PWM Technique for NPC Inverters by Means of Decoupled Control of Voltage Vector Components

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A new control method for Neutral-Point-Clamped (NPC) inverters based on closed-loop regulation of the line-to-line output voltages is proposed. The method uses independent hysteresis comparator controllers to regulate the direct and quadrature axis components of the three phase output voltages. The closed-loop control allows a high performance over the whole range of operation, even when low speed devices such as the GTO are used. A neutral-point potential control is described, which is capable of stabilizing the variations within fixed limits during steady and transient states. Further, a new vector selection technique which used four vectors to generate the sinusoidal output is investigated. The vector selection allows an almost decoupled control of the vector components. The principle of the method is discussed and the vector selection technique is presented. The effectiveness in the output-voltage-waveform generation and the balance of the DC-link capacitor voltages are verified by simulation and experiment.

Key words: Neutral-point-clamped inverter, Voltage vector, Decloupled control, Neutral voltage control

1. Introduction

In recent years, there has been great interest in the Neutral Point Clamped Pulse Width Modulated (NPC-PWM) inverters. The NPC inverter is particularly attractive in high power applications because the blocking voltage of each switching device is half that of the DC voltage and its output waveforms may contain less harmonics than that of conventional full-bridge inverters (1), (2).

The NPC inverters have an inherent problem of neutral point potential variations. Therefore, analysis of the neutral voltage variations is mandatory and a suppression method should be implemented (3). On the other hand, since the NPC inverter is usually aimed to high power applications, it is important to consider the effect of the minimum pulse width, and the maximum available switching frequency of slow switching devices as the GTOs (4)-(6). Moreover, as most control methods have adopted the carrier type PWM control, techniques to achieve smooth switching between synchronous and asynchronous PWM method are reported (7), (8).

This paper proposes a new control method for NPC inverters based on closed-loop regulation of the instantaneous line-to-line voltages. Instantaneous output voltages are sensed and transformed to synchronously rotating reference coordinates. Both the direct and quadrature axis components of the transformed voltages are controlled independently by using hysteresis comparators. Variations of the neutral potential are restrained into certain limits by a hysteresis controller included in the control circuit. The closed-loop control allows a high-quality output voltage and smooth regulation of the control variables over the whole range of operation even when low-speed devices are used. Inherent to the closed-loop control, the method includes automatic compensation for DC-link voltage variations. Other properties of the method are simple configuration wider control range and fast response (9)-(11).

Any vector selection can be applied to the proposed control method. PWM techniques for NPC inverters commonly use three vectors to generate the voltage waveform. In this case the switching levels of the line-to-line voltages change from (0, E_d/2) to (E_d/2, E_d) in a certain angle (exchanging-point) which is related to the input reference and the DC source voltage.

This paper proposes a new vector selection. The output waveform is generated by switching among four vectors which are interchanged sequentially every \( \pi/6 \) period. The use of four vectors not only simplifies the control circuit but also improves the waveform generation around the exchanging point. Lastly, with the proposed vector selection, an almost decoupled control of the vector components can be executed.

This paper discusses the principles of the control strategy and the vector selection. The neutral point voltage variations are considered and a method to reduce the variation is...
Fig. 1. Neutral-Point-Clamped Inverter.

The proposed control system is explained in detail and some considerations for the controllers' design are given. Computer simulations and experimental results with a low power prototype are presented.

2. Control strategy

Fig. 1 shows the circuit configuration of the NPC inverter. Each arm of the inverter consists of four switching devices and produces three switching states according to Table 1. The combination of the switching states of the three arms generates the 27 NPC inverter voltage vectors. The NPC vectors are shown in Fig. 2. They are classified according to their amplitude. Thus, a-group corresponds to the large vectors; b-group refers to the medium size vectors; c-group and d-group categorize the small vectors. The last group (z) is composed of zero vectors. For balance conditions, vectors of c and d groups generate the same output waveform.

The mathematical description of the PWM technique uses space vectors defined in a coordinate system rotating with the angular speed \( \omega = \omega t \) as shown in Fig. 3.

