Quasi-Resonant Boost-Half-Bridge Converter with Reduced Turn-off Switching Losses for 16V Fuel Cell Application

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Abstract — The active-clamped current-fed converter is able to achieve not only lossless clamping but ZVS operation of switches. Among them, a boost-half-bridge(BHB) converter [7] is one of the most suitable candidate for high current high step up applications due to the advantages of low transformer turn ratio, reduced voltage rating of diodes, zero magnetizing dc offset and symmetrical structure of all components. In this paper a quasi-resonant(QR) switching technique for the BHB converter with active clamping is proposed to reduce turn off switching losses. Experimental results on a 16 V 1.2 kW prototype are provided to validate the effectiveness of the proposed concept.

Keywords- Current-fed; Quasi-resonant; Boost-half-bridge; Fuel cell

I. INTRODUCTION

The isolated boost DC-DC converter has been increasingly needed in many applications such as fuel cell systems, photovoltaic systems, hybrid electric vehicles, and UPS where the use of high frequency transformers for high step-up ratio and galvanic isolation are required. Due to the advantages of smaller input current ripple, lower diode voltage rating and lower transformer turns ratio, the current-fed converter is better suited to high step-up applications. The active-clamped current-fed converter has been introduced based on three topologies: push-pull [1,2], full-bridge [3,4] and L-type half-bridge [5-7]. They achieve not only lossless clamping of voltage spikes across the switch caused by leakage inductance of the transformer but ZVS operation of switches. The active-clamped push-pull converter has the simplest structure among them, but center-tapping of the transformer at the low voltage, high current side could be a challenge in manufacturing. The active-clamped full-bridge converter requires an additional switch for clamping, and the switching frequency of the clamp switch should be twice that of the main switch. The active-clamped L-type half-bridge converter does not need either additional clamp switch as in the active-clamped full-bridge converter or transformer center-tapping as in the active-clamped push-pull converter. Due to the circuit structure of parallel-connection of two BHB converters and series-connection of two voltage doubler rectifiers via two transformers, an improved active-clamped current-fed converter [7] is one of the most suitable candidate for high step up applications. The active-clamped BHB converter proposed in [7] has the following advantages over the other active-clamped current-fed converter: 1) low turn ratio of the high frequency transformer, 2) reduced voltage rating of diodes which results in possible use of inexpensive Schottky diodes, 3) no dc offset of the magnetizing current 4) low profile and easy thermal distribution due to symmetrical nature of all components. However, a major disadvantage of the active-clamped BHB converter [7] is high turn-off current of switches.

In this paper a QR switching technique is introduced for the active-clamped BHB converter [7] to reduce the turn off switching losses. The clamp capacitors are used as a resonant capacitor for the QR operation during switch turn-on process so as to reduce the turn-off current of the switches. The operating principle of the proposed method is in detailed, and experimental results on a 16 V 1.2 kW prototype are provided to validate the proposed scheme.

II. OPERATING PRINCIPLES

The circuit topology of the proposed converter is shown in Fig. 1. It consists of two BHB converters connected in parallel at primary side and two voltage doubler rectifiers connected in series at secondary side.

If each BHB converter shares the clamp capacitor the voltage-second unbalance of two input inductors resulting from duty cycle mismatch or difference in circuit parameters may cause current unbalance between the two parallel connected BHB converters. As shown in Fig. 1, separate use of clamp capacitors for each BHB converter helps mitigate this problem without using current sensors. Furthermore, due to the capacitor connection at both sides of the transformer there must be no dc offset in magnetizing current. In the proposed converter capacitors \(C_{r1} \sim C_{r4}\) are used not only as a clamp capacitor but also as a resonant capacitor with resonant inductors \(L_{r1}\) and \(L_{r2}\) during switch turn-on process so as to reduce the turn-off current of the switches. The resonant frequencies of \(L_{r1}C_{r1}\) and \(L_{r2}C_{r2}\) are defined by, respectively,
\[
f_{1} = \frac{1}{T_{1}} = \frac{\omega_{r1}}{2\pi} = \frac{1}{2\pi} \sqrt{\frac{L_{1}}{C_{r1}}} \tag{1}
\]
\[
f_{2} = \frac{1}{T_{2}} = \frac{\omega_{r2}}{2\pi} = \frac{1}{2\pi} \sqrt{\frac{L_{1}}{C_{r2}}} \tag{2}
\]

Figure 2 shows key waveforms of the proposed converter for illustration of the operating principle. The two legs at the low voltage side are interleaved with 180° phase shift resulting in reduced input current ripples, and the upper and lower switches of each leg are operated with asymmetrical complementary switching to regulate the output voltage. Figure 3 shows equivalent circuits of the five operating modes.

