New Single-Switch Three-Phase High Power Factor Rectifiers Using Multi-Resonant Zero Current Switching

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Abstract - New single-switch three-phase high power factor rectifiers are introduced which have continuous input and output currents. By the use of a multi-resonant scheme, the transistor operates with zero current switching and the diodes operate with zero voltage switching. These multi-resonant rectifiers with a single transistor are capable of drawing a higher quality input current waveform at nearly unity power factor and lower stress than the quasi-resonant rectifiers. Buck-type converters are used for the power stage, and hence the output voltage is lower than the input voltage. Moreover, these rectifiers have a wide load range and low stress on semiconductor devices. Simulation and experimental results are presented.

I. Introduction
Numerous publications [1-6] have treated the power factor correction of three-phase ac-dc power supplies. PWM three-phase high power factor rectifiers based on the discontinuous conduction mode have been introduced and analyzed in [1,2,5,6]. However, the PWM boost-type high power factor rectifiers have higher output voltage than the peak input voltage, together with a pulsating output current. Moreover, the switching power loss puts a practical upper bound on the usable frequency range.

The quasi-resonant zero current switching buck-type high power factor rectifiers have been introduced in [2,4]. It was shown in [2] that, even though a resonant circuit is used, the transistor currents are lower than those in an equivalent PWM 3φ-dc converter. In addition, the zero current switching property makes this approach well-suited for applications employing IGBT's. However, the buck-type rectifiers in [2] have pulsating input currents and hence require an extra input filter. The quasi-resonant zero voltage switching boost-type rectifiers of [3] exhibit nonpulsating input currents, but load current variations lead to high peak transistor voltages stress.

In this paper, a new family of 3φ single-switch high power factor multi-resonant buck-type rectifiers are introduced which have continuous input and output currents. Figure 1 shows the new 3φ multi-resonant zero current switching cell and the voltage waveform of the input resonant capacitor C_r. By the use of a multi-resonant scheme, the transistor operates with zero current switching and the diodes operate with zero voltage switching. Moreover, these multi-resonant rectifiers with a single transistor are capable of drawing an even higher quality input current waveform at nearly unity power factor and lower stress than the quasi-resonant rectifiers or PWM discontinuous conduction mode boost-type rectifiers. This is true because the portion of period b and period c in the one switching interval as shown in Fig. 1(c) is shorter than the other types of topologies. Hence, the dominance of period a, when the voltage V_Cr1 is proportional to the input current, produces a more linear input characteristic. Buck-type converters are used for the power stage, and hence the output voltage is lower than the input voltage. These rectifiers have a wide load range and low stress on the semiconductor devices.

In Section II, the benefits and the special characteristics of the new 3φ single-switch high power factor multi-resonant buck-type rectifiers are briefly discussed and compared with those of the high power factor rectifiers.
based on quasi-resonant rectifiers and PWM discontinuous conduction mode boost-type rectifiers.

In Section III, the principal operation of the 3φ single-switch high power factor multi-resonant buck-type rectifier is presented.

In Section IV, these rectifiers are analyzed and the results of this analysis are presented as graphical forms. The special properties of the multi-resonant topology for the high power factor also explained. Simulation and experimental results are presented in Section V.

II. New Single-Switch Three-Phase High Power Factor Rectifiers Using Multi-Resonant Zero Current Switching

Figure 1 shows the new 3φ multi-resonant zero current switching cell and the voltage waveform of the input side resonant capacitor \( C_r \). At the steady state, the average voltage of \( C_r \) during one switching period is same as the input voltage. Moreover, the peak voltage of \( C_r \) is proportional to the input current. If the switching frequency is much higher than the input line frequency, then the input current will follows the input voltage waveform. Hence, this phenomenon results in the high power factor and low harmonic input current characteristic.

By the use of a multi-resonant scheme, the transistor operates with zero current switching, and the diodes operate with zero voltage switching. Moreover, this multi-resonant scheme has great benefits for drawing a higher quality input current waveform than the quasi-resonant scheme or the discontinuous conduction mode PWM method. From Fig. 1(c), the resonant voltage waveform of \( V_{Cr} \) can be divided by three different periods. During the first period, \( V_{Cr} \) is increasing linearly with slope proportional to the input current \( i_L \). During the second period, the resonant capacitor \( C_r \) is ringing together with the resonant tank inductor \( L_r \) until \( V_{Cr} \) reaches zero voltage. Finally, \( V_{Cr} \) remains at zero for the third period. Hence, if the first period is longer than the sum of second and third periods, then the input current waveform becomes more proportional to the input voltage waveform. Indeed, it is a good design, if the first period is longer than the second and third periods. Therefore, these multi-resonant rectifiers are capable of drawing a high quality input current waveform at nearly unity power factor with a single transistor. Simulation and experimental results demonstrate total harmonic distortion of less than 4.5% at the full output power, and less than 1% at the 10% output power, in an open loop rectifier. Moreover, these rectifiers have a wide load range with low voltage stress on the semiconductor devices.

