sequence gives the least ripple indicate the zone of superior performance of that particular sequence. It can be observed from Fig.8 that the flux ripple given by 0121 and 7212 are close to each other, and 0121 results in less ripple in the first half of the sector while 7212 yields less ripple in the second half. Based on this discussion, a hybrid PWM technique can be designed to give reduced current ripple, which is shown in Fig.11(a). This technique is referred to as the three-zone hybrid PWM technique [6,8].

4.2 Comparison of sequences in terms of switching losses

In a similar manner, various switching sequences can be compared in terms of switching losses based on the switching energy lost in each sub-cycle. The normalized switching energy loss in a sub-cycle \( E_{SUB \_SEQ} \) is a function of \( \alpha \) and power factor angle \( \phi \). The normalized switching energy loss in a sub-cycle is plotted for sequences 0127, 0121 and 7212 over the whole range of \( \alpha \) in Fig.9 \( (V_{REF} = 0.866 \text{p.u.}, I_m = 1.0 \) and \( \phi = 30^\circ \) lagging) and Fig.10 \( (V_{REF} = 0.866 \text{p.u.}, I_m = 1.0 \) and \( \phi = 0^\circ \)). From Fig.9 and Fig.10, it can be observed that 7212 produces lower switching losses than 0127 for all values of \( \alpha \), for power factors 0.866 lagging and unity. Whereas 0121 gives better performance in terms of switching losses only for some values of \( \alpha \) at different power factors. For lagging power factor operation, which is typical in induction motor drives, using 7212 instead of 0121 will be a better option in terms of switching loss performance. This will not significantly affect the current ripple due to the PWM technique as 0121 and 7212 have closer characteristics in terms of flux ripple (Fig.8). Therefore, the three-zone hybrid PWM technique in Fig.11(a) can be modified into a simpler technique shown in Fig.11(b) to give better performance in terms of switching losses. This technique is referred to as the two-zone hybrid PWM technique [7].
Figure 9: Normalized switching energy loss over a sub-cycle Vs $\alpha$ for sequences 0127, 0121 and 7212 ($V_{\text{REF}} = 0.866 \ \text{p.u.}, \ I_m = 1.0 \ \text{and} \ \phi = 30^\circ \ \text{lagging}$).

Figure 10: Normalized switching energy loss over a sub-cycle Vs $\alpha$ for sequences 0127, 0121 and 7212 ($V_{\text{REF}} = 0.866 \ \text{p.u.}, \ I_m = 1.0 \ \text{and} \ \phi = 0^\circ$).

Figure 11: PWM techniques considered for evaluation - 2: (a) Three-zone hybrid PWM (b) Two-zone hybrid PWM

5. EVALUATION OF PWM TECHNIQUES IN TERMS OF HARMONIC DISTORTION AND SWITCHING LOSSES

In section III, evaluation of sequences at a sub-cycle level is presented. This forms the basis for design and evaluation of various PWM techniques. Comparative evaluation of PWM techniques is presented here.

The techniques considered for evaluation are conventional space vector PWM (CSVPWM), $60^\circ$ clamp PWM, $30^\circ$ clamp PWM, three-zone hybrid PWM and two-zone hybrid PWM. These techniques are summarized in Fig.7 and Fig.11.

5.1 Evaluation of harmonic distortion

The mean square flux ripple over a sub-cycle can be averaged over a sector to give the mean square flux ripple over a sector. Thus, the RMS flux ripple over a sector $\Psi_{RMS}$, for a given $V_{\text{REF}}$ and $T_s$ is given by

$$\Psi_{RMS} = \sqrt{\frac{3}{\pi}} \int_0^{\frac{\pi}{3}} \Psi_{RMS \ \text{SEQ}}(\alpha) \ d\alpha$$  \hspace{1cm} (8)

where $SEQ$ is the sequence applied in that sub-cycle, which
Leading p.f. - 0
Lagging p.f. - 90

Figure 14: Normalized switching power loss for different PWM Techniques Vs Power factor angle for \( V_{REF} = 0.866 \text{ p.u.} \) (1.0 p.u. = \( V_{DC} \)). Normalization is w.r.t. CSVPWM.

depends on the PWM technique employed.

The total harmonic distortion (THD) in current \( I_{THD} \) is defined as the ratio of RMS value of the ripple current to RMS value of the fundamental component of the total current.

\[
I_{THD} = \frac{I_{RMS}}{I_{RMS}} = \sqrt{\left( \frac{I_{RMS}^2 - I_{RMS}^2}{I_{RMS}} \right)}
\]

As current ripple is proportional to stator flux ripple, the harmonic distortion in stator flux can be taken as the analytical measure of harmonic distortion in line current. The THD in stator flux \( \Psi_{THD} \) is given by

\[
\Psi_{THD} = \frac{\Psi_{RMS}}{\Psi_{RMS}}
\]

where \( \Psi_{RMS} \) is the RMS value of the fundamental component of stator flux, which is directly related to \( V_{REF} \). Hence, for a given switching frequency, \( \Psi_{THD} \) can be calculated for various PWM techniques over the range of \( V_{REF} \).

