Abstract—A bidirectional isolated full-bridge dc–dc converter with a conversion ratio around nine times, soft start-up, and soft-switching features for battery charging/discharging is proposed in this paper. The converter is equipped with an active flyback and two passive capacitor–diode snubbers, which can reduce voltage and current spikes and reduce voltage and current stresses, while it can achieve near zero-voltage-switching and zero-current-switching soft-switching features. In this paper, the operational principle of the proposed converter is first described, and its analysis and design are then presented. A 1.5-kW prototype with a low-side voltage of 48 V and a high-side voltage of 360 V has been implemented, from which measured results have verified the discussed features.

Index Terms—Bidirectional, snubbers, soft switching.

I. INTRODUCTION

In renewable dc supply systems, batteries are usually required to back up power for electronic equipment. The voltage levels of the batteries are typically much lower than the dc-bus voltage. Bidirectional converters for charging/discharging the batteries are therefore required. In past studies, bridge-type bidirectional isolated converters have been widely applied to fuel cell and electric vehicle driving systems [1]–[6]. For raising power level, a dual full-bridge configuration is usually adopted [7]–[18], and their low and high sides are typically configured with boost- and buck-type topologies, respectively. However, component stress, switching loss, and electromagnetic interference (EMI) noise are increased due to diode-reverse-recovery current and MOSFET drain–source voltage, resulting in low reliability. A more severe issue is due to leakage inductance of the isolation transformer, which will result in high-voltage spikes during switching transition. A possible solution is to pre-excite the leakage inductance to raise its current level up to that of the current-fed inductor, which can reduce their current difference and, in turn, reduce voltage spike. However, since the current level varies with load condition, it is hard to tune the switching timing to match these two currents. Thus, a passive or an active snubber circuit is still needed.

Passive and active clamp circuits were proposed to suppress the voltage spike due to the current difference between the current-fed inductor and leakage inductance currents [13], [14]. A conventional passive approach is employing a resistor–capacitor–diode snubber to clamp the voltage, and the energy absorbed in the buffer capacitor is dissipated on the resistor, resulting in low efficiency. On the other hand, a simple active clamping circuit [13] was proposed, which is shown in Fig. 1. However, the resonant current will flow through the main switches, increasing current stress significantly. An isolated bidirectional converter with a flyback snubber was therefore proposed [18], [19], which is shown in Fig. 2. The flyback snubber can recycle the absorbed energy which is stored in the clamping capacitor $C_C$, while without current flowing through the main switches. It can also clamp the voltage to a desired value just slightly higher than the voltage across the low-side transformer. Since the snubber current does not circulate through the main switches, current stress can be reduced a lot under heavy-load condition. Furthermore, the flyback snubber can be controlled to precharge the high-side capacitor to avoid in-rush current during a start-up period. However, the low- and high-side switches are operated with hard switching turnover, resulting in high-voltage spikes.

To solve the aforementioned problem, we first introduce two buffer capacitors ($C_{b1}$ and $C_{b2}$) connected in parallel with the upper legs of the voltage-fed bridge, as shown in Fig. 3. With
these two buffer capacitors, the low- and high-side switches can operate with near zero-voltage switching (ZVS) and zero-current switching (ZCS). However, when it is operated in step-down conversion, these capacitors will resonate with leakage inductance of the transformer, causing EMI noise and increasing switching loss. Thus, two passive capacitor–diode snubbers are proposed to supplement the active flyback snubber, as shown in Fig. 4. The proposed snubber configuration cannot only reduce the voltage spike caused by the current difference between the leakage inductance and current-fed inductor currents but also can relieve the drawbacks of high-current and high-voltage stresses imposed on the main switches at both turn-on and turnoff transitions. Moreover, it can achieve near ZVS and ZCS for the switches on both sides of the transformer.