Figure 2 shows the basic circuit diagram and ideal waveforms of the new single-switch 3φ multi-resonant ZCS HPFP rectifiers. The inductance \( L_r \) and the capacitors \( C_{r1-2} \) and \( C_d \) form the multi-resonant tank circuit and lead to zero current switching in the transistor and zero voltage switching in the diodes. Moreover, the voltage waveforms of the resonant tank capacitors \( C_{r1-3} \) are pulsating sinusoidal with peaks proportional to the input line currents. This property yields an average or low frequency component in the phase voltage approximately proportional to the line current. Hence, low harmonic rectification is obtained. Inductors \( L_a, L_b \), and \( L_c \) are filter inductors with small current switching ripples.

The converter is basically a buck topology, with output voltage controllable between zero and approximately the peak ac line-line voltage. Similar converters based on other dc-dc parent converters can also be derived which have other voltage conversion ratios.

The proposed rectifiers have many advantages. These advantages include
1. High power factor, low harmonic rectification is performed naturally.
2. Input and output currents are continuous.
3. Because of the buck-type rectifier property, the output voltage is lower than the peak input line-line voltage and hence the voltage stress on the transistor is lower than in the boost-type rectifier.
4. Wide range of the load power variation is achieved.

Fig. 2. The basic converter circuit and ideal waveforms of the new single transistor three-phase multi-resonant zero current switching high power factor rectifier.
5. Use of a single controlled switch operating at zero current switching with good switch utilization.

6. Simple control circuit. The switch turn-on time is almost constant for complete output load range, hence the control of the switch turn-off time is only required for frequency control.

As shown in Fig. 3, several other rectifiers based on other dc-dc parent converters are derived which have different voltage conversion ratios. The converters shown in Fig. 3(a) and (b) show the buck-boost derived and the flyback derived converters respectively. The dual of SEPIC converter and the forward converter derived versions are also shown in Fig. 3(c) and (d). The leakage inductance of the isolation transformer is effectively series with the resonant tank inductor. Hence, the parasitic inductance problems of the transformer version of the rectifiers are minimized.

III. Principle of Operation

The new single transistor three-phase multi-resonant zero current switching high power factor rectifier shown in Fig. 2 is analyzed. It is sufficient to consider a 30° interval of the ac input line waveform, if the three phase input voltages are symmetric, well balanced and if the switching frequency is much higher than the input line frequency. The 30° interval where \( i_2 > 0 > i_c > i_b \) is described here.

**Interval 1, \( t_0 \leq t \leq t_1 \), all switches are OFF except \( D_d \)**

In this interval, each tank capacitor \( C_{rl-3} \) charges up linearly at a rate proportional to its respective line current. This will continue until the switch \( S_1 \) is turned on. During this moment, the resonant tank inductor \( L_r \) supplies the output load current. When the switch \( S_1 \) turns on, the bridge rectifier input line-to-line voltage \( V_{ab} \) is maximum, forcing \( D_1, D_5 \) to conduct.

**Interval 2, \( t_1 \leq t \leq t_2 \), \( D_1, D_5, D_d, \) and \( S_1 \) are ON**

In this interval, the capacitor voltage \( V_{C3} \) continues to increase, while the other two capacitor voltages ring along with the resonant tank inductor \( L_r \). This will continue until \( V_{C3} \) becomes equal to \( V_{C2} \). Diode \( D_6 \) then also conducts.

**Interval 3, \( t_2 \leq t \leq t_3 \), \( D_1, D_5, D_6, D_d, \) and \( S_1 \) are ON**

In this interval, the three resonant tank capacitors \( C_{rl-3} \) and resonant tank inductor \( L_r \) ring until the tank inductor current increases to zero. Then diode \( D_6 \) turns off, initiating the next interval.

Depending on the magnitudes of \( i_b \) and \( i_c \), the order of interval 2 and 3 may be reversed. Interval 3 occurs before interval 2 when currents \( i_b \) and \( i_c \) are similar in magnitude.

