Evaluation of Two Modulation Techniques for Two-level and Three-level Inverters in Application of Active Power Filters

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Abstract

One of the effective factors in performance of Active Power Filter (APF) is its modulation technique. In this paper, two modulation techniques namely, the Hysteresis Current Control Pulse Width Modulation (HCC PWM) and Carrier-based Pulse Width Modulation (CPWM), are analyzed for a shunt APF. The shunt APF is implemented with a Two-Level Voltage Source Inverter (2L-VSI) and Three-Level Neutral Point Clamped Voltage Source Inverter (3L-NPC VSI). The reference signals extraction technique is based on instantaneous reactive power theory (p-q theory) in order to eliminate harmonic components and compensate reactive power. Impressive parameters in switching frequency are designed carefully to achieve the same switching frequency for all the case studies. The analysis is based on respecting the same maximum switching frequency and results of simulations are compared through Total Harmonic Distortion (THD) value of source currents, dynamic performance and switching numbers of switches.

Keywords: Active power filter, hysteresis current control PWM, carrier-based PWM, switching frequency.

1. Introduction

In recent years, power quality issue in industrial utilities has become a subject of serious concern due to the intensive use of power electronic equipment [1-3]. In other words, nonlinear loads such as, arc furnaces, adjustable speed drives and power electronic
converters draw harmonic currents and reactive power from power supply networks. These non-sinusoidal currents have some drawbacks including, voltage distortion on the load side, losses in transmission lines and protective devices failures. Conventionally passive filters are used to solve these problems but, due to some disadvantages such as, tuning difficulties, aging problems, resonance and large size, it is preferred to use APF [4, 5]. APF is a very useful tool for eliminating harmonic pollution in power systems. The shunt APF is a device that is connected in parallel to the nonlinear loads for canceling their harmonic currents. It consists of a three-phase voltage inverter with current regulation, which is controlled as current sources for producing harmonic components generated by the load but, 180° out of phase. Different types of shunt and series APF have been used effectively but, shunt APF is more popular than series filter, because most of the industrial applications require current harmonics compensation [6, 7].

The performance of an APF basically depends on three parameters: a) the control method implemented to generate the reference signals, b) type of inverter topology used and c) the modulation technique. Several topologies of inverters have attracted attention and have been largely used in industrial application during the last years [8, 9].

In addition to inverter topologies, several modulation and control techniques have been developed to achieve some targets like, less switching losses, less THD in the spectrum of switching waveforms and fast dynamic response [10, 11]. Two popular modulation schemes CPWM and HCC PWM are considered and used for a shunt APF which is implemented with a 2L-VSI and 3L-NPC VSI as shown in Figure 1.

In this paper, section 2 explains the reference signals extraction strategy applied to simulated APF. Section 3 describes the principles of two mentioned modulation techniques. In section 4, the proposed system is simulated and results are discussed. Finally, section 5 presents conclusion of simulations.
2. Reference signals extraction technique

The reference signals extraction algorithm for APF is based on the instantaneous reactive power theory (p-q theory) which introduced for three-phase systems by Akagi in 1984 [12]. The p-q theory uses the $\alpha\beta0$ transformation also, known as the Clarke transformation which consists of matrix that, transforms three-phase voltages and currents into the $\alpha\beta0$ stationary reference frames. In this method, the instantaneous source voltages ($V_{sa}$, $V_{sb}$, $V_{sc}$) and load currents ($i_{La}$, $i_{Lb}$, $i_{Lc}$) are transferred from abc coordinates to the $\alpha\beta0$ coordinates as follows:

$$\begin{pmatrix} V_{\alpha}(\alpha t) \\ V_{\beta}(\alpha t) \end{pmatrix} = \sqrt{2 \over 3} \begin{pmatrix} 1 & -1 \over 2 & -1 \over 2 \\ 0 & \sqrt{3} \over 2 & -\sqrt{3} \over 2 \end{pmatrix} \begin{pmatrix} V_{sa}(\alpha t) \\ V_{sb}(\alpha t) \\ V_{sc}(\alpha t) \end{pmatrix}$$