Fig.12 shows \( \Psi_{THD} \) for different PWM techniques as a function of fundamental frequency. As mentioned in section 2, the sampling interval \( T_s = \frac{2}{f_s} \) for sequences 012 and 721, while \( T_s = T \) for other sequences. The duration \( T \) is chosen as 200\( \mu \)s, yielding an average switching frequency \( f_{sw} = \frac{2}{f_s} = 2.5 \text{kHz} \). Fig.13 shows normalized total harmonic distortion for different PWM techniques. The normalization here is with respect to the THD corresponding to CSVPWM.

5.2 Evaluation of switching losses

In order to evaluate the switching loss performance of different PWM techniques, the total switching loss in the inverter has to be calculated. The average switching loss per phase over a fundamental cycle is equal to the total switching loss (three phases together) averaged over a sector. Presently, the normalized switching loss is used for comparison.

The normalized switching energy loss averaged over a sector for a given power factor angle \( \phi \) is given by

\[
E_{AV} = \frac{3}{\pi} \int_0^{\phi} E_{SUB \ SEQ}(\alpha) \, d\alpha
\]

where \( \text{SEQ} \) is the sequence applied in that sub-cycle, which depends on the PWM technique employed.

For calculating the normalized switching power loss, the normalized switching energy loss needs to be multiplied by the sampling frequency. The normalized switching power loss is calculated for the four PWM techniques considered and it is once again normalized w.r.t. CSVPWM. The variation of this normalized switching power loss over the full range of power factor angle is shown in Fig.14 for all the PWM techniques. The variation of normalized switching power loss for different PWM techniques with fundamental frequency at a power factor 0.866 lagging is shown in Fig.15.

5.3 Discussion

It is evident from Fig.12 and Fig.13 that all the techniques discussed give a better performance in terms of THD compared to CSVPWM when the drive is operating at higher modulation indices. The frequency range of superior performance is typically 35Hz to 50Hz. The 30\(^\circ\) clamp PWM gives the best performance in terms of THD over the range of 35Hz to 45Hz. The two-zone and three-zone PWM techniques are almost identical in THD performance. These lead to a reduction of 42\% in THD when compared to CSVPWM at 50Hz.

Referring to the normalized switching loss curve in Fig.14, the two-zone PWM technique is the best in terms of switching loss for lagging power factor angles ranging from 0\(^\circ\) to 60\(^\circ\). It gives 30\% reduction in switching loss over CSVPWM at a power factor angle close to 10\(^\circ\) lagging. For leading power factors ranging from unity to 0.707, 60\(^\circ\) clamp PWM is the best technique, giving a maximum reduction of 22\% at unity p.f.

6. EXPERIMENTAL RESULTS

CSVPWM, two-zone PWM and three zone PWM techniques are implemented using ALTERA CycloneII based FPGA platform [12]. A 2.2kW, 415V, 50Hz, squirrelcage induction motor is operated at no load from a 10kVA IGBT based two level VSI with a DC bus voltage of 600V. The average switching frequency used for the PWM techniques is 2.5kHz. The experimental motor current waveforms are
Figure 16: Line current waveforms of the motor at no load. The measured THD values are: (a) 10.07%, (b) 6.00% and (c) 6.43%.

The THD measurements of these waveforms show that at 50Hz, the THD in line current for CSVPWM is 10.07%. The three-zone technique gives a THD of 6.00%, which is 40.4% less compared to CSVPWM. This is the best reduction in THD among the techniques discussed. The two-zone technique gives a THD of 6.43%, which is 36.15% less compared to CSVPWM. These techniques yield superior performance at high modulation indices. The typical range of superior performance is 35-50 Hz. The measured THD values over this range of frequencies is shown in Fig.17. It is observed that the reduction in harmonic distortion starts at 38Hz, and increases towards the rated operating point.

7. CONCLUSION

A method of evaluating different PWM techniques in terms of harmonic distortion as well as switching losses is presented. The analysis for any technique can be pursued along similar lines. Four different hybrid PWM techniques are compared considering CSVPWM as the benchmark. The two-zone and three-zone techniques are proved to be better than CSVPWM, both in terms of THD as well as switching losses. While the distortion due to two-zone and three-zone PWM are comparable, the former results in a significantly reduced switching loss at high lagging power factors among the techniques considered. Hence, two-zone PWM technique is well-suited for motor drive applications.

8. REFERENCES