**Interval 4, \( t_3 \leq t \leq t_4 \), \( D_1, D_5, D_6, \) and \( S_1 \) are ON**

In this interval, the three resonant tank capacitors \( C_{rl-3} \), the parallel capacitor \( C_d \), and the resonant tank inductor \( L_r \) form a resonant tank circuit. This interval ends when the resonant tank capacitor voltages \( V_{Crl-3} \) discharges to zero.

**Interval 5, \( t_4 \leq t \leq t_5 \), all switches are ON except \( D_d \)**

In this interval, the parallel capacitor \( C_d \) and the resonant tank inductor \( L_r \) ring until the tank inductor current decreases to the negative load current. At this point, the input bridge rectifiers become reverse biased. Hence, the switch current becomes zero.

**Interval 6, \( t_5 \leq t \leq t_6 \), all switches are OFF**

Interval 6 is actually a subset of interval 1 as shown in the waveforms of Fig. 2. The voltage of capacitor \( C_d \) linearly decreases until it reaches zero voltage and diode \( D_d \) turns on.
IV. Analysis of Single Transistor 3φ Multi-Resonant ZCS HPF Buck-Type Rectifiers

A. Approximate simplified model

According to the symmetry of the 3φ input voltages, the system behavior of the entire line period can be described by the extension of the interval [0, π/6]. To simplify the three phase input and single ended output circuit shown in Fig. 2, one operating point at the time of π/2 is chosen, and hence the simplified single input and single output model is obtained by this assumption. At this moment, the phase voltage $V_{an}$ is the peak voltage of the sine wave input voltage. The phase voltages $V_{an}$ and $V_{cn}$ are both equally one half of the negative $V_{an}$. Hence, these two capacitors are charged and discharged exactly in the same manner at this condition and the capacitors $C_{r2}$ and $C_{r3}$ can be considered as parallel-connected capacitors. Also, the phase current $i_{an}$ equal to its peak value, and $i_{bn}$ and $i_{cn}$ are both $-0.5x_i_{an}$. Hence, the input voltage source and the input filter inductor can be replaced by the current source $I_g$ as shown in Fig. 4, where $I_g$ is the peak phase current $i_{an-peak}$. The input resonant capacitors $C_{r1-3}$ can be replaced by an effective capacitor $C_s$ which is equal to the series connection of $C_{r1}$ and the parallel connected $C_{r2}$ and $C_{r3}$. The diodes $D1$ and $D2$ of Fig. 4 represent the three phase input bridge diodes. The output filter inductor is replaced by current source $I$. Finally, the simplified single input and single ended circuit diagram is shown in Fig. 4.

The relations between the actual three phase input circuit and this simplified single input and single ended output model are described approximately: (a) $C_s = C_s \times 2/3$ where the input resonant capacitors $C_{r1-3}$ have the same values and are represented as $C_s$. (b) $I_g = \text{peak phase current } i_{an-peak}$. (c) $V_g = 3/2 \times \text{peak phase voltage } V_{an-peak}$ where $V_g$ is also the same as the average voltage of $V_{cx}$ during one switching period, and (d) three-phase input power $P_{in} = V_g \times I_g = 3/2 \times (V_{an-peak} \times i_{an-peak})$.

B. Normalization based on the simplified model

In this paper, the converter equations are normalized with respect to the dc output voltage instead of the ac input voltage. This allows the system waveforms to be expressed as functions of the dc operating point. The normalizing base quantities are then described as,

$$\text{base voltage} = V, \quad \text{base current} = V/R_o,$$

$\text{base impedance } R_o = \frac{L_r}{C_x}, \quad \text{base frequency } f_o = \frac{1}{2\pi \sqrt{L_rC_x}}$.

where $V$ is the output voltage of the rectifier. Hence, if the rectifier is an ideal loss free system, then the normalized values of the input and output quantities are described as

$$M_g = \frac{V_g}{V}, \quad J_g = \frac{I_g}{R_o/V}, \quad J = IR_o/V = M_g J_g, \quad \alpha = 2\pi f_o(t_1-t_o), \quad \beta = 2\pi f_o(t_2-t_1), \quad \gamma = (\alpha+\beta)/2,$$

where $t_1$-$t_o$ is the off time of the switch $S1$ and $t_2$-$t_1$ is the on time of the switch $S1$ as shown in Fig. 2.