II. CONFIGURATION AND OPERATION

The proposed soft-switching bidirectional isolated full-bridge converter with an active flyback and two passive capacitor–diode snubbers is shown in Fig. 4. It can be operated with two types of conversions: step-up conversion and step-down conversion. Fig. 4 consists of a current-fed switch bridge, an active flyback snubber at the low-voltage side, a voltage-fed switch bridge, and a passive snubber pair at the high-voltage side. Inductor $L_m$ performs output filtering when power flows from the high-voltage side to the low-voltage side, which is denoted as a step-down conversion. On the other hand, it works in the step-up conversion. Moreover, snubber capacitor $C_C$ and diode $D_C$ are used to absorb the current difference between current-fed inductor current $i_L$ and leakage inductance current $i_P$ of isolation transformer $T_P$ during switching commutation. The flyback snubber is operated to transfer the energy stored in snubber capacitor $C_C$ to buffer capacitors $C_{b1}$ and $C_{b2}$, and voltage $V_C$ can drop to zero. Thus, the voltage stresses of switches $M_1 \sim M_4$ can be limited to a lower level, achieving near ZCS turnoff. The main merits of the proposed snubber include no spike current circulating through the switches and achieving soft-switching features. Note that high spike current can result in charge migration, over current density, and extra magnetic force which will deteriorate in MOSFET carrier density, channel width, and wire bonding and, in turn, increase its conduction resistance.

In the step-up conversion, switches $M_1 \sim M_4$ are operated with current-fed inductor current $i_L$, and the body diodes of switches $M_5 \sim M_8$ serve as a rectifier. In the step-down conversion, switches $M_5 \sim M_8$ are controlled, and the body diodes of switches $M_1 \sim M_4$ operate as a full-bridge rectifier. To simplify the steady-state analysis, several assumptions are made as follows.

1) All components are ideal except that the transformer is associated with leakage inductance.
2) Inductor $L_m$ is large enough to keep the current $i_L$ constant over a switching period.
3) Snubber capacitor $C_C$ is much larger than the parasitic capacitance of switches $M_1 \sim M_8$.

A. Step-Up Conversion

In the step-up conversion, switches $M_1 \sim M_4$ are operated like a boost converter, where switch pairs $(M_1, M_2)$ and $(M_3, M_4)$ conduct to store energy in $L_m$. At the high-voltage side, body diodes $D_5 \sim D_8$ of switches $M_5 \sim M_8$ will conduct to transfer power to $C_{HV}$. When switch pairs $(M_1, M_2)$ and $(M_3, M_4)$ are switched to $(M_1, M_4)$ or $(M_2, M_3)$, current difference $i_C = i_L - i_P$ will charge capacitor $C_C$ until $i_P$ rises up to $i_L$, and capacitor voltage $V_C$ will be clamped to $V_{HV} \cdot N_P / N_S$, achieving near ZCS turnoff for $M_2$ or $M_4$. In the meantime, high-side current $i_S$ has the priority flowing through one of the two passive capacitor–diode snubbers, and either $C_{b1}$ or $C_{b2}$ will be fully discharged before diode $D_3$ or $D_7$ conducts. When switch pair $(M_1, M_4)$ or $(M_2, M_3)$ is switched back to $(M_1, M_2)$ and $(M_3, M_4)$, switch $M_2$ or $M_4$ can have near ZCS turn-on feature due to leakage inductance $L_l$ limiting the $di/dt$ of high-side diode-reverse-recovery current. The flyback snubber operates simultaneously to discharge snubber capacitor $C_C$ and transfer the stored energy to buffer capacitors $C_{b1}$ and $C_{b2}$. With the flyback snubber, the energy absorbed in $C_C$ will not flow through switches $M_1 \sim M_4$, which can reduce their current stresses dramatically when the leakage inductance of the isolation transformer is significant.

The key voltage and current waveforms of the converter operated in the step-up conversion are shown in Fig. 5. A detailed description of the converter operation over a half-switching cycle is presented as follows.

1) **Mode 1** [$t_0 \leq t < t_1$]: Before $t_0$, all of the four switches $M_1 \sim M_4$ are turned on. Inductor $L_m$ is charged by $V_{HV}$. At $t_0$, $M_1$ and $M_4$ remain conducting, while $M_2$ and $M_3$ are turned off. Then, clamping diode $D_C$ conducts, and snubber capacitor $C_C$ is charged by the current difference $i_C$. In this mode, the flyback snubber still stays in the OFF state. The equivalent circuit is shown in Fig. 6(a).

