Three-Phase Interleaved Parallel Vienna Rectifier Based on Coupling Transformer

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Abstract: This paper presents a three-phase interleaved parallel Vienna rectifier employing a coupling transformer. The application of coupling transformer figures out the current sharing problem between two parallel Vienna converters successfully, meanwhile, the volume of input EMC filters and passive components are reduced for the coupling transformers. The proposed rectifier also presents the inherent benefits: high power factor, higher power density, lower input current ripple and lower output voltage ripple. The operation of the PFC rectifier is described, including the design guidelines for the coupling transformer.

Introduction

With the continuous improvement of power electronic devices in voltage and power levels in recent years, multi-level converters are widely used in the field with high voltage and high power[1,2]. Multilevel is generated in three-phase rectifiers to improve the power quality in high ac-dc converters, and the prevalent method in multi-level rectifiers is the neutral-point-clamped type[3,4]. However, the considerable number is an economic issue to be considered, semiconductor switches are needed to generated N levels employing neutral-point-clamped rectifier, while is required only in which proposed rectifiers.

Traditional methods of suppressing circulating currents are mainly Voltage Droop Control (VDC)[5,6], which is only applied to the situation with simple and lower requirements; master slave control[7,8], which has a better effect on circulation inhibition but a complicated control strategy; automatic current sharing average current control[9]. The interleaved parallel based on coupling transformer can suppress the circulating current well with a simpler control method, which can also decrease the ripple of the input current and improve the ability to defend electromagnetic interference.

A three-phase interleaved parallel Vienna rectifier employing coupling transformers is presented and discussed in this paper. The proposed rectifier can be designed to have a simple circuit, outstanding circulation suppression abilities, low input current ripple.

Three-phase parallel Vienna rectifier with coupling transformers

The proposed three-phase interleaved parallel Vienna rectifier with coupling transformer is shown in Fig.1.
The proposed three-phase parallel Vienna converter operates as a boost rectifier, which is fed by a three-phase three-wire ac power system \((e_i, i = a, b, c)\). The boost inductor \((L_k, k = a, b, c)\) in each phase enhances input phase voltage of the rectifier. The coupling transformer (CT) with \(N\) windings \((N_j, j = 2, 3...N)\) connects the corresponding boost inductor \(L_k\) to \(W\) switching networks. CT can share the input current \((i_k, k = a, b, c)\) among \(N\) windings, which can transmit larger power from ac-link to dc-link and decrease the current stress on the semiconductor switches. In this paper, CT mainly has two function: 1) suppress the loop currents among parallel switching network; 2) average the voltages on each winding to fulfill multilevel of input voltage. \(W\) Vienna-type Switching Networks (VTSN) in parallel, consist of three single-pole single-throw (SPST) switches \((S_{kj}, k = 1, 2...N; j = 1, 2, 3)\) and six diodes, which is used to generate three voltages in each winding of CT. The output is comprised two filter capacitors \((C_1 \text{ and } C_2)\), which is assumed that voltages \((V_{c1} \text{ and } V_{c2})\) on both are identical and balanced. We take \(N=2\) as the example to analyze the operational principle in the next work.

**Operation Modes.** A phase of the proposed rectifier with \(N=2\) is taken as an example to explain the operation process, which is discussed in two situations: duty cycle \(D < 0.5 \text{ and } D > 0.5\).

1) \(D > 0.5\)

The driving signals on the gate of the semiconductor switches and the main current waves are
shown in Fig.2. Condition in the positive cycle of the input current and phase a with N=2 are only discussed here.

Model 1 ($t_1 < t < t_2$): During this interval, switch $S_{a1}$ is ON and $S_{a2}$ is OFF. The current $i_{a1}$ in winding 1 is rising, while the current $i_{a2}$ in winding 2 starts to fall from the maximum. In this mode, ac source and inductor transmit energy to the dc side together. As a result, the boost inductor begins to discharge and the current falls. The voltages on winding 1 and winding 2 are 0 and $\frac{V_0}{2}$, the voltage at the input side of coupling transformer will be $\frac{V_0}{4}$ for the voltage-averaging characteristic.

Model 2 ($t_2 < t < t_3$): In this model, switch $S_{a1}$ is OFF and $S_{a2}$ is ON. The current in winding 1 is still increasing and will reach the maximum at the end of this model, while current in winding 2 starts to rise from the minimum. The boost inductor begins to charge and the current rises after the combination of currents $i_{a1}$ and $i_{a2}$. Meanwhile, the energy is transmitted to inductor and output side from ac source. The voltages on winding 1 and winding 2 are both clamped to $\frac{V_0}{2}$, and the voltage at the input side of coupling transformer becomes $\frac{V_0}{2}$.

