New Control Methods for Single Phase PWM Regenerative Rectifier with Power Decoupling Function

Kuo-Hen Chao  
Student Member, IEEE  
Center for Advanced Power Technologies  
Dept. of Electrical Engineering  
National Tsing Hua University  
Hsinchu, Taiwan  
chaokuohen@gmail.com

Po-Tai Cheng  
Senior Member, IEEE  
Center for Advanced Power Technologies  
Dept. of Electrical Engineering  
National Tsing Hua University  
Hsinchu, Taiwan  
ptcheng@ieee.org

Toshihisa Shimizu  
Senior Member, IEEE  
Department of Electrical Engineering,  
Tokyo Metropolitan University,  
1-1 Minami Oosawa,  
Hachioji, Tokyo, 192-0397  
JAPAN

Abstract—This paper presents a new method for power decoupling control of single phase AC/DC converters using three switch poles to achieve unity power factor operation at AC side and power decoupling function at DC side. The method proposed compensates the pulsating power at twice the grid frequency by absorbing low frequency ripple current to a DC decoupling capacitor through the third switch pole operation for bidirectional power flow. This approach can eliminate the need of large filter capacitor in DC side which is required to suppress the low frequency ripples. This paper will explain the principles of operations of the proposed method, and experimental results will also be presented for validation.

Keywords- AC/DC converter, power decoupling

I. INTRODUCTION

With the dramatic increase of electronics loads, the harmonic pollution of AC grid receives more and more attentions nowadays. In order to address this concern, AC/DC converters with power factor correction (PFC) capability are often adopted. However, for single phase applications, the PFC functionality causes significant pulsating power of twice the grid frequency at the DC side, and a large capacitor bank or a L-C tuned filter may be required to suppress the low frequency ripple. As a result, these bulky filters lead to increased size and weight of the converter.

Conventionally the pulsation power problem is solved by using passive filter [1]. With the significant improvement of semiconductor of power device nowadays, the power decoupling problem can be solved by active decoupling circuit with acceptable cost and reliability. There are many circuit topologies for active power decoupling, most of these circuits decouple the reactive power by drawing sinusoid current at twice the grid frequency from the DC bus and store the energy in capacitors [2, 3, 4], in inductor [5]. But most of these circuit topologies are based on flyback type converter and only have unipolar power flow. Topologies with three switch poles converters have also been proposed for power decoupling functions [6,7] in single phase AC/DC applications.

Figure 1 shows the circuit configuration of the single-phase three-pole AC/DC converter with the power decoupling capability. The u and v phases operate as conventional PWM rectifier, and the z phase accomplishes the power decoupling capability by transferring the pulsating power at DC side to the power decoupling capacitor $C_{dp}$. The proposed control coordinates the operations of the u, v, and z phases to achieve the desired power decoupling function. The power decoupling method proposed for this circuit is a relatively simple and cheap solution. The effectiveness of the power decoupling method is confirmed by simulation and experimental results.

II. OPERATION PRINCIPLE

The proposed method is based on the single-phase, three-switch poles AC/DC converter as in Fig. 1. The u-phase and v-phase poles are connected to the AC grid. This H-bridge configuration allows bi-directional power flow so the DC bus voltage can be properly maintained whether the load is consuming or regenerating power. The z-phase pole accomplishes the power decoupling function by circulating the $2\omega_0$ pulsation power between the main DC bus and the power decoupling capacitor $C_{dp}$. The control block diagram of the converter is given in Fig. 2 and the operation principles are presented in the following sections.

A. PWM Rectifier

The control block diagram of the single-phase three-pole converter is shown in upper side of Fig. 2 The PWM converter controller is to control the DC bus voltage through unity power factor operation at the AC side. A proportional-integral (PI) closed-loop control regulates the main DC bus voltage $E_{dc}$ at the desired level. A feedforward command of load power consumption is added on top of the closed-loop control output in order to improve the $E_{dc}$ dynamic performances. The resulting current command amplitude is
The voltage and the current of the converter are expressed added to produce the final voltage reference for the PWM. Amplified by a proportional gain AC/DC converter. The error of the output current is command phase-locked to the AC grid voltage. The final current then multiplied by a normalized cosine wave which is phase-locked to the AC grid voltage. The final current command $i_z$ ensures the unity power factor operation of the AC/DC converter. The error of the output current is amplified by a proportional gain $K_p$ and the line voltage is added to produce the final voltage reference for the PWM.