2) **Mode 2** [$t_1 \leq t < t_2$]: In this mode, leakage inductance current $i_P$ will start to track current $i_L$, and buffer capacitor $C_{b1}$ will start to release energy. At time $t_2$, current $i_P$ is equal to current $i_L$, the voltage of switches $M_2$ and $M_3$ and capacitor $C_C$ will reach the maximum value simultaneously, and its equivalent circuit is shown in Fig. 6(b).

---

**Fig. 3.** Bidirectional isolated full-bridge dc–dc converter with a flyback snubber and parallel snubber capacitors $C_{b1}$ and $C_{b2}$ (type B).

**Fig. 4.** Proposed soft-switching bidirectional isolated full-bridge converter with an active flyback and two passive capacitor–diode snubbers (type C).
Fig. 5. Key voltage and current waveforms of the proposed converter operated in the step-up conversion.

Fig. 6(b). A near ZCS soft switching is therefore attained during $t_0$ to $t_2$.

3) **Mode 3** [$t_2 \leq t < t_3$]: Before $t_3$, the energy stored in buffer capacitor $C_{b1}$ is not fully discharged yet. Thus, the capacitor will not stop discharging until $V_{b1}$ drops to zero. The equivalent circuit is shown in Fig. 6(c).

4) **Mode 4** [$t_3 \leq t < t_4$]: When the energy stored in $C_{b1}$ has been completely released to the output at $t_3$, diode $D_5$ will conduct. The circuit operation over this time interval is identical to a regular turnoff state of a conventional current-fed full-bridge converter. The equivalent circuit is shown in Fig. 6(d).

5) **Mode 5** [$t_4 \leq t < t_5$]: At $t_4$, all of the four switches $M_1 \sim M_4$ are turned on again, and switch $M_S$ of the flyback snubber is turned on synchronously. Switches $M_2$ and $M_3$ achieve a ZCS turn-on soft-switching feature due to $L_{ll}$, and current $i_P$ drops to zero gradually. In the flyback snubber, the energy stored in capacitor $C_C$ will be delivered to the magnetizing inductance of transformer $T_S$. The equivalent circuit is shown in Fig. 6(e).

6) **Mode 6** [$t_5 \leq t < t_6$]: When switch $M_S$ is turned off at $t_5$, capacitor voltage $V_C$ drops to zero, and the energy stored in the magnetizing inductance will be transferred to buffer capacitor $C_{b1}$. In this mode, the time interval of driving signal $V_{gs(M_S)}$ is slightly longer than the discharging time of capacitor $C_C$. The purpose is to ensure that the energy stored in capacitor $C_C$ can be completely released, creating a ZCS operational opportunity for switch $M_2$ or $M_4$ at the next turnoff transition. The equivalent circuit is shown in Fig. 6(f).

7) **Mode 7** [$t_6 \leq t < t_7$]: At $t_6$, the energy stored in the magnetizing inductance of transformer $T_S$ was completely transferred to buffer capacitor $C_{b1}$, and the circuit operation is identical to a regular turn-on state of a conventional current-fed converter. Its equivalent circuit is shown in Fig. 6(g). The circuit operation stops at $t_7$ and completes a half-switching cycle.
B. Step-Down Conversion

In the analysis, the leakage inductance of the transformer at the low-voltage side is reflected to the high-voltage side, in which equivalent inductance $L_{eq}$ equals $(L_{th} + L_{tt} \cdot N_h^2/N_p^2)$.

In the step-down conversion, switches $M_5 \sim M_8$ are operated like a buck converter in which switch pairs ($M_5$, $M_6$) and ($M_7$, $M_8$) take turns conducting to transfer power from capacitor $C_{HV}$ to battery $B_{LV}$. For alleviating leakage inductance effect on voltage spike, switches $M_5 \sim M_8$ are operated with phase-shift control, achieving ZVS turn-on features. Although there is no need to absorb the current difference between $i_L$ and $i_P$, capacitor $C_C$ can help clamp the voltage ringing due to $L_{eq}$ and the parasitic capacitance of $M_1 \sim M_4$. The two passive capacitor–diode snubbers, switches $M_6$ and $M_8$ can achieve near ZCS turnoff.