Model 3 ($t_3 < t < t_4$): In model 3, switch $S_{a1}$ and $S_{a2}$ are both ON. The current in winding 1 becomes falling and will approach to the minimum at the end of this model, the current in winding 2 is still rising up to maximum. In this interval, the boost inductor begins to discharge and the current falls after the combination of currents $i_{a1}$ and $i_{a2}$. The energy is transmitted to output side from ac
source and the inductor. The voltages on winding 1 and winding 2 are clamped to 0 and \( \frac{V_0}{2} \), respectively. The voltage at the input side of coupling transformer becomes \( \frac{V_0}{4} \).

\textit{Model 4 (t_4 < t < t_5)}: In this model, switch \( S_{a1} \) and \( S_{a2} \) are both ON. The current in winding 1 starts to rise from the minimum, and the current in winding 2 is still increasing. The boost inductor begins to charge and the current rises after the combination of currents \( i_{a1} \) and \( i_{a2} \). Meanwhile, the energy is transmitted to inductor and output side from ac source. The voltages on winding 1 and winding 2 are both clamped to \( \frac{V_0}{2} \), and the voltage at the input side of coupling transformer becomes \( \frac{V_0}{4} \).

2) \( D < 0.5 \)

The situation in \( D < 0.5 \) is similar with the one in \( D > 0.5 \), which is seen in Fig.3 and the analysis of operation modes is neglected here.

![Figure 3. Main current waveforms with duty cycle \( D < 0.5 \).](image)

**Operational Principle of Coupling Transformer**

Circulating Current Suppression Principle. It’s reasonable that each VTSN is equivalent to an ideal current source parallel with a resistance according to Norton Theorem. The equivalent circuit of the interleaved parallel mode is presented in Fig.4.

![Figure 4. The equivalent circuit of the interleaved parallel mode.](image)
In the equivalent circuit, $G_1$ and $G_2$ are equivalent transadmittance. It’s apparent that circulating current, between two current $i_{u1}$ and $i_{u2}$ are both equivalent no-load current of VTSN, parallel modules, will be generated when $i_{u1} \neq i_{u2}$ or $G_1 \neq G_2$.

It’s proved to be feasible to decompose the currents $i_1$ and $i_2$ into the common mode and the differential mode. The common mode current $i_c$, and the differential mode currents $i_{d1}$ and $i_{d2}$ are defined as follows.

$$
\begin{align*}
    i_c &= \frac{1}{2} (i_1 + i_2) \\
    i_{d1} &= i_1 - i_c \\
    i_{d2} &= i_2 - i_c
\end{align*}
$$

(1)

Where the differential mode current is namely circulating current, and $i_d = i_{d1} = -i_{d2}$. Therefore, circulating current is shown in Fig.5.

Figure 5. Schematic diagram of circulating current between modules.

Here, we take coupling transformer $T_1$ as the example to verify how the CT suppresses the generation of the circulating current. Considering a symmetrical structure without leakage flux, it is assumed that the self-inductance of each winding and the mutual-inductance between arbitrary two windings is equal respectively. Therefore, we decompose the coupling transformer into self-inductance $L$ and mutual-inductance $M = \frac{L}{N-1}$. The voltage-current relation is obtained as follows:

$$
\begin{align*}
    v_{u1} &= L \frac{di_{u1}}{dt} - \frac{1}{N-1} L \frac{di_{u2}}{dt} - \cdots - \frac{1}{N-1} L \frac{di_{uN}}{dt} \\
    v_{u2} &= -\frac{1}{N-1} L \frac{di_{u1}}{dt} + L \frac{di_{u2}}{dt} - \cdots - \frac{1}{N-1} L \frac{di_{uN}}{dt} \\
    & \quad \vdots \\
    v_{uN} &= -\frac{1}{N-1} L \frac{di_{u1}}{dt} - \frac{1}{N-1} L \frac{di_{u2}}{dt} - \cdots + L \frac{di_{uN}}{dt}
\end{align*}
$$

(2)

Equation (3) can be written as follow:
\[ v_{a1} = \frac{N}{N-1} L \frac{di_{a1}}{dt} - \frac{1}{N-1} L \frac{di_a}{dt} \]
\[ v_{a2} = \frac{N}{N-1} L \frac{di_{a2}}{dt} - \frac{1}{N-1} L \frac{di_a}{dt} \]
\[ \vdots \]
\[ v_{aN} = \frac{N}{N-1} L \frac{di_{aN}}{dt} - \frac{1}{N-1} L \frac{di_a}{dt} \]  