C. Analysis results

Analysis results are shown in Figs. 5 and 6 in graphical form. Figure 5 shows the input characteristic of the converter, or normalized peak input current $I_g$ vs. normalized peak input voltage $M_g$ for a given $\alpha$. The normalized switch off time $\alpha$ is the control variable for this rectifier; the switch on period is essentially constant for the complete output load range. Hence, one can set the switch on period as a constant value and control the switch off period $\alpha$.

![Fig. 5. The normalized input characteristic of the new single transistor 3φ multi-resonant ZCS HPF rectifier. The normalized transistor off time $\alpha$ is expressed in radians.](image-url)
The characteristics of Figs. 5 and 6 end, for large $I_g$, at the zero current switching boundary. The zero current switching property is lost at large currents. Also, the characteristics are not plotted for $M_g$ less than 1; the multiresonant buck rectifier does not function when $M_g < 1$, i.e., when $3/2V_{an-peak}$ is less than the dc output voltage $V$.

Figure 5 is useful for determining the converter steady-state operating point. Given a specified range of values of peak input voltage and load current, the range of variations of $M_g$ and $I_g$ can be determined from the definitions of Sections IV A and B, and this region is overlayed on Fig. 5. The range of variation of the control parameter $\alpha$ can then be read graphically from the figure. The range of variations of the switching period $\gamma = (\alpha + \beta)/2$ can then be estimated, since the switch conduction angle $\beta$ is nearly constant, and is typically approximately one half tank resonant period, or $\pi$ radians. The switching frequency is therefore approximately $f_s = 2\pi f_c/\gamma$.

Figure 6 shows the voltage stress and the current stress of the switch $S_1$ and the output diode $D_d$ at different operating conditions. Again, voltages and currents are normalized using the base quantities defined in Section IV B. The voltage stress of switch $S_1$ is shown in Fig. 6(a) and the current stress of the switch $S_1$ is shown in Fig. 6(c). At large voltage conversion ratios, the current stress decreases at the light load condition. Figure 6(b) shows the voltage stress of the output diode. The current stress of the resonant tank inductor can also be found by subtracting the normalized output current $J$ ($M_gJ_g = J$) from this current stress of the switch. The output diode voltage stress is the same as the voltage stress of $C_d$ as shown in Fig. 6(b) and the current stress of the output diode is approximately equal to the output current $J$.

**Table 1. Comparison of transistor stresses between 3ϕ multi-resonant rectifier and conventional PWM rectifier**

<table>
<thead>
<tr>
<th>$P = 6kW$</th>
<th>Blocking Voltage (V)</th>
<th>Peak Current (A)</th>
<th>Total Stress (kVA)</th>
<th>Utilization</th>
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</thead>
<tbody>
<tr>
<td>3ϕ multi-resonant</td>
<td>$V_{in} = 240V_{rms}$</td>
<td>685</td>
<td>77.4</td>
<td>53</td>
</tr>
<tr>
<td>6-switch PWM</td>
<td></td>
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<td>83.23</td>
<td>0.072</td>
</tr>
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</table>

**Table 2. Stresses of the switch and the output diode at the maximum and minimum load**

<table>
<thead>
<tr>
<th>Output Power [kW]</th>
<th>Switch Voltage $V_{ps}[V]$</th>
<th>Current $I_{ps}[A]$</th>
<th>Diode Voltage $V_{pd}[V]$</th>
<th>Peak Diode Current $I_{pd}[A]$</th>
<th>$I_{ps}R_{di}[V]$</th>
</tr>
</thead>
<tbody>
<tr>
<td>6</td>
<td>685</td>
<td>4.66</td>
<td>77.4</td>
<td>5.16</td>
<td></td>
</tr>
<tr>
<td>0.6</td>
<td>640</td>
<td>4.35</td>
<td>43.6</td>
<td>2.91</td>
<td></td>
</tr>
<tr>
<td>Output Power [kW]</td>
<td>Diode Voltage $V_{pd}[V]$</td>
<td>Current $I_{pd}[A]$</td>
<td>Diode Voltage $V_{pd}[V]$</td>
<td>Peak Diode Current $I_{pd}[A]$</td>
<td>$I_{ps}R_{di}[V]$</td>
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</tr>
<tr>
<td>6</td>
<td>461</td>
<td>3.14</td>
<td>41</td>
<td>2.73</td>
<td></td>
</tr>
<tr>
<td>0.6</td>
<td>551</td>
<td>3.75</td>
<td>43.1</td>
<td>0.287</td>
<td></td>
</tr>
</tbody>
</table>

Fig. 6. The normalized stresses of the new single transistor new single transistor 3ϕ multi-resonant ZCS HPF rectifier: (a) voltage stress of the switch $S_1$, (b) voltage stress of the output diode $D_d$, and (c) current stress of the switch $S_1$. The current stress of the output diode is equal to the output current.
impedance $R_o$ is $9.8\Omega$. The total harmonic distortion of the input current becomes $0.98\%$ at minimum load and $4.38\%$ at maximum load. Figure 7 shows the input current and the input voltage waveforms at the maximum load and the minimum load.