The key voltage and current waveforms of the converter operated in the step-down conversion are shown in Fig. 7. A detailed description of its operation over a half-switching cycle is presented as follows.

1) **Mode 1** ($t_0 \leq t < t_1$): In this mode, switches $M_5$ and $M_6$ are turned on, while $M_7$ and $M_8$ are in the OFF state. The high-side voltage $V_{HV}$ is crossing the transformer, and it is, in fact, crossing the equivalent inductance $L_{eq}$ and drives current $i_S$ rise with the slope of $V_{HV}/L_{eq}$. With the transformer current increasing toward the load-current level at $t_1$, the body diodes ($D_1$ and $D_4$) are conducting to transfer power, and the voltage across the transformer terminals on the low-voltage side changes immediately to reflect the voltage from the high-voltage side. The equivalent circuit is shown in Fig. 8(a).

2) **Mode 2** ($t_1 \leq t < t_2$): At $t_1$, switch $M_8$ remains conducting, while $M_5$ is turned off. The body diode of $M_6$ then starts conducting the freewheeling leakage current. The transformer current $i_S$ reaches the load-current level at $t_1$, and $V_{AB}$ rises to the reflected voltage ($V_{HV} \cdot N_p/N_S$). Clamping diode $D_C$ starts conducting the resonant current of $L_{eq}$ and the parasitic capacitance of $M_1 \sim M_4$. At the same time, switch $M_S$ of the flyback snubber is turned on and starts transferring the energy stored in capacitor $C_C$ to buffer capacitors $C_{b1}$ and $C_{b2}$. The process ends at $t_2$ when the resonance goes through a half resonant cycle and is blocked by clamping diode $D_C$. With the flyback snubber, the voltage of capacitor $C_C$ will be clamped to a desired level just slightly higher than the voltage of $V_{ds(M4)}$. The equivalent circuit is shown in Fig. 8(b).

3) **Mode 3** ($t_2 \leq t < t_3$): At $t_2$, the body diode of switch $M_6$ is conducting, and switch $M_b$ can be turned on with ZVS. The equivalent circuit is shown in Fig. 8(c).

4) **Mode 4** ($t_3 \leq t < t_4$): At $t_3$, switch $M_6$ remains conducting, while $M_8$ is turned off. Buffer capacitor $C_{b2}$ is discharging by the freewheeling current. When $C_{b2}$ is fully discharged, a near ZCS turnoff condition is therefore attained, and the body diode of $M_7$ then starts conducting the freewheeling current. The equivalent circuit is shown in Fig. 8(d).
5) Mode 5 \([t_4 \leq t < t_5]\): At \(t_4\), with the body diode of switch \(M_7\) conducting, \(M_7\) can be turned on with ZVS. Over this time interval, the active switches change to the other pair of switches, and the voltage across the transformer reverses its polarity. The circuit operation stops at \(t_5\) and completes a half-switching cycle. The equivalent circuit is shown in Fig. 8(e).

### III. DESIGN AND PRACTICAL CONSIDERATION OF SNUBBERS

The purpose of using an active flyback snubber is to transfer energy from snubber capacitor \(C_C\) to buffer capacitors \(C_{b1}\) and \(C_{b2}\), which can attain a near ZCS soft-switching feature. To reduce high-voltage spike occurring on switch \(M_{S}\), the flyback snubber is operated in discontinuous conduction mode, and the key components of the proposed snubber are designed as follows.

#### A. Snubber Capacitor \(C_C\)

For clamping the switch voltage at the low-voltage side, snubber capacitor \(C_C\) needs to satisfy the following inequality:

\[
C_C \geq \frac{L_{eq}}{V^2_C} \left( i_L - i_p \right)^2
\]  

(1)

where \(L_{eq} = L_{sll} + L_{zh} \cdot N^2_p / N^2_s\).