(3)

Where \( i_a = i_{a1} + i_{a2} + \cdots + i_{aN} \), likewise, decomposing the winding currents into the common mode \( i_c \) and the differential mode \( i_{dj} \) (j=1, 2…N), and

\[ i_c = i_{a1} + i_{a2} + \cdots + i_{aN} \]
\[ i_{d1} = i_{a1} - i_c \]
\[ \vdots \]
\[ i_{dN} = i_{aN} - i_c \]  

(4)

Then, (3) can be written as follows.

\[ v_{a1} = \frac{N}{N-1} L \frac{di_{d1}}{dt} \]
\[ v_{a2} = \frac{N}{N-1} L \frac{di_{d2}}{dt} \]
\[ \vdots \]
\[ v_{aN} = \frac{N}{N-1} L \frac{di_{dN}}{dt} \]  

(5)

From equations (5), we can get that common mode inductance is null, common mode current can go through the coupling transformer without barrier, while the differential component is restrained, that is to say the circulating current is suppressed.

Voltage Average Principle. In the Vienna-type switching network, the SPST connects corresponding winding to the midpoint 0 of the dc-link. When the SPTT is turned on, the voltage on the pole is zero, while it is determined by the direction of the input current reversely. In the positive half cycle of the input current wave, the voltage on the pole is \( V_{c1} \). On the contrary, the voltage is \( V_{c2} \). Considering the assumption that the dc-link voltages are balanced and constant, then the partial voltage is \( V_{c1}=V_{c2} = \frac{V}{2} \).

Coupling transformer \( T_1 \) is taken for an example yet. The voltage on the input side of \( T_1 \) can be denoted as follow:
\[ v_{in,a} = v_{a1} + v_{i0}, \quad \vdots \]
\[ v_{in,a} = v_{a2} + v_{20} \]
\[ v_{in,a} = v_{aN} + v_{N0} \]

And, from (3) and equation \( i_a = i_{a1} + i_{a2} + \cdots + i_{aN} \), the following formula is true.

\[ v_{a1} + v_{a2} + \cdots + v_{aN} = 0 \] (7)

From (6) and (7), we can get:

\[ v_{in,a} = \frac{1}{N}(v_{a1} + v_{a} + \cdots + v_{aN}) \] (8)

It is apparent that the coupling transformer can average the voltage of the parallel nodes. The voltage levels at the input side of the coupling transformer rely on the switching states, which are determined by the implemented modulation strategy.

Design of Coupling Transformer. The rating of the coupling transformer is analyzed with two windings. Firstly, we decompose the winding currents \( i_{a1} \) and \( i_{a2} \) into common component and differential component respectively, i.e., \( i_c \), \( i_{d1} \) and \( i_{d2} \). Then the equations as follow are tenable.

\[
\begin{cases}
I_{a1} = \sqrt{I_c^2 + I_{d1}^2} \\
I_{a2} = \sqrt{I_c^2 + I_{d2}^2}
\end{cases}
\] (9)

Where, currents in the equations are rms values.

The rms values of winding voltages are obtained where differential currents are assumed sinusoidal waves.

\[
\begin{cases}
V_{S1} = 2\omega_s L_s I_{d1} \\
V_{S2} = 2\omega_s L_s I_{d2}
\end{cases}
\] (10)

Where, \( L_s \) is the self-inductance of the windings.

It is impossible for coupling transformer to eliminate the differential current totally, which is suppressed to minimum. Therefore, it is assumed that the differential current is restrained to \( \alpha \) times the common one. Then

\[
\begin{cases}
I_{d1} = \alpha I_c \\
I_{d2} = \alpha I_c
\end{cases}
\] (11)

The winding voltages and current rms values can be expressed by common current.

\[
\begin{cases}
V_{S2} = V_{S1} = 2\omega_s \alpha L_s I_c \\
I_{a2} = I_{a1} = \sqrt{1 + \alpha^2} I_c
\end{cases}
\] (12)
The rated capacity of the coupling transformer can be expressed as follows.

\[ S = V_{s1} I_{s1} + V_{s2} I \]  \hspace{1cm} (13)

When the rated capacity is determined, then self-inductance of the winding will be:

\[ L_s = \frac{S}{4\omega \alpha \sqrt{1 + \alpha^2 I_e^2}} \]  \hspace{1cm} (14)

Control Strategy. With the expansion of the parallel units, the number of semiconductor switches will increase. To simplify the implementation and expansion in commercial applications, SPWM modulation based on carrier phase-shifting (CPS) is further discussed in this study. In detail, the carrier of each VTSN is shifted \( \frac{2\pi}{N} \) from each other for \( N \) parallel units, without changing the carrier frequency and amplitude.