The coordinate transformation is given by Eq. (1).

\[
\begin{bmatrix}
    e_p \\
    e_q
\end{bmatrix} = \frac{2}{3} \begin{bmatrix}
    \sin(\theta) & \sin(\theta + 2\pi/3) & \sin(\theta - 2\pi/3) \\
    \cos(\theta) & \cos(\theta + 2\pi/3) & \cos(\theta - 2\pi/3)
\end{bmatrix} \begin{bmatrix}
    e_{uv} \\
    e_{vw} \\
    e_{wu}
\end{bmatrix}
\]

Where \( e_{uv}, e_{vw}, e_{wu} \) represent the line to line voltages and the term \( (e_p, e_q) \) represent the instantaneous voltage vector in the \( P, Q \) coordinates. The advantage of this model is that sinusoidal waveforms in balanced conditions are represented as DC quantities:

\[
\sqrt{2} A \begin{bmatrix}
    \sin(\theta + \phi) \\
    \sin(\theta + 2\phi / 3) \\
    \sin(\theta - 2\phi / 3)
\end{bmatrix} = \begin{bmatrix}
    e_p \\
    e_q
\end{bmatrix} = A \begin{bmatrix}
    \cos(\phi) \\
    \sin(\phi)
\end{bmatrix}
\]

In particular the angle \( \phi \) can be equal to zero. In this case, the quadrature component \( e_q \) is equal zero and direct axis component \( e_p \) is equal to the RMS value of the sinusoidal voltage.

For non sinusoidal or unbalanced voltages, both \( e_p \) and \( e_q \) are subdivided into two quantities. One does not depend on \( \theta \) called the DC component and the other which is a time-dependent variable called the ripple component. The principle is focused on minimizing the ripple components by adequate selection among the available vectors.

\[
\begin{bmatrix}
    e_p \\
    e_q
\end{bmatrix} = \begin{bmatrix}
    e_p (DC) + e_p (ripple) \\
    e_q (DC) + e_q (ripple)
\end{bmatrix}
\]

Table 1. Switching states.

<table>
<thead>
<tr>
<th>Switching states</th>
<th>( S_{X1} )</th>
<th>( S_{X2} )</th>
<th>( S_{X3} )</th>
<th>( S_{X4} )</th>
</tr>
</thead>
<tbody>
<tr>
<td>P</td>
<td>ON</td>
<td>ON</td>
<td>OFF</td>
<td>OFF</td>
</tr>
<tr>
<td>O</td>
<td>OFF</td>
<td>ON</td>
<td>ON</td>
<td>OFF</td>
</tr>
<tr>
<td>N</td>
<td>OFF</td>
<td>OFF</td>
<td>ON</td>
<td>ON</td>
</tr>
</tbody>
</table>
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Table 2. Transformed NPC inverter vectors.

<table>
<thead>
<tr>
<th>Group</th>
<th>Vector</th>
<th>Vector components</th>
</tr>
</thead>
<tbody>
<tr>
<td>a</td>
<td>1 (PPN) 2 (PPN)</td>
<td>( e_p = \frac{\sqrt{3}}{3} E_d \sin \left( \theta \cdot \frac{(2n-1)\pi}{6} \right) )</td>
</tr>
<tr>
<td></td>
<td>3 (NPN) 4 (NPP)</td>
<td>( e_q = \frac{\sqrt{3}}{3} E_d \cos \left( \theta \cdot \frac{(2n-1)\pi}{6} \right) )</td>
</tr>
<tr>
<td>b</td>
<td>1 (POP) 2 (OPN)</td>
<td>( e_p = \frac{\sqrt{3}}{2} E_d \sin \left( \theta \cdot \frac{\pi}{3} \right) )</td>
</tr>
<tr>
<td></td>
<td>3 (NOP) 4 (ONP)</td>
<td>( e_q = \frac{\sqrt{3}}{2} E_d \cos \left( \theta \cdot \frac{\pi}{3} \right) )</td>
</tr>
<tr>
<td>c</td>
<td>1 (POO) 2 (PPO)</td>
<td>( e_p = \frac{\sqrt{3}}{3} e_q \sin \left( \theta \cdot \frac{(2n-1)\pi}{6} \right) )</td>
</tr>
<tr>
<td></td>
<td>3 (OPO) 4 (OPP)</td>
<td>( e_q = \frac{\sqrt{3}}{3} e_p \cos \left( \theta \cdot \frac{(2n-1)\pi}{6} \right) )</td>
</tr>
<tr>
<td>d</td>
<td>1 (ONN) 2 (OON)</td>
<td>( e_p = \frac{\sqrt{3}}{3} e_q \sin \left( \theta \cdot \frac{(2n-1)\pi}{6} \right) )</td>
</tr>
<tr>
<td></td>
<td>3 (NON) 4 (NOO)</td>
<td>( e_q = \frac{\sqrt{3}}{3} e_p \cos \left( \theta \cdot \frac{(2n-1)\pi}{6} \right) )</td>
</tr>
<tr>
<td>z</td>
<td>(PPP) (OPO) (NNN)</td>
<td>( e_p = 0 ), ( e_q = 0 )</td>
</tr>
</tbody>
</table>