The peak switch voltages and the peak currents of the new rectifier and six switch PWM approaches are compared in Table 1 for a $6kW 3\phi$ac $240V_{L-L}$ application. Total switch stress summed over all active switches in the rectifiers are shown in Table 1. Also shown is the silicon utilization, defined as the converter output power divided by the total switch stress [2]. Table 2 shows the stresses of the switch and the output diode at the maximum load and the minimum load. It can be seen that the active device silicon is utilized more effectively by the proposed multi-resonant rectifier than by the conventional six-switch PWM approach. The penalty in component stress due to the resonant approach is more than compensated for by the fact that there is only one device. The normalized stresses shown in Table 2 agree well with the analysis results shown in Fig. 6.

B. Experimental Results

The new single-switch $3\phi$ multi-resonant ZCS HPF rectifier as shown in Fig. 2 was built to provide experimental proof of the proposed approach. The dc output voltage $V$ is $62V$ with an rms ac input voltage of $100V_{L-L}$. The maximum output power $0.36kW$ is obtained at the switching frequency $56kHz$ with switch on time $11\mu s$, and the minimum output power $36W$ is obtained at the switching frequency $19.2kHz$ with switch on time $11\mu s$. The input filter inductors $L_a, L_b$, and $L_c$ are $1mH$ each and the output filter inductor is $1.1mH$. The input side resonant tank capacitors $C_{1,2,3}$ are connected in a delta configuration with $45nF$ each, and are equivalent to the $Y$-connection with $135nF$ each. Hence, the effective capacitor $C_2$ as mentioned in Section IV is $90nF$. The value of the output side resonant tank capacitor $C_d$ is chosen to be equal to $C_x$ or $90nF$. This choice leads to a good compromise between low transistor voltage stress and low input current harmonics. The resonant tank inductor $L_t$ is $78\mu H$. Hence, the resonant frequency $f_o$ is $59.8kHz$ and characteristic impedance $R_o$ is $29.6\Omega$. The

<table>
<thead>
<tr>
<th>Table 3. Measured stresses of the switch and the output diode at the maximum and minimum load</th>
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<tbody>
<tr>
<td>output power [W]</td>
</tr>
<tr>
<td>------------------</td>
</tr>
<tr>
<td>360</td>
</tr>
<tr>
<td>36</td>
</tr>
<tr>
<td>360</td>
</tr>
<tr>
<td>36</td>
</tr>
</tbody>
</table>
Fig. 8. Measured waveforms of the input line current with its ac phase voltage for experimental rectifier circuit: (a) at maximum output power 0.36kW (vertical scale: 20V/div, 1A/div and horizontal scale: 2msec/div), (b) at minimum output power 36W (vertical scale: 20V/div, 0.5A/div and horizontal scale: 2msec/div).

load resistance $R_L = 10.8\Omega$ is connected for the maximum load and $R_L = 108\Omega$ is connected for the minimum load. Figure 8 shows the measured waveforms of the input line current with its ac phase voltage and Table 3 shows the measured stresses of the switch and the output diode at maximum and minimum load conditions. Experimental results show the feasibility of the wide load range control with low stresses and agree well with the simulation and the analysis results.

VI. Conclusion

In this paper, a new family of 3φ single-switch high power factor multi-resonant buck-type rectifiers has been introduced. These rectifiers are capable of drawing a high quality input current waveform at nearly unity power factor and have continuous input and output currents. By the use of a multi-resonant scheme, the transistor operates with zero current switching and the diodes operate with zero voltage switching. Moreover, these rectifiers have a wide load range and low stress on the semiconductor devices.

The input characteristics of these rectifiers are derived and explain well how the open-loop line current waveform is naturally proportional to the input voltage. The normalized component stresses are derived and presented in Section IV. The wide range of the output power variation and near unity power factor are verified by the simulation and the experimental results.

References