(1)

$$\begin{pmatrix} i_{\alpha}(\alpha t) \\ i_{\beta}(\alpha t) \end{pmatrix} = \sqrt{2 \over 3} \begin{pmatrix} 1 & -1 \over 2 & -1 \over 2 \\ 0 & \sqrt{3} \over 2 & \sqrt{3} \over 2 \end{pmatrix} \begin{pmatrix} i_{La}(\alpha t) \\ i_{Lb}(\alpha t) \\ i_{Lc}(\alpha t) \end{pmatrix}$$

(2)

The active power and reactive power for three phase system are calculated as shown in Equation (3).

$$\begin{pmatrix} p(\alpha t) \\ q(\alpha t) \end{pmatrix} = \begin{pmatrix} V_{\alpha}(\alpha t) & V_{\beta}(\alpha t) \\ -V_{\beta}(\alpha t) & V_{\alpha}(\alpha t) \end{pmatrix} \begin{pmatrix} i_{\alpha}(\alpha t) \\ i_{\beta}(\alpha t) \end{pmatrix}$$

(3)

The instantaneous real power and reactive power, can be expressed as:

$$p(\alpha t) = \overline{p}(\alpha t) + \tilde{p}(\alpha t)$$

$$q(\alpha t) = \overline{q}(\alpha t) + \tilde{q}(\alpha t)$$

(4)

These two powers have constant values and oscillating components. Where $\overline{p}(\alpha t)$ and $\tilde{p}(\alpha t)$ represent the average and oscillating parts of $p(\alpha t)$ respectively, whereas $\overline{q}(\alpha t)$ and $\tilde{q}(\alpha t)$ represent the average and oscillating parts of $q(\alpha t)$.
The APF compensating currents \( (i_{F_A}^*, i_{F_B}^*, i_{F_C}^*) \) can be derived from the extracted oscillating power and reactive power supplied by the APF. Normally, a low pass filter is used for separating average and oscillating parts of power. To calculate the reference compensating currents, Equation (3) is inverted and the powers to be compensated \( (\bar{p}(\omega t), \bar{q}(\omega t)) \) are used as follows:

\[
\begin{pmatrix}
  i_{a}^*(\omega t) \\
  i_{b}^*(\omega t)
\end{pmatrix} = \frac{1}{V_a^2(\omega t) + V_b^2(\omega t)} \begin{pmatrix}
  V_a(\omega t) & -V_b(\omega t) \\
  V_b(\omega t) & V_a(\omega t)
\end{pmatrix} \begin{pmatrix}
  \bar{p}(\omega t) + \bar{P}_{\text{loss}}(\omega t) \\
  \bar{q}(\omega t)
\end{pmatrix}
\]

\( (5) \)

\[
\begin{pmatrix}
  i_{F_A}^*(\omega t) \\
  i_{F_B}^*(\omega t) \\
  i_{F_C}^*(\omega t)
\end{pmatrix} = \begin{pmatrix}
  1 & 0 \\
  -\frac{1}{\sqrt{3}} & \frac{\sqrt{3}}{2} \\
  \frac{1}{\sqrt{3}} & -\frac{\sqrt{3}}{2}
\end{pmatrix} \begin{pmatrix}
  i_{a}^*(\omega t) \\
  i_{b}^*(\omega t)
\end{pmatrix}
\]

\( (6) \)

\( \bar{P}_{\text{loss}}(\omega t) \) is the output of DC-link voltage controller. Since any fluctuation in this voltage will cause improper performance of inverters, the charging and discharging of DC-link capacitor should be in such a way that, make the DC-link voltages constant. To stabilize this voltage, a conventional PI controller is used. As demonstrated in Figure 2, PI controller compares the reference voltage and DC-link voltage to produce the proper control signals. The reference current extraction and control system block diagram are demonstrated in Figure 2.
Figure 1: Power system with APF

Figure 2: Control system of APF
3. Modulation Techniques

Several PWM techniques have been extended in the literature for controlling the active devices in converters. The most popular and easiest techniques to implement are CPWM and HCC PWM. This section, presents the considered control techniques applied to control the current injection of shunt APF.

