A Fully Soft-Switched Single Switch Isolated DC–DC Converter

Minjae Kim, Student Member, IEEE, and Sewan Choi, Senior Member, IEEE

Abstract—This paper proposes a soft-switched single switch isolated converter. The proposed converter is able to offer low cost and high power density in step-up application due to the following features: zero-current switching (ZCS) turn-on and zero-voltage switching (ZVS) turn-off of switch and ZCS turn-off of diodes regardless of voltage and load variation; low rated lossless snubber; reduced transformer volume compared to flyback-based converters due to low magnetizing current. Experimental results on a 100 kHz, 250 W prototype are provided to validate the proposed concept.

Index Terms—Isolated step-up dc–dc converter, single switch, soft switching.

I. INTRODUCTION

ISOLATED step-up dc–dc converters are used in many applications, such as photovoltaic module-integrated converter (MIC) systems, portable fuel cell systems, and vehicle inverters where high efficiency, high power density, and low cost are required [1]–[4]. Owing to smaller input current ripple, lower diode voltage rating and lower transformer turns ratio, the current-fed isolated converter is better suited for step-up applications. The current-fed isolated converter has two types: passive-clamped [5]–[7] and active-clamped [8]–[13]. The passive-clamped current-fed converter has simple structure and small switch count, but suffers from excessive power losses dissipated in the RCD snubber and associated with hard switching of main switch: Active-clamped current-fed converters have actively been developed based on three basic topologies: push–pull [8], full-bridge [9], [10], and half-bridge [11]–[13]. They achieve not only lossless clamping of voltage spikes caused by transformer leakage inductance but also zero-voltage switching (ZVS) turn on of switches. However, they may not be expected to achieve high efficiency and low cost in relatively low power application since they need at least four switches and gate driver circuits.

Isolated converters with reduced switch count have been proposed for low power application [14]–[26]. Isolated dc–dc converters with one main switch and one clamp switch achieve zero turn on of switches, but switches are turned off with hard switching [14]–[17]. Isolated single switch dc–dc converters are more attractive to achieve low cost [18]–[26]. Z-source converter [18], [19] and flyback converter [20]–[23] are hard switched at both turn-on and turn-off instants. Frequency-controlled flyback converter [24] and series-connected forward-flyback converter [25] achieve zero-current switching (ZCS) turn-on of switch, but the switch is hard switched at turn-off instant. The aforementioned single switch topologies have increased transformer volume since magnetizing inductor is used for energy transfer. An isolated single-switch resonant converter [26] achieves both ZCS turn-on and ZCS turn-off of switch, but need high transformer turn ratio for step-up application due to low voltage gain and hence is not suited to step-up application.

In this paper, a soft-switched single switch isolated converter is proposed for step-up application. The proposed converter has the following features: 1) ZCS turn-on and ZVS turn-off of switch regardless of voltage and load variation; 2) ZCS turn-off of all diodes leading to negligible voltage surge associated with the diode reverse recovery; 3) small input current ripple due to CCM operation; 4) reduced transformer volume due to low magnetizing current; and 5) low-rated lossless snubber, which makes it possible to achieve high efficiency and low cost for step-up application. Experimental results on a 100 kHz, 250 W prototype are provided to validate the proposed concept.

II. PROPOSED CONVERTER

Fig. 1 shows the circuit diagram of the proposed converter. The proposed converter consists of input filter inductor \(L_i\), switch \(S_i\), a lossless snubber which includes capacitor \(C_s\), inductor \(L_s\), and diodes \(D_{s1}\) and \(D_{s2}\), and clamp capacitor \(C_c\) at the primary side and \(L_c–C_c\) series resonant circuit and diodes \(D_1\) and \(D_2\) at the secondary side. The lossless snubber makes it possible to achieve ZVS turn-off of switch as well as clamp the voltage spikes of the switch by leakage inductance. Also, the \(L_c–C_c\) series resonant circuit makes it possible to achieve ZCS turn-off of diodes. Fig. 2 shows three resonance operations according to the variations of resonant frequency \(f_r\), which is expressed as in (1): the above-resonance operation \((DT_s < 0.5f_r)\), the resonance operation \((DT_s = 0.5f_r)\), and

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input voltage $V_i$. In the below-resonance operation, nine modes exist within $T_s$.

Mode 1 ($t_0$–$t_1$): This mode begins when switch $S_1$ is turned on. Equivalent circuit of this mode is shown in Fig. 5(a). $L_s$ and $C_s$ start resonating and resonant current $i_{L_s}$ flows through $L_s$, $D_{s1}$, $C_s$, and $S_1$. The voltage and current of resonant components are determined, respectively, as follows:

$$i_{L_s}(t) = \frac{C_s}{L_s} \sin(\omega_{r2}(t - t_0)), \quad t_0 < t < t_1$$ \hspace{1cm} (2)
$$v_{C_s}(t) = \frac{C_s}{L_s} \cos(\omega_{r2}(t - t_0)), \quad t_0 < t < t_1$$ \hspace{1cm} (3)

where $\omega_{r2} = 1/\sqrt{L_s C_s}$. Since induced voltage $V_{C_{r, min}} - nV_{C_r}$ across $L_r$ makes time interval from $t_0$ to $t_1$ very short, current $i_{r}$ appears to decrease almost linearly. Current through $S_1$ increases with the slope of $i_{L_s}$, resulting in ZCS turn-on of $S_1$. The turn-on loss of switch associated with energy stored in MOSFET’s output capacitance is negligible in this low input voltage application [27]–[29]. This mode ends when current $i_{L_r}$ reaches 0 A. It is noted that diode $D_1$ is turned OFF under ZCS condition.