**Mode 1 [t1-t2]**: Switch \( S_{1} \) is turned off at \( t_{1} \). External capacitor \( C_{q1} \) across \( S_{1} \) is charged and parasitic capacitor \( C_{oss} \) of \( S_{2} \) is discharged, respectively, by \( i_{r1}(t) \). Then the body diode of \( S_{2} \) is turned on. Switch voltage \( V_{S2} \) increases linearly with a slope of \( i_{r1}(t) / (C_{q1} + 2C_{oss}) \). Switch \( S_{1} \) is almost turned off with ZVS if external capacitor \( C_{q1} \) is chosen large enough to hold the switch voltage \( V_{S1} \) at near zero at the switching instant. The dead time required to achieve ZVS turn on of \( S_{2} \) is determined by,

\[
t_{d,s2} \approx \frac{V_{m}(2C_{oss} + C_{q1})}{i_{r1}(t) - (1-D)} \tag{3}
\]

This state ends at the instant reflected capacitor voltage to secondary \( n \cdot V_{C2} \) becomes greater than capacitor voltage \( V_{col} \). The duration of this mode \( t_{e} \) can be obtained by solving the following equation,

\[
\frac{(C_{q1} + C_{oss})}{C_{oss} - C_{q1}} \cdot \sin(\omega_{r2}(1-D)t_{e} - t_{1}) + t_{1} = 0 \tag{4}
\]

**Mode 2 [t2-t3]**: At \( t_{2} \) inductor \( L_{1} \) starts to resonate with \( C_{q2} \) and diode \( D_{1} \) starts conducting. Switch \( S_{1} \) carries difference between input inductor current \( i_{r1}(t) \) and resonant inductor current \( i_{r1}(t) \). Fig. 4(a) shows equivalent resonant circuit assuming that input inductor current is constant during this mode. The resonant inductor current is obtained by,

\[
i_{r1}(t) = I_{ds}(1 - \cos \omega_{r2} \cdot t) \tag{5}
\]

This state ends when the gate signal is removed from \( S_{2} \) at \( t_{3} \). It should be noted that turn off current of \( S_{2} \) is smaller than that of the conventional scheme[7].

**Mode 3 [t3-t4]**: At \( t_{3} \) external capacitor \( C_{q3} \) across \( S_{1} \) is discharged and parasitic capacitor \( C_{oss} \) of \( S_{1} \) is charged, respectively, by \( i_{r1}(t) - i_{r2}(t) \). Then, the body diode of \( S_{3} \) is turned on. Switch voltage \( V_{S3} \) increases linearly with a slope of \( (i_{r1}(t) - i_{r2}(t)) / (C_{q3} + 2C_{oss}) \). Switch \( S_{2} \) is almost turned off with ZVS if external capacitor \( C_{q3} \) is chosen large enough to hold the switch voltage \( V_{S3} \) at near zero at the switching instant. The dead time required to achieve ZVS turn on of \( S_{1} \) is determined by,

\[
t_{d,s1} \approx \frac{V_{m}(2C_{oss} + C_{q3})}{\left(i_{r1}(t) - i_{r2}(t)\right) - (1-D)} \tag{6}
\]

The diode current \( i_{D1} \) decreases linearly with a slope of \( \left(V_{C1} + (V_{D1} + \frac{V_{oss}}{2}) \right) / L_{oss} \). Diode \( D_{1} \) could be turned off with ZCS when \( L_{oss} \) is moderately large. This state ends when current \( I_{D1} \) becomes 0A.
where the peak value of the resonant current $I_{r, pk}$ is

$$I_{r, pk} = \frac{n \cdot i_{r}}{2C_{r} \cdot f_{c}} \sqrt{\frac{C_{r}}{L_{r}}}$$  \hspace{1cm} (0.5T_{n} \leq D \cdot T_{s}) \hspace{1cm} (8)$$

$$I_{r, pk} = \frac{n \cdot i_{r}}{C_{r} \cdot f_{c} \cdot \cos(D \cdot T_{s} \cdot \omega_{s})} \sqrt{\frac{C_{r}}{L_{r}}}$$  \hspace{1cm} (0.5T_{n} > D \cdot T_{s}) \hspace{1cm} (9)$$

This resonant mode ends when current $i_{r}(t)$ becomes 0A. Note that diode $D_{2}$ is turned off with ZCS.