#### B. Leakage Inductance

When the proposed converter is operated under step-down conversion, diode reverse recovery will cause high-current spike during switching transition. The leakage inductance of the transformer can not only limit diode-reverse-recovery current but also can achieve ZVS turn-on with a phase-shift operation manner. The leakage inductance needs to satisfy

\[
i_S > \sqrt{\frac{2}{L_{eqh}}} \left( \frac{4}{3} C_{\text{MOS}} V^2_{\text{in}} + \frac{1}{2} C_{\text{TR}} V^2_{\text{in}} \right)
\]  

(2)

where \(L_{eqh} = L_{zh} + L_{sll} \cdot N^2_s / N^2_s\). However, large leakage inductance will cause high current difference during the step-up conversion, and the flyback snubber needs to process a high power level.

#### C. Flyback Snubber

During the interval of \([t_0, t_2]\), a high transient voltage occurs under the step-up conversion due to the current difference between the leakage inductance and current-fed inductor currents, which can be suppressed by \(D_C\) and \(C_C\). The energy stored in capacitor \(C_C\) is transferred to buffer capacitors \(C_{b1}\) and \(C_{b2}\) by the flyback snubber. The power rating of the flyback snubber can be expressed as

\[
P_{FB} = 0.5 C_C \cdot V^2_C \cdot f_S
\]  

(3)

where \(f_S\) is the switching frequency.

#### D. Magnetizing Inductance of Flyback Snubber

To ensure that snubber capacitor \(C_C\) can be fully discharged by the flyback snubber, the turn-on time of \(M_S\) \((T_{ON})\) cannot be longer than \(1/4\) resonant cycle of the magnetizing inductance and \(C_C\). Thus, the magnetizing inductance of the flyback snubber should satisfy the following inequality:

\[
L_{m} = \frac{4T^2_{ON}}{\pi^2 C_C}
\]  

(4)

where \(T_{ON}\) is the conduction time of switches \(M_1 \sim M_4\) in the step-up conversion.

#### E. Buffer Capacitors \(C_{b1}\) and \(C_{b2}\)

When the converter is operated in the step-down conversion, capacitors \(C_{b1}\) and \(C_{b2}\) can share current \(i_S\), and \(V_{ds(M6)}\) and \(V_{ds(M8)}\) will rise up with a lower slope at switches \(M_6\) and \(M_8\) turnoff transition, reducing switching loss. If the power rating becomes higher, buffer capacitors \(C_{b1}\) and \(C_{b2}\) can be enlarged to achieve near ZCS turn-off. On the other hand, when the converter is operated in the step-up conversion, snubber capacitor \(C_C\) will be fully discharged by the proposed operation scheme. When switches \(M_1 \sim M_4\) are switched to either \((M_1, M_4)\) or \((M_2, M_3)\) conducting, current difference \(i_C = i_L - i_p\) will first charge capacitor \(C_C\). Meanwhile, current \(i_p\) can start to rise up because of \(C_{b1}\) and \(C_{b2}\) holding voltage \(V_{HV}\), which can reduce duty loss. For \(C_{b1}\) and \(C_{b2}\) being charged up to \(V_{HV}\) by the flyback snubber, they need to satisfy the following inequality:

\[
C_{b1}, C_{b2} \leq \frac{C_C \cdot V^2_C}{V^2_{HV}}
\]  

(5)

#### F. Snubber Diodes \(D_{b1}\) and \(D_{b2}\)

In the analysis, if capacitors \(C_{b1}\) and \(C_{b2}\) are connected in parallel with the upper legs of the voltage-fed bridge directly, they will resonate with the leakage inductance during switching transition in the step-down conversion, increasing switching loss. Thus, two snubber diodes \(D_{b1}\) and \(D_{b2}\) are introduced to connect with \(C_{b1}\) and \(C_{b2}\) in series, respectively, and the ringing current through the high-side switches can be blocked effectively.

#### G. Soft Start-Up

High inrush current is a start-up problem with the step-up conversion. The initial high-side voltage \(V_{HV}\) should not be lower than \(V_{HV} \cdot N_S / N_P\) to avoid inrush current. When \(V_{HV}\) is not high enough, the controller will drive the flyback snubber to precharge high-side capacitor \(C_{HV}\) to achieve a soft start-up feature.