The \( e_p \) and \( e_q \) components of the transformed NPC vectors are listed in Table 2. Fig. 4 shows the waveforms of the transformed vectors for the interval \([0, \pi/3]\). For small values in the input reference, the best choice is to select vectors \( e_5/d_5, e_5/d_5, e_5/d_5 \) (shown with bold lines in Fig. 4), because they produce the smallest step changes in both \( e_p \) and \( e_q \) components. In mode-low operation, the instantaneous vector is chosen using the selection illustrated in Table 3; where \( E_{pr} \) correspond to the voltage reference, \( h_p \) is the hysteresis limit of the \( e_p \) controller, and \( h_q \) is the hysteresis limit of the \( e_q \) controller. Since only vectors of \( c, d \) and \( z \) groups are used, three kinds of voltage levels (\( E_d/2, 0, -E_d/2 \)) exists in the line-to-line output voltage. Therefore, the waveform of the output voltage will be restrained to the shaded area of Fig. 5.

Equations (4)-(6) define the control operation for mode-low. The average value of \( e_p \) and \( e_q \) (\( \bar{e}_p \), \( \bar{e}_q \)) are controlled to be equal to \( E_{pr} \) and zero respectively.

\[
\bar{e}_p = \alpha_1 \left( \frac{\sqrt{3}}{2} E_d \sin \theta \right) + \alpha_2 \left( \frac{\sqrt{3}}{2} E_d \sin \left( \theta - \frac{\pi}{6} \right) \right) + \alpha_3 \left( \frac{\sqrt{3}}{2} E_d \sin \left( \theta + \frac{\pi}{6} \right) \right) + \alpha_4 [0] = E_{pr}
\]

(4)

\[
\bar{e}_q = \alpha_1 \left( \frac{\sqrt{3}}{2} E_d \cos \theta \right) + \alpha_2 \left( \frac{\sqrt{3}}{2} E_d \cos \left( \theta - \frac{\pi}{6} \right) \right) + \alpha_3 \left( \frac{\sqrt{3}}{2} E_d \cos \left( \theta + \frac{\pi}{6} \right) \right) + \alpha_4 [0] = 0
\]

(5)

\[
\alpha_1 + \alpha_2 + \alpha_3 = 1
\]

(6)

Here, \( \alpha_1, \alpha_2, \alpha_3 \) are the dwell time of \( e_5/d_5, e_5/d_5 \) and \( z_5/z_5/z_5 \) respectively. Having solved Eqs. (4)-(6), we find the expressions for the dwell-time of these vectors as:

\[
\alpha_1 = \frac{4}{V_2} \frac{E_{pr} \cos \theta \cdot \frac{\pi}{6}}{E_d}
\]

(7)

\[
\alpha_2 = \frac{4}{V_2} \frac{E_{pr} \sin \theta \cdot \frac{\pi}{6}}{E_d}
\]

(8)

Table 3. Vector selection for mode-low.

<table>
<thead>
<tr>
<th>( 0 \leq \theta \leq \frac{\pi}{3} )</th>
<th>( \bar{e}_p &gt; h_p )</th>
<th>( \bar{e}_q &lt; -h_q )</th>
</tr>
</thead>
<tbody>
<tr>
<td>( \bar{e}<em>p - E</em>{pr} &gt; h_p )</td>
<td>( \bar{e}_q &lt; -h_q )</td>
<td></td>
</tr>
<tr>
<td>( \bar{e}<em>p - E</em>{pr} &lt; h_p )</td>
<td>( \bar{e}_q &lt; -h_q )</td>
<td></td>
</tr>
</tbody>
</table>

Fig. 4. Transformed NPC vectors: (a) \( e_p \) component; (b) \( e_q \) component.