The power flow of the converter is analyzed as follows. The voltage and the current of the converter are expressed as:

$$V_{uv} = V_u - V_v = \bar{V}_{uv} \cos(\omega_0 t + \phi)$$ \hspace{1cm} (1)

$$i_z = \bar{i}_z \cos(\omega_0 t + \phi + \psi)$$ \hspace{1cm} (2)

Where $\psi$ is the phase angle difference between $v_{uv}$ and $i_z$, and $\psi$ is caused by the voltage drop across on the inductor $L_{cp}$. Assuming the converter is lossless, the power flow at DC side is

$$P_{dc} = \frac{\bar{V}_{uv} \bar{i}_z}{2} \cos \psi + \frac{\bar{V}_{uv} L_{cp}}{2} \cos 2(\omega_0 t + \phi + \frac{\psi}{2}) \hspace{1cm} (3)$$  

$$= P_{active} + P_{reactive}$$

Equation (3) shows that the converter produces a pulsating power at $2\omega_0$ while consumes (or regenerates) active power. The $z$-phase switch pole is to compensate this pulsating power as explained in the next section.

### B. Power decoupling Method

The $z$-phase switch pole circulates the reactive power by producing a DC voltage and a $2\omega_0$ sinusoidal voltage. The resulting $2\omega_0$ current component and the DC voltage can compensate the $2\omega_0$ pulsating power. The control block is shown as the lower half of the Fig. 2. The $z$-phase voltage is given as in equation (4),

$$i_z = \bar{i}_z \cos(2\omega_0 t + \phi + \gamma)$$ \hspace{1cm} (4)

At steady state, the current $i_z$ can be expressed as

$$v_z = \bar{i}_z \sin(2\omega_0 t + \phi + \gamma) + \bar{i}_s \sin(2\omega_0 t + \phi + \gamma) \left( -\frac{1}{2\omega_0 C_{cp}} + v_b \right) + v_b$$ \hspace{1cm} (5)

where $X = \left( -\frac{1}{2\omega_0 C_{cp}} + \frac{2}{2\omega_0 L_{cp}} \right)$ is the impedance of the path, and $v_b$ is the DC average voltage of $E_z$. The $z$ phase power can be calculated by multiplying $i_z$ and $v_z$ as follow:

$$P_z = \bar{i}_z \bar{v}_z \cos(2\omega_0 t + \phi + \gamma) + i_z v_b \cos(2\omega_0 t + \phi + \gamma)$$

The $2\omega_0$ component of $P_z$ is used to compensate the pulsating power at the DC bus. From equation (3) and (6) can find

$$i_z v_b = \frac{\bar{V}_{uv} \bar{i}_s}{2} \Rightarrow i_z = \frac{\bar{V}_{uv} \bar{i}_s}{2 v_b} = I_{Load} K_{p3}$$ \hspace{1cm} (7)

$$\gamma = \frac{\psi}{2}$$ \hspace{1cm} (8)

The lower side of Fig. 2 shows the $z$ phase control block. From the result of equation (7) and (8), the current command $i_z$ can be constructed by creating a cosine wave of frequency $2\omega_0$ with $\gamma$ phase shift and the amplitude $\bar{i}_s$. Simple proportional gain $K_{p3}$ amplifies the current error and then bias voltage $v_b$ is added to produce the final voltage reference of the $z$ phase PWM.

The power decoupling method can mathematically prove that the $z$ phase circuit can produce $2\omega_0$ ripple power. If the amplitude and phase can be controlled correctly, this ripple power can eliminate the ripple power comes from the AC grid caused by unity power factor.