### IV. EXPERIMENTAL RESULTS

To verify the operational principle and performance of the proposed converter, three experimental prototypes of 1.5 kW, the converter shown in Fig. 2 (type A), the one shown in Fig. 3 (type B), and the proposed, shown in Fig. 4 (type C), were designed and built. The low-side voltage is 42–54 V, and the...
TABLE I
SPECIFICATIONS OF PROTOTYPE

<table>
<thead>
<tr>
<th>Specification</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Low-side Voltage</td>
<td>48 V (nominal value)</td>
</tr>
<tr>
<td>High-side Voltage</td>
<td>360 V</td>
</tr>
<tr>
<td>Maximum Output Power</td>
<td>1.5 kW</td>
</tr>
<tr>
<td>Switching Frequency</td>
<td>25 kHz</td>
</tr>
<tr>
<td>Low-side Switches</td>
<td>M1 ~ M4: IRFB4321Pbf</td>
</tr>
<tr>
<td>High-side Switches</td>
<td>M5 ~ M8: IRFP26N60LBF</td>
</tr>
<tr>
<td>Turns Ratio</td>
<td>Np / Np = 1 / 4.26</td>
</tr>
<tr>
<td>Leakage Inductance</td>
<td>Ld1 = 0.5 μH, Ld2 = 9 μH</td>
</tr>
<tr>
<td>Current-fed Inductor</td>
<td>Lm = 500 μH</td>
</tr>
<tr>
<td>High-side capacitor</td>
<td>470 μF * 2</td>
</tr>
<tr>
<td>Core of Isolation Transformer</td>
<td>EE-55</td>
</tr>
<tr>
<td>Core of Flyback Transformer</td>
<td>EI-40</td>
</tr>
<tr>
<td>Snubber Capacitor</td>
<td>Cc: 100 nF</td>
</tr>
<tr>
<td>Buffer Capacitor</td>
<td>Cb1 &amp; Cb2: 4.7 nF</td>
</tr>
<tr>
<td>Turns Ratio (Flyback)</td>
<td>Nf / Nf / Nf = 1 / 3 / 3</td>
</tr>
<tr>
<td>Leakage Inductance (Flyback)</td>
<td>1.3 μH</td>
</tr>
</tbody>
</table>

Measured waveforms of low-side voltage $V_{LV}$ and current $i_P$ under step-up conversion are shown in Fig. 10. For the voltage variation of batteries, the proposed bidirectional converter can cover input voltage range from 42 to 54 V.

Measured voltage waveforms of snubber capacitor $V_C$ and $V_{ds(M4)}$ from type A and the proposed one under step-up conversion are shown in Fig. 11. It can be observed from Fig. 11(a) that $V_C$ is regulated to a desired value just slightly higher than $V_{ds(M4)}$ in type A. Since the regulated voltage $V_C$ is slightly higher than $V_{ds(M4)}$, the energy transferred by the flyback snubber is just a small amount. However, the parasitic capacitance of the switches and the stray inductance of the circuit will result in high-voltage spike. In the proposed converter, capacitor $C_C$ is fully discharged, which could result in higher energy transferred by the flyback snubber than type A. Thus, capacitor $C_C$ should be chosen with a smaller capacitance. For type A, the value of $C_C$ is 1 μF, while that of the proposed one is 0.1 μF. Referring to (2), the processed power $P_{FB}$ by the flyback snubber of the proposed one is 36 W, just around 2.4% of the maximum power level. Moreover, the proposed operating scheme can achieve near soft-switching feature for switches $M_1 \sim M_4$ at turnoff transition, and the overshoot voltage can be suppressed to a lower level, as shown in Fig. 11(b). Fig. 12 shows the measured waveform of current $i_P$ from type A and the proposed one. Although capacitor $C_C$ is fully discharged, the proposed passive snubber can hold voltage $V_{b1}$ or $V_{b2}$.
Referring to the following equations, $\frac{dp}{dt}$ of the proposed one is higher than that of type A, which can reduce duty loss

$$\frac{dp}{dt} = \frac{V_{AB} - V_{CD} \times \frac{N_p}{N_S}}{L_{eq}}$$

$$= \frac{V_{CC} - (V_{HV} \times \frac{N_p}{N_S})}{L_{eq}} \quad \text{(Type A)}$$

$$\frac{dp}{dt} = \frac{V_{AB} - V_{CD} \times \frac{N_p}{N_S}}{L_{eq}}$$

$$= \frac{V_{CC} - (V_{HV} - V_{C61}) \times \frac{N_p}{N_S}}{L_{eq}} \quad \text{(proposed)}.$$