3.1. Hysteresis Current Control PWM

Hysteresis band current control technique is basically an instantaneous feedback current control of PWM where, the actual current continually tracks the reference current within a hysteresis band. HCC PWM provides excellent dynamics response and high accuracy with minimum hardware [13, 14].

Figure 3: Switching function generation by HCC PWM

Figure 3 demonstrates the performance of HCC PWM. The difference between reference current \( i_{ref}(t) \) and actual current \( i_{act}(t) \) is error current \( e(t) \). The sign function \( u(t) \) is generated by crossing \( e(t) \) through hysteresis band controller. Finally, switching function \( V_f(t) \) is obtained by \( V_f(t) = u(t) \times V_{dc} \).

3.1.1. Two-Level Hysteresis Current Control PWM Scheme

Two-level HCC PWM technique compares the error current against fixed hysteresis bands whereas, actual current \( i_F \) tracks reference current \( i_F^* \) as shown in Figure 4. If the error current exceeds the upper limit of the hysteresis band \( +h \), the upper switch of the 2L-VSI arm will turn OFF and the lower switch will turn ON. If the error current crosses...
the lower limit of the hysteresis band (-\(h\)), the lower switch of the 2L-VSI arm will turn OFF and the upper switch will turn ON [15].

<table>
<thead>
<tr>
<th>Switching states</th>
<th>(V_{xo}) ((x=a, b, c))</th>
</tr>
</thead>
<tbody>
<tr>
<td>(S_{1x}) ON</td>
<td>(S_{2x}) OFF</td>
</tr>
<tr>
<td>(S_{1x}) OFF</td>
<td>(S_{2x}) ON</td>
</tr>
</tbody>
</table>

The output voltage across the points \(x (x = a, b, c)\) and node ‘o’ is given in Table 1 for various switch combinations.

**Figure 4: Currents tracking in two-level HCC PWM**

In order to calculate the switching frequency of HCC PWM method, ignoring the interface resistance, the following equations can be written in the switching intervals \(t_1\) and \(t_2\).
\[
\frac{di_F^+}{dt} = \frac{1}{L_f} (0.5V_{dc} - U_o) 
\]

(7)

\[
\frac{di_F^-}{dt} = -\frac{1}{L_f} (0.5V_{dc} + U_o) 
\]

(8)

\(U_o\) is the supply voltage. From Figure 4 we can get the following equations:

\[
t_1 \frac{di_F^-}{dt} - t_1 \frac{di_F^+}{dt} = 2h 
\]

(9)

\[
t_2 \frac{di_F^-}{dt} - t_2 \frac{di_F^+}{dt} = -2h 
\]

(10)

The switching period and switching frequency are as follows:

\[
t_1 + t_2 = T_s = \frac{1}{f_s} 
\]

(11)

Adding equation (9) and (10) and using equation (11) give the following equation:

\[
t_1 \frac{di_F^-}{dt} + t_2 \frac{di_F^-}{dt} - \frac{1}{f_s} \frac{di_F^+}{dt} = 0 
\]

(12)

Subtracting equation (10) and (9), gives:

\[
t_1 \frac{di_F^-}{dt} - t_2 \frac{di_F^-}{dt} - (t_1 - t_2) \frac{di_F^+}{dt} = 4h 
\]

(13)

Substituting equation (8) in (13), gives:

\[
(t_1 + t_2) \frac{di_F^+}{dt} - (t_1 - t_2) \frac{di_F^-}{dt} = 4h 
\]

(14)

Similarly, substituting equation (8) in (12), gives:

\[
\frac{di_F^+}{dt} = \frac{dt}{t_1 - t_2} 
\]

\[
\frac{di_F^-}{dt} = \frac{dt}{t_1 + t_2} 
\]

(15)

Finally, substituting equation (15) in (14), gives:
From the above equation, the switching frequency is the function of hysteresis bandwidth, DC-link voltage, supply voltage, slope of instantaneous compensation current and interface inductance.