### C. Power decoupling capability

The power decoupling capability is influenced by various factors, such as the average duty $D$ of the $z$-phase switch pole, the inductor $L_{pd}$ and power decoupling
capacitor \( C_{pd} \). Due to the buck circuit nature of the z-phase, the main DC bus voltage \( E_{dc} \) must be higher than \( E_z \) on a steady-state basis as in equation (9)

\[
E_{dc} \geq V_i + E_z + \Delta E_z
\]

where DC average of \( E_z \) is equal to \( 2\omega_0 L_{cp} i_z \) and the voltage drop across the inductor is \( \frac{1}{2} \omega_0 L_{cp} i_z \) and the voltage ripple of the capacitor is \( \frac{i_z}{2\omega_0 C_{cp}} \). So equation (9) can be re-written as:

\[
E_{dc} \geq -2\omega_0 L_{cp} i_z + DE_{dc} + \frac{i_z}{2\omega_0 C_{cp}}
\]

The \( 2\omega_0 \) oscillation of power is absorbed by the z-phase, so the resulting \( i_z \) is expressed as

\[
i_z = \frac{\Delta v}{2\omega_0}\]

\[
E_{dc} \geq -2\omega_0 L_{pdc} P_{dc} \theta_{cc} \]

\[
E_{dc} \geq \frac{P_{dc}(1-4\omega_0^2 L_{pdc} C_{cp})}{2\omega_0 C_{cp} D(1-D)}
\]

\[
P_{dc} \geq \frac{2\omega_0 C_{cp} D(1-D) E_{dc}^2}{1-4\omega_0^2 L_{pdc} C_{cp}}
\]

(9)

Where \( D \) is the ratio of \( v_i \) and \( E_{dc} \). From equation (9), the power decoupling capability mainly depends on the power decoupling capacitor \( C_{pd} \) and average duty of the power decoupling circuit. The inductor \( L_{pd} \) also helps to increase the power decoupling capability but the effect is small.

III. EXPERIMENTAL RESULT

The proposed controls are tested on the prototype of the single-phase three-pole converter in the laboratory. The testbench parameters are as follows:

- AC grid: 110Vrms; 60Hz; filter inductor \( L_s \approx 3\mathrm{mH} \);
- Converter: IGBTs hard-switching at 10kHz; DC bus command \( E_{dc}^{*} = 200\mathrm{V} \); DC bus filter capacitor \( C_{dc} = 120\mu\mathrm{F} \) and the load consumption \( P_o = 650\mathrm{W} \).
- Power decoupling circuit: inductor \( L_{pd} \approx 3\mathrm{mH} \); power decoupling capacitor \( C_{cp} \approx 300\mu\mathrm{F} \). Average duty at 0.75..

Fig. 3 (a) shows the test results of the converter when the power decoupling is disabled. Severe \( 2\omega_0 \) oscillations on the DC bus \( E_{dc} \) occurs due to the single-phase AC/DC conversion. The AC current is also distorted as the closed-loop control of the \( E_{dc} \) tries to suppress the \( 2\omega_0 \) oscillations. Fig. 3 (b) and (c) show the test results when the power decoupling function is enabled under the power consuming mode and the regeneration mode. When the z phase is engaged, the \( 2\omega_0 \) pulsation power is absorbed as the z-phase draws the \( 2\omega_0 \) current from the main bus, and stores the pulsating power in the power decoupling capacitor \( C_{cp} \). The proposed power decoupling control successfully reduces the low frequency ripple voltage to 5% of the DC bus voltage under both the loading mode and regeneration mode. And the total harmonic distortion of the input current \( i_i \) can also be reduced below 4% under power decoupling control.