**Mode 5 [t_{5} \sim t_{6}]:** During this mode all currents are zero except that switch $S_{2}$ carries input inductor current. This state ends when the gate signal is removed from $S_{1}$ at $t_{6}$. It should be noted that turn off current of $S_{2}$ is the same as the input inductor current at $t_{6}$ which is smaller than that of the conventional scheme[7]. The other half cycle begins at time $t_{6}$ and is repeated except with the correspondingly opposite set of legs.

III. DESIGN METHODOLOGY

In the proposed converter capacitor $C_{r1}$ resonates with inductor $L_{r1}$ at turn-on of switch $S_{1}$ while capacitor $C_{r2}$ resonates with inductor $L_{r2}$ at turn-on of switch $S_{2}$. In this section a design example is provided to determine the optimal resonant frequencies of resonant tanks $L_{r1}-C_{r1}$ and $L_{r1}-C_{r2}$. The example specifications are as follows:

- $P_{o} = 1140$ W,
- $V_{in} = 12.8 \sim 26$ V (nominal input : 16 V),
- $V_{out} = 360$ V,
- $D = 0.29 \sim 0.69$ (nominal duty $D_{n}$ : 0.6),
- $f_{s} = 30$ kHz,
- $\Delta I_{in} = 3\%$,
- $\Delta V_{out} = 2\%$

Figure 5 shows switch current waveforms according to variation of resonant periods $T_{r1}$ and $T_{r2}$. In cases of (i) and (ii) the turn-off currents of switch $S_{1}$ are the same, but it is larger for (iii). Similarly, in cases of (iv) and (v) the turn-off currents of switch $S_{2}$ are zero, but in case of (vi) there exists a magnitude of turn off current.

$$\begin{align*}
\text{(i)} & \quad 0.5T_{r1} < D \cdot T_{s} \\
\text{(ii)} & \quad 0.5T_{r1} = D \cdot T_{s} \\
\text{(iii)} & \quad 0.5T_{r1} > D \cdot T_{s}
\end{align*}$$

$$\begin{align*}
\text{(iv)} & \quad 0.75T_{r2} < (1-D)T_{s} - t_{a} \\
\text{(v)} & \quad 0.75T_{r2} = (1-D)T_{s} - t_{a} \\
\text{(vi)} & \quad 0.75T_{r2} > (1-D)T_{s} - t_{a}
\end{align*}$$

Figure 5 Switch current waveforms according to variation of resonant period $T_{r1}$ and $T_{r2}$.
Table I. Derived equations of RMS current and turn off current of switches according to variation of resonant period

<table>
<thead>
<tr>
<th>Resonant period</th>
<th>RMS current</th>
<th>Turn off current</th>
</tr>
</thead>
<tbody>
<tr>
<td>Lower switches</td>
<td></td>
<td></td>
</tr>
<tr>
<td>$0.5T_{on} \leq DT_4$</td>
<td>$I_{1on} = \sqrt{\frac{\frac{1}{T_4} - \frac{1}{T_5}}{\frac{1}{T_4} - \frac{1}{T_5}} + \frac{D}{4} - \frac{2D}{\pi} + \frac{D}{16}}$</td>
<td>$I_{1on}$</td>
</tr>
<tr>
<td>$0.5T_{on} &gt; DT_4$</td>
<td>$I_{1on} = \sqrt{\frac{\frac{1}{T_4} - \frac{1}{T_5}}{\frac{1}{T_4} - \frac{1}{T_5}} + \frac{D}{4} - \frac{2D}{\pi} + \frac{D}{16}}$</td>
<td>$I_{1on}$</td>
</tr>
<tr>
<td>upper switches</td>
<td></td>
<td></td>
</tr>
<tr>
<td>$0.75T_{on} \leq (1-D)T_{on} - t_o$</td>
<td>$I_{1on} = \sqrt{\frac{\frac{1}{T_4} - \frac{1}{T_5}}{\frac{1}{T_4} - \frac{1}{T_5}} + \frac{D}{4} - \frac{2D}{\pi} + \frac{D}{16}}$</td>
<td>$I_{1on}$</td>
</tr>
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<td>$0.75T_{on} &gt; (1-D)T_{on} - t_o$</td>
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<td>$I_{1on}$</td>
</tr>
</tbody>
</table>