Fig. 5. Line-to-line voltage shape for mode-low operation.

Table 3. Vector selection for mode-low.
Table 4. Vector selection for mode-high.

<table>
<thead>
<tr>
<th>Interval</th>
<th>$\delta_p$</th>
<th>$\delta_q$</th>
<th>$\delta_q &gt; h_q$</th>
<th>$\delta_q &lt; -h_q$</th>
</tr>
</thead>
<tbody>
<tr>
<td>$0 \leq \theta \leq \pi/6$</td>
<td>$\delta_p - E_{pr} &gt; h_p$</td>
<td>$\delta_q/d_5$</td>
<td>$\delta_q/d_6$</td>
<td></td>
</tr>
<tr>
<td>$\pi/6 \leq \theta \leq \pi/3$</td>
<td>$\delta_p - E_{pr} &lt; -h_p$</td>
<td>$a_5$</td>
<td>$b_5$</td>
<td></td>
</tr>
<tr>
<td>$\pi/3 \leq \theta \leq \pi$</td>
<td>$\delta_p - E_{pr} &gt; h_p$</td>
<td>$\delta_q/d_5$</td>
<td>$\delta_q/d_6$</td>
<td></td>
</tr>
<tr>
<td>$\pi/3 \leq \theta \leq \pi$</td>
<td>$\delta_p - E_{pr} &lt; -h_p$</td>
<td>$b_5$</td>
<td>$a_6$</td>
<td></td>
</tr>
</tbody>
</table>

Since the dwell time cannot be smaller than zero the control region of this vector selection becomes:

$$\alpha_3 = 1 - \frac{4}{\sqrt{2}} \frac{E_{pr}}{E_d} \sin \left[ \theta + \frac{\pi}{3} \right] \quad 0 \leq \theta \leq \frac{\pi}{3}$$

Equations (11) and (12) define the control operation for mode-high during the interval $[0, \pi/6]$.

$$\alpha_3 = 1 - \frac{4}{\sqrt{2}} \frac{E_{pr}}{E_d} \sin \left[ \theta + \frac{\pi}{3} \right] \quad 0 \leq \theta \leq \frac{\pi}{3}$$

For higher values in the reference (mode-high operation), the control variables are regulated by switching among vectors of groups $a$, $b$, $c$, $d$. Vectors $a$, $b$, $c$, $d$ are used during the interval $[0, \pi/6]$ and vectors $a$, $b$, $c$, $d$ during the interval $[\pi/6, \pi/3]$ as shown in Fig. 6. The instantaneous vector selection is given in Table 4. The output-voltage waveform for high input references will be restrained to the shaded area of Fig. 7.

Fig. 7. Line-to-line voltage shape for mode-high operation.

Here $\beta_1 - \beta_4$ refers to the dwell time of the associated vector. An additional condition is given by the instantaneous vector selection of Table 4. The average value of $\delta_q$ can be controlled to be zero by switching between $a_5$ and $b_5$ for the condition $\delta_p - E_{pr} < -h_p$. For the condition $\delta_p - E_{pr} > h_p$ the $\delta_q$ component is controlled by switching among $c_5/d_5$ and $d_6$. Equations (13) and (14) describe the control operation of the $\delta_q$ component. Since $\delta_q$ can successfully be controlled no matter the instantaneous value of $\delta_p$, an almost decoupled control is accomplished.

$$\beta_1 \left[ \frac{\sqrt{6}}{3} \frac{E_d}{\sin \left( \theta + \frac{\pi}{3} \right)} \right] + \beta_2 \left[ \frac{\sqrt{2}}{2} \frac{E_{pr}}{\sin \left( \theta - \frac{\pi}{3} \right)} \right] = 0$$

$$\beta_3 \left[ \frac{\sqrt{6}}{3} \frac{E_d}{\sin \left( \theta + \frac{\pi}{3} \right)} \right] + \beta_4 \left[ \frac{\sqrt{2}}{2} \frac{E_{pr}}{\sin \left( \theta - \frac{\pi}{3} \right)} \right] = 0$$