Fig. 13 shows the measured waveforms of voltage $V_{ds(M4)}$ and current $I_{ds(M4)}$ from type A and the proposed one at $M_4$ turnoff transition, in which Fig. 13(a) and (b) are with a load of 1.5 kW and Fig. 13(c) is with 500 W. It can be observed that the voltage spike in type A is up to 197 V, due to large buffer capacitors $C_{b1}$ and $C_{b2}$, voltage $V_{ds}$ of type B is ringing at switching transition, as shown in Fig. 13(b). Moreover, Fig. 13(b) and (c) shows that the soft-switching feature can be achieved at both light- and heavy-load conditions. Fig. 14 shows the measured waveforms of voltage $V_{ds(M8)}$ and current $I_{ds(M8)}$ from type A and the proposed one at $M_8$ turnoff transition. It can be observed that, with $C_{b1}$ and $C_{b2}$, $V_{ds(M8)}$ of the proposed converter rises up with a lower slope and the switching loss of $(V_{ds(M8)} \cdot I_{ds(M8)})$ becomes lower, achieving near ZCS turnoff soft-switching feature.

Fig. 15 shows the measured waveforms of voltages $V_{ds(M5)}$ and $V_{ds(M6)}$ from type B and the proposed one under step-down conversion. It can be observed from Fig. 16(a) that, because of large buffer capacitors $C_{b1}$ and $C_{b2}$, there exists high ringing current in type B at switching transition, which will result in high EMI noise and switching loss. On the other hand, the ringing current is much lower in the proposed converter, as shown in Fig. 16(b).

Fig. 17 shows the measured waveform of current $i_p$ under step-up conversion from the converter with an active clamp circuit [13] and the proposed one. It can be observed from Fig. 17(a) that the peak current of $i_p$ under the full-load condition is 48.1 A, which means the discharging current $(48.1 - 30 = 18.1 A)$ of $C_C$ flowing through the main switches and increasing stress significantly. On the contrary, the energy stored in $C_C$ is recycled by the proposed passive and active snubbers, which can reduce the current stress.

Fig. 18 shows the plots of conversion efficiency versus power level of type A and the proposed one operated in the step-up conversion. The processed power $P_{FB}$ by the flyback snubber in the proposed converter is 36 W, while $P_{FB}$ is 52.5 W in type A. The conversion efficiency of the proposed converter under the full-load condition is about 91.5%. However, since capacitor
Fig. 17. Measured current $i_p$ waveforms from (a) the converter with an active clamp circuit [13] and (b) the proposed one under step-up conversion and with 1.5-kW power rating.

Fig. 18. Plots of conversion efficiency of type A and the proposed one operated in the step-up conversion.

Fig. 19. Plots of conversion efficiency of type A and the proposed one operated in the step-down conversion.

<table>
<thead>
<tr>
<th>Key Components</th>
<th>Loss Estimation</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>Step-up Conversion</td>
</tr>
<tr>
<td>Low-side Switches $M_1$ - $M_4$</td>
<td>21.6 W</td>
</tr>
<tr>
<td>High-side Switches $M_5$ - $M_8$</td>
<td>4.0 W</td>
</tr>
<tr>
<td>Current-fed Inductor $L_{sw}$</td>
<td>51.9 W</td>
</tr>
<tr>
<td>Isolation Transformer $T_p$</td>
<td>26.8 W</td>
</tr>
<tr>
<td>Flyback Snubber</td>
<td>10.8 W</td>
</tr>
<tr>
<td>Snubber Diodes $D_{s1}$ &amp; $D_{s2}$</td>
<td>0.2 W</td>
</tr>
<tr>
<td>Total</td>
<td>115.3 W</td>
</tr>
</tbody>
</table>

