Novel Switching Sequences for a Space Vector Modulated Three-Level Inverter

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Abstract— Conventional space vector pulsewidth modulation (CSVPWM) for a three-level inverter employs conventional switching sequence, which divides the zero vector time equally between the two zero states in every subcycle. This paper proposes a new PWM method for a three-level inverter, where one phase switches twice, while a second phase switches once and the third phase is clamped. For a given inverter average switching frequency, the proposed PWM method results in a significant reduction in harmonic distortion at high modulation indices. This is demonstrated theoretically as well as experimentally. The novel sequences can be employed at high modulation indices, while the conventional sequence can be used in the lower range of modulation index.

Index Terms— Harmonic distortion, multilevel inverter, pulsewidth modulation, space vector, stator flux ripple, voltage source inverter.

I. INTRODUCTION

Nowadays three-level voltage source inverters (VSI) are increasingly employed for dc-ac power conversion. A three-level diode-clamp inverter, shown in Fig. 1, has numerous advantages over a conventional two-level inverter. One of the advantages is the improved output waveform quality or reduction in total harmonic distortion (THD) in the line current, compared to a two-level inverter, under similar operating conditions [1]. With devices of the same voltage rating, a three-level inverter can handle a dc bus voltage roughly twice that of a two-level inverter. Hence this topology is favoured in medium voltage application [2]-[3]. It also provides better performance in terms of semiconductor losses, filtering arrangement and common mode voltage over two-level converter in low-voltage applications with medium to high switching frequencies [1].

The quality of waveforms produced by a voltage source inverter strongly depends on the pulse width modulation (PWM) method employed. Sine-triangle PWM (SPWM) and conventional space vector PWM (CSVPWM) are popular modulation methods for a two-level inverter [4]. These methods have been variously extended to three-level inverter [5]-[7].

This paper proposes a new PWM method for a three-level inverter. Compared to conventional space vector PWM, this method reduces the THD in the line current by about 30 percent at high modulation indices, as shown by theoretical as well as experimental results.

II. TWO LEVEL VOLTAGE SOURCE INVERTER

A two-level inverter has eight states, comprising of two zero states (+ + + and ---) and six active states. The active voltage vectors are of magnitude equal to DC bus voltage (Vdc), and the zero voltage vector is of magnitude zero as shown in Fig. 2.

For a voltage reference vector of magnitude Vref and angle α in sector –I, as shown in Fig. 2, the active vector V1, the active vector V2 and the zero vector V0 should be applied for durations T1, T2 and T0, respectively, as given by (1).

\[
T_1 = \frac{V_{\text{ref}}}{V_{dc}} \cdot \sin(60^\circ - \alpha) \cdot T_{*} \\
T_2 = \frac{V_{\text{ref}}}{V_{dc}} \cdot \sin(\alpha) \cdot T_{*} \\
T_0 = T_{*} - (T_1 + T_2) 
\]

Fig. 2 Vector diagram of a two-level inverter. I, II, III, IV, V and VI are sectors.
When a two-level inverter is switched using CSVPWM, the switching sequence is as shown in Fig. 3(a). The inverter state changes in a sequence, starting from one zero state and ending with the other zero state in a half carrier cycle or subcycle. The two zero states are applied for an equal duration of time, i.e. 0.5T. In case of SPWM, the two zero states are applied for different durations of time. However, the sequence is still the same.

Alternative switching sequences have been proposed recently for two-level inverters [8], [9]. Among these novel switching sequences, the two shown in Fig. 3(b) and Fig. 3(c) are noteworthy, since these lead to a substantial reduction in THD at high modulation indices [8].

In sequence 0127 (as in SPWM and CSVPWM), every phase switches once in a subcycle. However, with the sequences 0121 and 7212 [Fig. 3(b) and Fig. 3(c)], one phase switches twice, while a second phase switches once and the third phase is clamped [8].

This paper identifies and investigates switching sequences for a three-level inverter, which are equivalent to the sequences 0121 and 7212. The equivalent sequences are readily identified by reducing a three-level inverter into an equivalent two-level inverter, as brought out in the following section.

III. SWITCHING SEQUENCES FOR A THREE-LEVEL INVERTER

A three-level inverter has twenty seven states with three zero states (+ + +, 000 and ---) and twenty four active states as illustrated in Fig. 4. The zero states produce a zero vector, while the twenty four active states produce eighteen distinct voltage vectors as shown in Fig. 4. Among the active vectors, six voltage vectors (V1-V6) have a magnitude of 0.5Vdc, other six voltage vectors (V7-V12, V13-V18) are of magnitude of 0.866Vdc, while the rest of the voltage vectors (V19, V21, V23, V25 and V27) have magnitudes equal to Vdc as shown in Fig. 4.

In general, a phase is switched only between positive dc bus (P) and dc bus mid point (O) during its positive half cycle and only between ‘O’ and negative dc bus (N) during its negative half cycle in a three-level inverter [7]. So a fundamental three-phase cycle can be divided into six such regions. In each region, the three-level inverter can be viewed as an equivalent two-level inverter. This equivalent two-level inverter has six active vectors and a zero vector, as shown in Fig. 2, but of magnitude 0.5Vdc.