Figure 3: Input current, voltage, z phase current and DC bus voltage under 650W (a) without any power decoupling; (b) converter works in rectifier mode (c) converter works in regeneration mode (\( E_{dc} \) 50V/div; \( i_i \) 5A/div; \( i_{10} \)A/div; \( V_s \) 200V/div; time 5ms/div)
The z-phase draws the low frequency \(2\omega_0\) current, the current flowing into the main DC bus capacitor contains only high frequency component, thus a small DC bus capacitor can be employed to maintain a smooth DC bus voltage. Fig. 4 shows the experimental waveforms of the DC bus current \(I_{dc}\) before and after the power decoupling function is engaged. Significant \(2\omega_0\) ripple can be observed in Fig 4(a) as the z-phase is disabled. In Fig 4(b) the \(2\omega_0\) ripple is effectively reduced when the z-phase engages its power decoupling control function.

Fig. 5 shows the transients of a step load change from 340W to 740W (Fig. 5(a)) and from 740W to 340W (Fig. 5(b)). With the load current feed-forward command, the converter can quickly react to such a dramatic load change. The z-phase also adjusts its current to cope with the load change, thus the DC bus voltage dip and overshoot can be maintained within 10% of the main DC bus voltage.

The spectrum of the DC bus current \(I_{dc}\) shows that the \(2\omega_0\) ripples is successfully suppressed by proposed power decoupling control from 33% to 4% of the DC component. This feature allows the use of small capacitors for the main DC bus.

Figure 4: DC bus current (a) without any power decoupling; (b) with power decoupling \((I_{dc}\ 5A/div; i_s\ 20A/div; V_s\ 200V/div; time\ 5ms/div)\)

Figure 5: Step load change (a) from 340W to 740W; (b) from 740W to 340W \((E_{dc}\ 50V/div; i_{load}\ 2A/div; i_s\ 20A/div; V_s\ 200V/div; time\ 10ms/div)\)

Figure 6: Spectra of the DC bus current \(i_{dc}\) before and after power decoupling function is enabled.
The power decoupling function can be accomplished by placing capacitor on the AC side as shown in Fig. 7 [7]. The AC side power decoupling circuit in Fig 7 is tested under the identical conditions, and the results are given in Fig. 8. Fig. 8 (a) shows that the main DC bus ripple is successfully suppressed, and the grid current $i_s$ contains some distortion due to the interaction between the AC side power decoupling capacitors and the grid inductor. The $u$ phase and $v$ phase current in Fig 8(b) are composed of input current $i_s$ and $z$ phase current. Thus their RMS values and peak values become higher than $i_s$.

Table 1 shows the comparison of the peak current of each phases between the AC side capacitor power decoupling method and the proposed method. The power decoupling capacitor at AC side method has larger peak current at $u$ phase and $v$ phase because the $z$ phase current will flow through them, the peak current in $u$-phase and $v$-phase reach almost twice as much as the peak value of $i_s$. On the other hand, the proposed method places the power decoupling capacitor on the DC side, thus $u$ phase and $v$ phase do not carry any power decoupling related current.

<table>
<thead>
<tr>
<th></th>
<th>$u$ phase</th>
<th>$v$ phase</th>
<th>$z$ phase</th>
</tr>
</thead>
<tbody>
<tr>
<td>AC capacitor method</td>
<td>20.5A</td>
<td>20.5A</td>
<td>10A</td>
</tr>
<tr>
<td>Proposed method</td>
<td>8.5A</td>
<td>8.5A</td>
<td>4.5A</td>
</tr>
</tbody>
</table>

Table 1: Peak current value of each phases under different power decoupling circuit.

**Summary**

A power decoupling method is proposed for the bi-directional AC/DC front-end in this paper. The proposed power decoupling circuit and its control can be easily integrated with the conventional H-bridge AC/DC PWM converter to achieve unity power factor operation at the AC grid side and power decoupling at DC side. The power decoupling function performed by the extra $z$-phase switch pole can absorb the low frequency $2\omega_0$ ripple power, thus allows the use of much smaller capacitors to maintain the main DC bus voltage. The conventional electrolytic capacitors in the main DC bus can be replaced by film capacitors for extended service cycle of the converter system. This feature will receive more attention as PV converters applications continue to grow. The proposed circuit also allows the use of the commercially available three-phase IGBT modules for the single-phase AC/DC application. The proposed power decoupling method is validated by mathematical analysis and experimental result.

**REFERENCE**