### 3.1.2. Three-Level Hysteresis Current Control PWM Scheme

Three-level HCC PWM technique is set as upper and lower hysteresis band \((h)\) and inner bands called dead zone \((\delta)\) [16] as shown in Figure 4. Whenever, the current error crosses an outer hysteresis bands, \(u(t)\) set to +1 or -1 to force a reversal of the current error. Similarly, whenever the current error reaches inner hysteresis bands, \(u(t)\) is set to a zero condition and the current error will be forced to reverse direction without meeting the next outer boundary. If the selection of a zero output does not reverse the current error, it will continue through the inner boundary to the next outer hysteresis band. In each of the inner and outer bands, inverter’s switches will be commanded to reverse current.

<table>
<thead>
<tr>
<th>Switching states</th>
<th>(V_{x0}) ((x=a, b, c))</th>
</tr>
</thead>
<tbody>
<tr>
<td>S1x</td>
<td>S2x</td>
</tr>
<tr>
<td>ON</td>
<td>ON</td>
</tr>
<tr>
<td>OFF</td>
<td>ON</td>
</tr>
<tr>
<td>OFF</td>
<td>OFF</td>
</tr>
</tbody>
</table>

The output voltage across the points \(x (x = a, b, c)\) and node ‘o’ is given in Table 2 for various switch combinations.
The switching frequency of HCC PWM method can be calculated using the following equations in the switching intervals $t_1$ and $t_2$.

\[
\frac{di_F^+}{dt} = \frac{1}{L_f} (0.5V_{dc} - U_o) \tag{17}
\]

\[
\frac{di_F^-}{dt} = -\frac{1}{L_f} (0 + U_o) \tag{18}
\]

$U_o$ is the supply voltage. From Figure 5 we can get the following equations:

\[
t_1 \frac{di_F^+}{dt} - t_1 \frac{di_F^-}{dt} = h - \delta \tag{19}
\]

\[
t_2 \frac{di_F^+}{dt} - t_2 \frac{di_F^-}{dt} = -(h - \delta) \tag{20}
\]

The switching period and switching frequency are as follows:
\[ t_1 + t_2 = T_f = \frac{1}{f_s} \]  

(21)

Adding equation (19) and (20) and using equation (21) give the following equation:

\[ \frac{\frac{df_F^*}{dt}}{t_1} + \frac{df_F^*}{t_2} - \frac{1}{f_s} \frac{df_F^*}{dt} = 0 \]

(22)

Subtracting equation (20) and (19), gives:

\[ \frac{df_F^*}{dt} - \frac{df_F^*}{dt} - (t_1 - t_2) \frac{df_F^*}{dt} = 2(h - \delta) \]

(23)

Substituting equation (18) in (23), gives:

\[ \frac{(t_1 - t_2) df_F^*}{dt} - (t_1 - t_2) \frac{df_F^*}{dt} = 2(h - \delta) + 0.5t V_{dc} \]

(24)

Similarly, substituting equation (18) in (22), gives:

\[ \frac{df_F^* - 0.5t V_{dc}}{dt} = \frac{df_F^*}{f_s} \]

(25)

Finally, substituting equation (25) in (24), gives:

\[ f_s = \frac{1}{2(h - \delta) + 0.5t V_{dc}} \left\{ \frac{df_F^*}{dt} \left[ 1 - \left( \frac{df_F^*}{dt} \right)^2 - 0.5t V_{dc} \frac{df_F^*}{dt} \right] \right\} \]

\[ = \frac{1}{2(h - \delta) + 0.5t V_{dc}} \left( \frac{L_f}{L_f} \left( \frac{df_F^*}{dt} - 0.5t V_{dc} \frac{df_F^*}{dt} \right) \right) \]

(26)

As illustrated in Figure 5, the Three-level HCC PWM scheme does not use both the lower and the upper band simultaneously. So for the same hysteresis band the switching frequency in three-level scheme will be more than two-level.
3.2 Carrier-based PWM