Fig. 5 shows the equivalent two-level vector diagram of a three-level inverter in the hextant centered around the active vector V1. Now the vector V1 can be viewed as the zero vector of the equivalent two-level inverter. The vector V1 can be subtracted from the reference vector (Vref) for the three-level inverter to obtain the reference vector (Vref) corresponding to the two-level inverter.

The reference vector can be generated by applying the three nearest voltage vectors for appropriate durations of time that will result in volt-second balance as shown by (2).

$$ V_p T_1 = V_7 T_7 + V_8 T_8 + V_1 T_1 $$

where V7, V8 and V1 are the three nearest voltage vectors to the reference vector Vp. The dwell times T7, T8 and T1 of these vectors can be calculated in a fashion similar to that of two-level inverter as shown in (3).
Consider the reference vector \( V_{\text{ref}} \) for the three-level inverter or the equivalent reference vector \( V_{\beta} \) for the two-level inverter as shown in Fig. 5. Now the three-level inverter states 0--, +--, +0- and +00 correspond to the two-level inverter states of 0, 1, 2 and 7, respectively.

Hence, sequence 0--, +--, +0-, +00 corresponds to sequence 0127. Similarly, sequences 0--, +--, +0- and +00 correspond to the novel sequences 0121 and 7212, respectively. For convenience, these three sequences for a three-level inverter will henceforth simply be indicated as 0127, 0121 and 7212. The following sections investigate the use of these novel switching sequences for a three-level inverter.

IV. ANALYSIS OF CURRENT RIPPLE

In a PWM inverter fed drive, there is an error between the instantaneous applied voltage vector and the reference voltage vector. Fig. 6 shows the error voltage vectors, corresponding to the applied voltage vectors \( V_7 \), \( V_8 \) and \( V_1 \) for a reference vector \( V_{\text{ref}} = 0.7 \angle 10^\circ \). The error voltage vectors can be expressed as:

\[
\begin{align*}
V_{\text{err},7} &= V_7 - V_{\text{ref}} = V_7 - (V_{\beta} + V_1) \\
V_{\text{err},8} &= V_8 - V_{\text{ref}} = V_8 - (V_{\beta} + V_1) \\
V_{\text{err},1} &= V_1 - V_{\text{ref}} = -V_{\beta}
\end{align*}
\]

The error voltages see the motor as an inductance equal to its total leakage inductance. Hence there is a proportional relationship between the integral of error voltage and the ripple current. The time integral of the error voltage vector is defined as the ‘stator flux ripple vector’, which is a measure of line current ripple [9]-[10].

When sequence 0--, +--, +0-, +00 (or sequence 0127) is employed, the trajectory of the tip of the stator flux ripple vector \( (\psi) \) is as shown in Fig. 6. The magnitude of the stator flux ripple is zero both at the start and the end of the subcycle. The trajectory of the tip of the vector is triangular as seen.

The stator flux ripple vector can be resolved along two orthogonal axes. The orthogonal axes chosen here are the x-axis and the y-axis indicated in Fig. 6. The x-axis is aligned with \( V_{\beta} \), the reference vector for the equivalent two-level inverter. The y-axis is perpendicular to the x-axis as shown. This choice of axes simplifies calculation and is similar to the q-axis and d-axis in the analysis of PWM for two-level inverter [8]. The components of the stator flux ripple vector \( (\psi) \) along the x-axis and y-axis are shown in Fig. 7(b) for sequence 0127.

Fig. 7 compares the trajectories of the tips of the stator flux ripple vectors, corresponding to sequences 0127, 0121 and 7212, for the given reference vector \( V_{\text{ref}} = 0.7 \angle 10^\circ \).

The corresponding x-axis and y-axis components are also shown, where the quantities \( X_0, X_1, X_2 \) and \( Y \) are as defined below.

\[
\begin{align*}
X_0 &= [0.5V_{dc} \cos(\beta) - V_{\beta}] * T_i \\
X_1 &= [0.5V_{dc} \cos(60^\circ - \beta) - V_{\beta}] * T_i \\
X_2 &= -V_{\beta} * T_i \\
Y &= 0.5V_{dc} \sin(\beta) * T_i
\end{align*}
\]

Fig. 6 Error voltage vectors corresponding to active vector 1 (\( V_7 \)), active vector 2 (\( V_8 \)) and zero vector (\( V_1 \)), and stator flux ripple vector for sequence 0127.
Observe that $\psi_x$ and $\psi_y$ are piecewise linear functions of time. When squared, these become piecewise parabolic functions of time. Hence the mean square values of these can easily be computed using the geometric formulae for area under parabolic section [8]-[10]. The mean square value of the stator flux ripple over a subcycle is the sum of the mean square values of its x-component and y-component as shown in (6) for sequence 0120. The mean square values correspond to sequences 0121 and 7212 can similarly be expressed in terms of $X_0$, $X_1$, $X_2$ and $Y$.