Modulation strategy carriers disposition example for rectifier with \( N=2 \) is shown in Fig.6.

\[ \text{Figure 6. An example for SPWM based on CPS control strategy with N=2.} \]

From Fig.6, the gate ON signal for a given switch occurs in two situations:

1. In the positive half cycle, the modulation wave is positive and is lower than the carrier wave at any time.
2. In the negative half cycle, the modulation wave is negative and is higher than the carrier wave at any time.

Design Guidelines

The Boost Inductor. The boost inductor at ac side plays a significant role in the proposed rectifier. On the one hand, boost inductor suppresses the harmonics of the input current and make
input current more approximative sinusoidal curve; on other hand, it also reserves energy and enhances the input voltage, which makes the converter a boost-type rectifier. Therefore, the value for the boost inductance should be low enough to ensure that the input currents have a fast response characteristic, but not so low as to result in excessively high peak currents. The reference formula is provided as follow.

$$\frac{(2V_0 - 3V) \cdot V}{2 f_s V_0 \Delta i} \leq L_0 \leq \frac{2V_0}{3 \omega i_m}$$  \hspace{1cm} (15)

Where, $V_0$ is the voltage reference in dc-link, $V$ is the peak voltage of input phase voltage, $f_s$ is the switching frequency, $\Delta i$ and $i_m$ is the ripple and peak current in input side, $\omega$ is the angular frequency of the grid.

The Filter Capacitor. There is a low frequency sinusoidal voltage component remaining in dc side, which is generated by the three-phase ac sources. Considering this condition, the minimum capacitance is given as follow.

$$C_{1_{\text{min}}} \geq C \geq \frac{V_{i_m}}{2 \omega V_0 \Delta V_0}$$  \hspace{1cm} (16)

Simulation Result Analysis

A simulation model of the proposed rectifier with $N=2$ has been built in Matlab environment. The proposed rectifier is designed according to the following specifications presented in table 1.

<table>
<thead>
<tr>
<th>Table 1. Design specifications.</th>
</tr>
</thead>
<tbody>
<tr>
<td>Output power</td>
</tr>
<tr>
<td>Input phase voltage</td>
</tr>
<tr>
<td>Output voltage</td>
</tr>
<tr>
<td>Main frequency</td>
</tr>
<tr>
<td>Output voltage ripple</td>
</tr>
<tr>
<td>Input current ripple</td>
</tr>
<tr>
<td>Theoretical efficiency</td>
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<tr>
<td>Switching frequency</td>
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</tbody>
</table>

The ac phase voltages and currents are shown in Fig.7.
From the Fig. 7, it is apparent that the input phase current and voltage has the same phase, which results in high power factor (PF). The curve of the input current approximates sine and low-current harmonic is obtained.

As shown in Fig. 8, the input phase voltages present five levels and there are nine line-to-line voltage levels.

As discussed earlier, the ripple of the current across the boost inductor is lower due to the combination of two windings currents, which are obtained by interleaved control. Meanwhile, the local changing frequency multiplication of current in the inductor is obtained by the modulation scheme. This is seen in Fig. 9, where the windings currents and current across the boost inductor are shown.
Finally, the harmonics content are shown in Fig.10, compared with the IEC 61000-3-2 class A. it’s obvious that the total harmonic distortion with a current total harmonic distortion of $THD = 1.53\%$.

![Figure 9. Main current waves (a) and their local waves (b).](image)

![Figure 10. Main current harmonic spectrum and IEC 61000-3-2 class A limits.](image)

**Conclusion**

The three-phase interleaved parallel Vienna rectifier employing coupling transformers has been presented, including appropriate operation principle, control strategy. In addition, detail operation modes and design principium are also presented in this paper. It’s apparent that coupling transformer can realize current sharing problem between two parallel Vienna rectifiers, meantime, the frequency of boost inductor is twice the one in signal Vienna rectifier, which reduces the volume of inductance, besides, the proposed rectifier has lower input current (with $THD < 2\%$) ripple and output current ripple, higher power density and easier to be extended.

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