TABLE II

KEY COMPONENT LOSS ESTIMATION UNDER 1.5-kW POWER RATING

This paper has presented a soft-switching bidirectional isolated full-bridge converter, which allows input voltage variation from 42 to 54 V, for battery charging/discharging applications. The proposed converter can reduce the voltage spike caused by the current difference between leakage inductance and current-fed inductor currents, the current spike due to diode reverse recovery, and the current and voltage stresses, while it can achieve near ZVS and ZCS soft-switching features. The passive snubber can hold voltage $V_{d1}$ or $V_{d2}$ and improve the slew rate of $dt/dt$, which can reduce duty loss. However, near ZVS turn-on transition cannot be achieved under light-load condition in step-down conversion. Experimental results measured from the three types of 1.5-kW isolated bidirectional full-bridge dc–dc converters have verified that the proposed converter (type C) can yield the performance of lower voltage and current spikes, higher efficiency, and less ringing. It is suitable for high-power applications with galvanic isolation.

REFERENCES


Tsai-Fu Wu (S’88–M’91–SM’98) received the B.S. degree in electronic engineering from National Chiao Tung University, Hsinchu, Taiwan, in 1983, the M.S. degree in electrical and computer engineering from Ohio University, Athens, OH, USA, in 1988, and the Ph.D. degree in electrical engineering and computer science from the University of Illinois, Chicago, IL, USA, in 1992.

From 1985 to 1986, he was a System Engineer with SAMPO, Inc., Kuei-Shan Hsiang, Taiwan, where he was involved in developing and designing graphic terminals. From 1988 to 1992, he was a Teaching and Research Assistant with the Department of Electrical Engineering and Computer Science, University of Illinois, Chicago, IL, USA. From 1993 to 2012, he was with the Department of Electrical Engineering, National Chung Cheng University, Minxiong, Taiwan. He is currently a Professor with the Department of Electrical Engineering, National Tsing Hua University, Hsinchu. His current research interests include developing and modeling of power converters; design of electronic dimming ballasts for fluorescent lamps, metal halide lamps and plasma display panels; design of solar-array supplied inverters for grid connection; and design of pulsed-electrical-field generators for transdermal drug delivery and food pasteurization.

Dr. Wu is a Senior Member of the Chinese Institute of Engineers. He was the recipient of three Best Paper Awards from the Taipei Power Electronics Association in 2003–2005. In 2006, he was named as an Outstanding Researcher by the National Science Council, Taiwan.

Jeng-Gung Yang was born in Taiwan in 1985. He received the B.S. and M.S. degrees in electrical engineering from National Chung Cheng University, Minxiong, Taiwan, in 2009 and 2011, respectively, where he is currently working toward the Ph.D. degree in the Department of Electrical Engineering.

His current research interests include the design and development of soft-switching power converters.

Chia-Ling Kuo was born in Taiwan in 1985. She received the B.S. and M.S. degrees in electrical engineering from National Chung Cheng University, Minxiong, Taiwan, in 2008 and 2010, respectively, where she is currently working toward the Ph.D. degree in the Department of Electrical Engineering.

Her research interests include the design and implementation of bidirectional inverters for dc-distribution applications.

Yung-Chun Wu was born in Kaohsiung, Taiwan, in 1971. He received the Ph.D. degree in electrical engineering from National Chung Cheng University, Minxiong, Taiwan, in 2000.

From 2000 to 2001, he was an Assistant Professor with the Department of Electrical Engineering, Chienkuo Institute of Technology, Changhua, Taiwan. From 2001 to 2007, he was an Assistant Professor with the Department of Electrical Engineering, I-Shou University, Kaohsiung, Taiwan. Since 2007, he has been with the Department of Aeronautical Engineering, National Formosa University, Huwei, Taiwan, where he is currently an Assistant Professor. His research interests include developing and designing converter topologies, electronic ballasts, and high-efficiency converters.