$$\psi_{0127}^2 = \frac{1}{3} \frac{1}{T} \left[ \psi_x^2 + (0.5X_0)^2 \psi_x + (0.5X_0)^2 \psi_y \right] + \frac{1}{3} \frac{1}{T} \left[ \psi_y^2 + (0.5X_0)^2 \psi_y + (0.5X_0)^2 \psi_x \right]$$

The total rms line current ripple over the entire range of modulation can be calculated using the total rms harmonic distortion factor based on the notion of stator flux ripple ($F_{DIST}$), which is independent of machine parameter [8]-[10]. The distortion factor is the rms stator flux ripple of a sector, normalized with respect to the fundamental flux $\psi_1$, as defined in (7).

$$F_{DIST} = \frac{1}{\psi_1} \sqrt{\frac{6}{\pi} \int_0^{\pi/6} F_{SEQ} d\alpha} \quad (7a)$$

$$\psi_1 = \frac{V_{ref}}{2\pi f_1} \quad (7b)$$

A theoretical comparison of the harmonic distortion among sequence 7212, sequence 0121 and sequence 0127 (CSVPWM techniques), assuming $V/f$ ratio to be fixed at its rated value, is presented in Fig. 8. The figure shows that there is a significant reduction in current ripple at higher frequency range (47.5-50 Hz) due to 0121 and 7212. The percentage of improvement in THD in line current due to sequence 7212 is 27.5 and that due to sequence 0121 is 29.3 at full modulation index of linear zone.

V. EXPERIMENTAL RESULTS

The above mentioned three sequences are evaluated in a 10 kVA inverter fed 2.2 kW, 415 V, 50 Hz induction motor drive. The PWM techniques are implemented on a TMS320LF2407A DSP controller.

Fig. 9 shows the harmonic spectra of line voltage waveforms at a fundamental frequency of 50 Hz and an average switching frequency of 1.5 kHz corresponding to sequence 0127. It is seen that the sidebands are around 1.5 kHz and 3 kHz. The dominant harmonic components are around 3 kHz.

Fig. 10 and Fig. 11, respectively, present the measured harmonic spectra of line voltage waveforms corresponding to novel sequences 7212 and 0121 at the same fundamental frequency and average switching frequency as in Fig. 9. There are no significant harmonic components around 1.5 kHz. The dominant harmonic components are around 3 kHz. Since the harmonic voltages are at higher frequencies, these are effectively filtered by the motor leakage inductance. Hence
the distortion in current waveform is better with these sequences than with 0127 as brought out by Fig. 12 to Fig. 14.

Fig.9 Measured harmonic spectra of line voltage at 50 Hz with sequence 0127 (scale: x axis-80v/div. y axis- 1kHz/div.)

Fig.10 Measured harmonic spectra of line voltage at 50 Hz with sequence 7212 (scale as in Fig.9)

Fig.11 Measured harmonic spectra of line voltage at 50 Hz with sequence 0121 (scale as in Fig.9)

Fig.12 Measured line current waveform at 50 Hz with sequence 0127 (scale: 2A/div)

Fig.13 Measured line current waveform at 50 Hz with sequence 7212 (scale: 2A/div)

Fig.14 Measured line current waveform at 50 Hz with sequence 0121 (scale: 2A/div)

The experimental no-load current waveforms, corresponding to the spectra in Figs. 9 to 11, are shown in Figs. 12 to 14, respectively.
The total harmonic distortion factor of the no-load current ($I_{THD}$) is defined in (8), where $I_1$ is the RMS fundamental no-load current and $I_n$ is the RMS $n^{th}$ harmonic current.

$$I_{THD} = \frac{1}{I_1} \sum_{n=1}^{\infty} \frac{I_n^2}{I_1^2}$$ (8)

The measured $I_{THD}$ values corresponding to different sequences are plotted against fundamental frequency in Fig. 15. Compared to the conventional sequence 0127, the novel sequences 0121 and 7212 result in lower THD at fundamental frequencies in the range 48 to 50 Hz. The experimental results agree well with the theoretical prediction in Fig. 8.

![Fig.15 Measured THD in line current corresponding to three sequences](image)

VI. CONCLUSION

Switching sequences, involving division of active vector time, are considered and compared with the conventional sequence for a three level diode clamp inverter. The maximum line voltage obtainable for a given dc bus voltage with the proposed technique is as high as conventional space vector modulation. For a given inverter average switching frequency, the proposed two sequences result in significant reduction (around 30 percent) in harmonic distortion at higher speeds of the induction motor drive. Hence conventional sequence (0127) can be used in lower range of fundamental frequency (below 48 Hz) and the novel sequence (0121 and 7212) can be employed at the fundamental frequency of 48 Hz and above range. Experimental results on a 2.2 kW induction motor drive are presented.

REFERENCES