CPWM technique is one of the most popular modulation techniques applied in power switching inverters. In CPWM, a reference voltage waveform is compared with a repetitive switching frequency triangular carrier waveform to generate gate signals for the switches of inverter. The principle of CPWM current control for 2L-VSI is illustrated in Figure 6. In this figure, the difference between the reference currents and the compensating actual currents are sent to the Proportional Integral (PI) controllers to generate the reference voltages. These reference voltages are intersected with the triangular carrier signal and compared by using the limit comparators to generate the switching pulses of switches. If the reference voltage is higher than the triangular carrier waveform then the comparator output is 1. If the reference voltage is less than the triangular carrier waveform then, the comparator output is 0.

![Figure 6: Block diagram of CPWM for 2L-VSI](image-url)
To increase the performance of multilevel inverters, the multi-carrier CPWM control methods are used. These methods have been classified in two categories: level-shifted known as Phase Disposition (PD) and Phase-Shifted (PS). Since PD is appropriate for NPC VSIs and it is the method that produces the minimum THD% in phase voltages [17, 18], in this paper, PD multi-carrier CPWM is chosen to switch 3L-NPC VSI. Block diagram of CPWM current control for 3L-NPC VSI is demonstrated in Figure 7. As shown in this figure, the reference voltage signals are intersected with two triangular carrier signals and compared, by using the limit comparators to generate the switching pulses of switches.

![Diagram](https://example.com/diagram.png)

**Figure 7:** Block diagram of CPWM for 3L-NPC VSI
4. Simulation Results and Analysis

To verify the performance of shunt APF, the system shown in Figure 1 has been simulated by MATLAB/SIMULINK software. The simulation system parameters are given in Table 3. Two aforementioned modulation techniques described in previous section, are applied to control the current injection of shunt APF which is implemented with a 2L-VSI and 3L-NPC VSI. The analysis is based on using the same switches with equal switching frequency range. Hence, effective parameters in switching frequency such as, hysteresis band width in HCC PWM ($h$), frequency of triangular carrier signal ($f_c$), amplitude modulation ratio ($m_a$) and PI controllers coefficients ($k_p$, $k_i$) in CPWM are designed carefully to achieve the same maximum switching frequency for all four cases. The selected values for these parameters are shown in Table 4. Since the three-level scheme does not use both the lower and the upper band simultaneously, the hysteresis band in this case is considered twice of the bands in two-level scheme.

Figure 8 indicates switching frequency variation for the top switch of phase-a leg ($S_{1a}$) in both described inverters, over a fundamental cycle of supply voltage. Hence, effective parameters in switching frequency such as, hysteresis band width in HCC PWM ($h$), frequency of triangular carrier signal ($f_c$), amplitude modulation ratio ($m_a$) and PI controllers coefficients ($k_p$, $k_i$) in CPWM are designed carefully to achieve the same maximum switching frequency for all four cases. The selected values for these parameters are shown in Table 4. Figure 8 indicates switching frequency variation for the top switch of phase-a leg ($S_{1a}$) in both described inverters, over a fundamental cycle of supply voltage.
Table 3: System Parameters

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Supply voltages ((V_{sa}, V_{sb}, V_{sc}))</td>
<td>100 V (L–N rms)</td>
</tr>
<tr>
<td>System frequency ((f))</td>
<td>50 Hz</td>
</tr>
<tr>
<td>Three phase linear load ((R_2, L_2))</td>
<td>(20 \Omega, 100 \text{ mH})</td>
</tr>
<tr>
<td>Nonlinear load ((R_1, L_1))</td>
<td>(50 \Omega, 100 \text{ mH})</td>
</tr>
<tr>
<td>DC link voltage ((V_{dc}))</td>
<td>320 V</td>
</tr>
<tr>
<td>DC link capacitor ((C_{dc}))</td>
<td>1.1 mF</td>
</tr>
<tr>
<td>Filter interface impedance ((R_c, L_c))</td>
<td>(0.1 \Omega, 4 \text{ mH})</td>
</tr>
<tr>
<td>Line impedance ((R_s, L_s))</td>
<td>(1 \Omega, 0.1 \text{ mH})</td>
</tr>
</tbody>
</table>

As it can be seen, switching frequency in HCC-PWM method is variable and irregular and also it is exposed to wide variation. Inversely, the switching frequency of CPWM technique is constant and it is equal to the triangular carrier signal frequency \((f_c=8 \text{ KHz})\). The maximum switching frequency in HCC-PWM technique is stabilized in 8 KHz according to Table 4.