The conditions of the dwell time ($\beta_1 - \beta_4 \neq 0$) define the validity region of this vector selection for this interval. Following analogous procedure for the other intervals the control region of mode-high operation can be determined as:

$$\frac{\sqrt{2}}{4 \cos \left( \theta - (2n-1) \pi/6 \right)} \frac{E_d}{\sin \left( \theta + \frac{\pi}{3} \right)} \leq E_{pr} \leq \frac{\sqrt{2}}{4 \cos \left( \theta - (2n-1) \pi/6 \right)}$$

Where $(n-1) \pi/3 \leq \theta \leq n \pi/3$ and $n = 1, 2, ..., 6$.

3. Neutral point voltage control

The neutral terminal is not a rigid voltage source. Thus, it is not certain that the DC-link capacitor voltages $e_{d1}$ and $e_{d2}$ would remain equal. Variations in the neutral point potential take place only when a vector of $b$, $c$, or $d$ groups is applied to the output. The increment or decrement in the neutral voltage depends on the selected vector, the load current, the pulse

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Capacitor $C_2$ discharges during generative conditions when a vector of c-group is used. In this case, voltage $e_{d1}$ decreases and voltage $e_{d2}$ increases as shown in Fig. 8 (a). In regenerative condition, as the current flows in the opposite direction, a contrary effect in the voltage variation also occurs. Similar action appears when a d-group vector is applied to the output (Fig. 8 (b)), but in this case $e_{d2}$ decreases during generative condition and increases during regenerative condition.

Since vectors of c and d groups produce the same output voltage respectively, interchanging between these two groups does not alter the output generation. Therefore, in generative condition, $(i_s > 0)$ variations can be minimized by selecting a c-group vector if $e_{d1} > e_{d2}$ and the respective d-group vector if $e_{d2} > e_{d1}$. During regenerative condition $(i_s < 0)$ an opposite selection is required. It is easy to add a hysteresis comparator to carry out this action. The operation of this controller is summarized in Table 5, where $h_0$ refers to the hysteresis-width of the controller.

4. Control system

The control system is indicated in Fig. 9. The line-to-line output voltages are sensed by isolation amplifiers and transformed to the P-Q reference coordinates ($e_p$ and $e_q$) using a multiplier type A/D converter. Both components are controlled independently by hysteresis comparators. Beforehand, the signals are time delayed by means of an integrator to control the average value of $e_p$ and $e_q$ with finite switching frequency. The hysteresis width of the comparators ($h_p$ and $h_q$) and the integrator time constant determine the accuracy of the waveform generation.

An additional controller is used to allow the commutation from mode-high to mode-low and vice versa. The hysteresis width of this controller ($H_p$) is wider than that of the $e_p$ controller ($H_p > h_p$), and its operation is illustrated in Table 6.

A hysteresis comparator controller ($h_n$) is used to balance the DC capacitor voltages. The direction of the source current ($i_s$) determines whether is generative or regenerative condition. The information of the working interval, given by $\theta_n$, completes the input data to the EPROM table that generates the command signals.

The design of the hysteresis width of the controllers and the integrator time constant is a trade-off between the accuracy of the waveform generation and the available switching frequency of the inverter elements. On the other hand, to avoid variations on the average value of the switching frequency with changes in the reference and/or the DC source voltage, the hysteresis width of the $e_p$ and $e_q$ controllers can be adjusted as a function of the relation $E_{pr}/E_d$. The effect of the dead time of the devices used can also be compensated adjusting the hysteresis width of the controllers.

Table 5. Operation of the neutral voltage controller.

<table>
<thead>
<tr>
<th>$h_0$ Control</th>
<th>$i_s &gt; 0$</th>
<th>$i_s &lt; 0$</th>
</tr>
</thead>
<tbody>
<tr>
<td>$e_{d1} - e_{d2} &gt; h_0$</td>
<td>c-group</td>
<td>d-group</td>
</tr>
<tr>
<td>$e_{d1} - e_{d2} &lt; -h_0$</td>
<td>d-group</td>
<td>c-group</td>
</tr>
</tbody>
</table>

Table 6. Selection of the operation mode.