Table I list the derived equations of RMS current and turn off current of switches according to the resonant period variation. Using equations of Table I, the RMS current and turn off current of switches as a function of resonant frequency are shown in Fig. 6. As shown in Fig. 6(a), the RMS current of the lower switch tends to slightly increase as the resonant frequency increases. The turn off current of lower switch is small for case (i) in which the resonant frequency is higher than 27 kHz. The turn off current rapidly increases as the resonant frequency decreases. Therefore, in this design example the optimal resonant frequency for lower switch $f_{on}$ is 27 kHz and determined by,

$$f_{on} = \frac{f_s}{2D_s} \quad (10)$$

Fig. 6(b) shows the RMS current of the upper switch is relatively constant without regard to variation of resonant frequency. The turn off current of upper switch is zero for case (iv) in which the resonant frequency is higher than 61 kHz. The turn off current rapidly increases as the resonant frequency decreases. Therefore, in this design example the optimal resonant frequency for upper switch $f_{on}$ is 61 kHz and determined by,

$$f_{on} = 3 \left[ \frac{\pi}{4} - \frac{f_s}{2D_s - f_s-t_o} \right] \quad (11)$$

### IV. EXPERIMENTAL RESULT

In order to verify the effectiveness of the proposed converter, 1.2 kW laboratory prototypes of both the PWM scheme [7] and the proposed QR scheme have been constructed with the specifications and parameters which are used in section III.

The turn ratio of the transformer is 1:5. The leakage inductance of transformer which is measured as 0.4 $\mu$H has been used as the resonant inductance $L_{res}$. From (10) and (11), the resonant capacitances are calculated as $C_{r1} = 90 \mu F$ and $C_{r2} = 17 \mu F$. Actual resonant capacitances used for the experiment are $C_{r1} = 95 \mu F$ and $C_{r2} = 20 \mu F$ (Pilkor Electronics).

Fig. 7 shows experimental waveforms obtained at 1140 W load. Figs. 7(a) and (b) show the drain-source voltage of lower switch $S_1$ and current $I_{L1} - I_{r1}$ of the PWM scheme and QR scheme, respectively. Figs. 7(c) and (d) show the drain-source voltage of upper switch $S_2$ and current $I_{L1} - I_{r1}$ of the PWM scheme and QR scheme, respectively. It can be seen that both lower and upper switches are being turned on with ZVS and turn-off currents of the proposed QR scheme are significantly reduced compared to those of the PWM scheme.

The measured efficiencies of both QR and PWM schemes using Yokogawa’s power analyzer WT3000 are shown in Fig. 8. The peak and full load efficiencies of the proposed QR scheme with $C_{eq}$ is 96.3% at 310 W and 93.7% at 1140 W, respectively. Efficiency improvement of the QR schemes over the PWM scheme is mostly resulting from reduced turn-off losses associated with the proposed resonance operation and the use of external capacitors $C_q$ across lower switches $S_1$ and $S_2$. Due to the external capacitor $C_q$, lower switches are turn on with hard switching at light load, which results in decreased...
light-load efficiency. However, this is not a problem in the fuel cell application where full load efficiency should be optimized since the system is operated at full load during most of time.

Fig. 7 Experimental waveforms at 1140W load (a) drain-source voltage of $S_1$ and current $i_{S1}$ of PWM scheme (b) drain-source voltage of $S_1$ and current $i_{S1}$ of QR scheme (c) drain-source voltage of $S_2$ and current $i_{S2}$ of PWM scheme (d) drain-source voltage of $S_2$ and current $i_{S2}$ of QR scheme

![Efficiency vs Load Power](image)

**Fig. 8** Measured efficiencies of the QR and PWM schemes

**V. CONCLUSIONS**

This paper proposes an improved switching method for an active-clamped BHB converter for high step-up application. The clamp capacitors are also used as a resonant capacitor for the QR operation during switch turn-on process so as to reduce the turn-off current of the switches. A small external capacitor across lower switches helps further reduce the turn-off switching losses. Experimental results on a 1.2kW prototype demonstrated 2.8% of full load efficiency improvement of the proposed scheme.

**REFERENCES**