Figure 8: Switching frequency variation

a) 2L-VSI and CPWM method
b) 2L-VSI and HCC PWM method
c) 3L-NPC VSI and CPWM method
d) 3L-NPC VSI and HCC PWM method
Figure 9: Simulation system waveforms
The relevant system’s waveforms of phase (a) for all case studies are represented in Figure 9. These waveforms are as follows. The load current ($i_{La}$), the reference compensation current extracted by p-q theory ($i_{Fa}^*$), the APF current ($i_{Fa}$), the source current ($i_{Sa}$) and the line-line output voltage of inverter ($V_{ab}$). The shunt APF has been entered to the system at $t=0.04$ second. As it can be seen, in all cases source currents after compensation are close to sinusoidal waveform. The observable difference in waveforms is output voltage of inverter topologies. The numbers of line-line output voltage level in 2L-VSI and 3L-NPC VSI are equal to 3 and 5 respectively.

The comparison indexes in this paper are based on total harmonic distortion percent (THD%) in the source current after compensation, switching numbers and switching losses of switches. From the simulation results illustrated in Table 4, THD% of the distorted source currents is improved from 18.4% to less than 5% after compensation. The results of comparison show that, for the same maximum switching frequency, CPWM is better than HCC PWM and it provides minimum THD% of source current. In three-level HCC PWM scheme, all the four switches in each leg do not change their states simultaneously. Hence, the average of switching frequency in 3L-NPC VSI is less than 2L-VSI. However, 3L-NPC VSI has better performance than 2L-VSI. The other proposed indicator is numbers of switching (N.S) for switches which are calculated over a cycle of supply voltage. The results represented in Table 4 show that, in HCC PWM method switching number of switches is minimum. It is because of irregular and variable switching frequency of HCC PWM method. By comparing the switching numbers of switches and average switching frequency it is obtained that, switching losses in CPWM is more than HCC PWM.
Table 4: Comparison of Modulation Techniques

<table>
<thead>
<tr>
<th>Indexes</th>
<th>HCC PWM</th>
<th>CPWM</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>2L-VSI</td>
<td>3L-NPC VSI</td>
</tr>
<tr>
<td></td>
<td>3L-NPC VSI</td>
<td>2L-VSI</td>
</tr>
<tr>
<td></td>
<td>h=0.7</td>
<td>h=1.4</td>
</tr>
<tr>
<td></td>
<td>m_a=0.8</td>
<td>f_c=8 KHz</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>f_{swmax} (KHz)</td>
<td>8.04</td>
<td>8.03</td>
</tr>
<tr>
<td>f_{swavg} (KHz)</td>
<td>4.4</td>
<td>4.1</td>
</tr>
<tr>
<td>THD_h % before compensation</td>
<td>18.4</td>
<td>18.4</td>
</tr>
<tr>
<td>THD_h % after compensation</td>
<td>4.23</td>
<td>4.84</td>
</tr>
<tr>
<td>Number of switching</td>
<td>62</td>
<td>32</td>
</tr>
</tbody>
</table>

Conclusion

In this paper, two different modulation techniques to control a shunt APF have been implemented and evaluated using two different inverter topologies: 2L-VSI and 3L-NPC VSI. Both of the strategies, hysteresis current control PWM and carrier-based PWM improved THD% of source current. The results of comparison show a certain superiority of the CPWM, in both two-level and three-level inverter topologies. Indeed, CPWM provides minimum THD% of source current. The results also show that the switching number of switches and average switching frequency in HCC PWM method is less than CPWM method. It is because of irregular and variable switching frequency of HCC PWM method.
References

Authors

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