<table>
<thead>
<tr>
<th>$H_p$ Control</th>
<th>Mode of operation</th>
</tr>
</thead>
<tbody>
<tr>
<td>$\hat{e}<em>p, E</em>{pr} &gt; H_p$</td>
<td>Mode-high</td>
</tr>
<tr>
<td>$\hat{e}<em>p, E</em>{pr} &lt; -H_p$</td>
<td>Mode-low</td>
</tr>
</tbody>
</table>
5. Simulation results

In this section, simulation results of the proposed control technique on an induction motor drive working in a constant Volt/Hertz ratio are presented.

For the simulations, a 400-V, four-pole, 110-kW induction motor load was used. $E_d$ was of 600-V and the capacitors $C_1$ and $C_2$ were of 4000-$\mu$F. The commutation dead time was $t_d=50-\mu$s, and the minimum pulse width was set to $t_w=100-\mu$s to allow the use of the GTO. The switching frequency varies slightly from period to period but its average value was around 500-Hz which is normally used for GTO inverters.

Fig. 10 shows the operation for a frequency of 50-Hz (mode-high). The good quality of the line-to-line voltage ($e_{uv}$) and current ($i_u$) waveforms can be appreciated. It can be seen that the capacitor voltages $e_{d1}$ and $e_{d2}$ remain approximately equal to $E_d/2$. The voltage vector components $\vec{e}_p$ and $\vec{e}_s$ vary between the hysteresis limits. The $S_{11}$ and $S_{12}$ signals represent the state (on-off) of the switching devices $S_{11}$ and $S_{12}$ respectively (see Fig. 1); notice that both have been switched only a few times during the period.

Fig. 11 illustrates the simulation waveforms for mode-low. The operating frequency is 2-Hz. In spite of the low-speed devices used and the reduced reference operation, the motor current ($i_u$) is nearly sinusoidal. In this case the switching frequency is reduced automatically by the closed-loop to avoid distortion in output waveform. The superior quality of these waveforms is certified by the frequency spectrum of Fig. 12. Figs. 11 and 12 demonstrate that the system can operate successfully at very low references.

6. Experimental results

A low power prototype was constructed and tested in the laboratory. IGBT were used as the switching devices. A 200-V, four-pole, 0.75-kW induction motor was used. $E_d$ was of 280-V and the capacitors $C_1$ and $C_2$ were of 470-$\mu$F. The line-to-line voltage and the line current for mode-high ($f=50$-Hz) and mode-low ($f=10$-Hz) are shown in Figs. 13 and 14 respectively. The shape of the waveforms agrees with the theoretical analysis. The feasibility of the waveform and the good quality of the voltage and current signals are verified.

Fig. 15 illustrates the response for a ramp input reference to the constant Volt/Hertz drive. The reference ($E_{pr}$) increases...
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Fig. 14. Experimental waveforms for mode-low (f = 10-Hz).

Fig. 15. Response for a ramp reference (no-load conditions).

from 5% to 90% in 400-ms. The figure proves that smooth control can be obtained over the whole range of operation.

Fig. 16 shows the transient response for 30% disturbance applied to the DC-link voltage. No variation in the current waveform can be detected due to the fast response of the controllers.

The effect of the neutral point voltage control is illustrated in Fig. 17. Starting from balance conditions, suddenly the neutral voltage controller is deactivated. This produces an increase difference between \( e_{d1} \) and \( e_{d2} \). When the control is reactivated, the neutral voltage recovers the balance conditions in a relatively short period. The response depends on the capacitor value and the load conditions.

7. Conclusions

A control method for NPC-PWM inverters based on closed-loop regulation of the line-to-line voltages using independent hysteresis controllers was presented. Some characteristics of the control system are:

- Smooth control over the whole range of operation.
- High quality waveform at reduced references in spite of the slow switching devices used.
- Effective to reduce the variation of the neutral voltage.
- Good transient and steady-state performance.
- Allows an almost decoupled control of the voltage vector components.
- Increased control range of the output voltage, compared with conventional phase control techniques.

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References


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