Mode 2 ($t_1$–$t_2$): This mode begins when current $i_{L_r}$ changes its direction. Equivalent circuit of this mode is shown in Fig. 5(b). $L_r$ and $C_r$ start resonating and resonant current $i_{L_r}$ flows through $L_r$, $C_r$, and $D_2$. The voltage and current of resonant components are determined, respectively, as follows:

$$i_{L_r}(t) = (V_{C_{r, min}} - nV_{C_r}) \sqrt{\frac{C_r}{L_r}} \sin(\omega_{r1}(t - t_1)), \quad t_1 < t < t_4$$ \hspace{1cm} (4)
$$v_{C_r}(t) = nV_{C_r} - (nV_{C_r} - V_{C_{r, min}}) \cos(\omega_{r1}(t - t_1)), \quad t_1 < t < t_4$$ \hspace{1cm} (5)

where $\omega_{r1} = 1/\sqrt{L_r C_r}$. When voltage across snubber capacitor $C_s$ equals $-V_{C_r}$, $L_s - C_s$ resonance ends.

Mode 3 ($t_2$–$t_3$): This mode begins when diode $D_{s2}$ is turned on. Current $i_{L_s}$ is determined by following equation, and this mode ends when current $i_{L_s}$ reaches 0 A.

$$i_{L_s}(t) = -\frac{V_{C_r}}{L_s} (t - t_2) + i_{L_s}(t_2), \quad t_2 < t < t_3.$$ \hspace{1cm} (6)

It is noted that diodes $D_{s1}$ and $D_{s2}$ are turned OFF under ZCS condition.

Mode 4 ($t_3$–$t_4$): The $L_r - C_r$ resonance keeps on during this mode and ends when current $i_{L_r}$ reaches 0 A. Note that diode $D_2$ is turned OFF under ZCS condition.
Mode 5 ($t_5$–$t_6$): During this mode, a constant current flows through $S_1$ whose value is the sum of the input current $I_i$ and the magnetizing current $I_{Lm}$.

Mode 6 ($t_6$–$t_7$): The mode begins when $S_1$ is turned OFF. Then, $I_i + I_{Lm}$ flows through $C_s$, $D_{s2}$, and $C_c$. Voltage across snubber capacitor $C_s$ which is determined by the following equation increases linearly with the slope of $(I_i + I_{Lm})/C_s$, resulting in ZVS turn-off of $S_1$.

$$v_{C_s}(t) = \frac{I_i + I_{Lm}}{C_s}(t - t_5) - V_{Cc,t_5}, \quad t_5 < t < t_6. \quad (7)$$

This mode ends when $v_{C_s}$ becomes equal to $(V_o - V_{Cr,max})/n$.

Mode 7 ($t_7$–$t_8$): This mode begins when diode $D_1$ is turned ON. Equivalent circuit of this mode is shown in Fig. 5(c). $L_r$ and $C_s$ start resonating and resonant current $i_{Lr}$ flows through $C_s$, $D_{s2}$, $L_r$, $D_1$, and $C_r$. Assuming that $C_s << n^2 C_r$, $v_{Cr}$ can be considered constant, and resonance frequency $\omega_{r3}$ can be determined by $C_s$ and $L_r$. Therefore, the voltage and current of resonant components are determined, respectively, as follows:

$$i_{Lr}(t) = \left(\frac{I_i + I_{Lm}}{n}\right)\left[1 - \cos(\omega_{r3}(t - t_6))\right], \quad t_6 < t < t_7 \quad (8)$$

$$v_{C_r}(t) = \frac{I_i + I_{Lm}}{n} \sqrt{\frac{L_r}{C_s}} \sin(\omega_{r3}(t - t_6)) + \frac{V_o - V_{Cr,max}}{n}, \quad t_6 < t < t_7 \quad (9)$$

where $\omega_{r3} = n/\sqrt{L_r C_s}$. This mode ends when current $i_{Lr}$ becomes equal to $(I_i + I_{Lm})/n$.

Mode 8 ($t_8$–$t_9$): This mode begins when diode $D_{s1}$ is turned ON. Equivalent circuit of this mode is shown in Fig. 5(d). $L_s$, $C_s$, $L_r$, and $C_r$ start resonating and resonant current $i_{Lr}$ flows through $L_s$, $D_{s1}$, $C_s$, $C_r$, $L_r$, $D_1$, and $C_r$. Assuming that $C_s << n^2 C_r$ and $L_s \gg L_r/n^2$, the voltage and current of resonant components are determined using the superposition principle, respectively, as follows:

$$i_{L_s}(t) = \left[\frac{V_{Cc}}{n} + \frac{V_o}{n} - \left(\frac{V_{C,s,max}}{n} + \frac{V_{Cr,max}}{n}\right)\right]$$

$$\times \sqrt{\frac{C_s}{L_s}} \sin(\omega_{r3}(t - t_5)), \quad t_7 < t < t_8 \quad (10)$$
\[ v_{C_r}(t) = \left[ V_{C_i} + \frac{V_o}{n} - \left( \frac{V_{C_r,\text{max}} + V_{C_r,\text{max}}}{n} \right) \right] \times [1 - \cos(\omega_r (t - t_r))] + V_{C_{r,\text{max}}}, \quad t_7 < t < t_8. \]  
\[ \text{(11)} \]

Assuming that \( i_{L_r} \approx (I_i + I_{L_m})/n \) during this mode, the voltage \( v_{C_r} \) is determined by the following equation:
\[ v_{C_r}(t) = -\frac{I_i + I_{L_m}}{nC_r}(t - t_7) - V_{C_r,\text{max}}, \quad t_7 < t < t_9. \]  
\[ \text{(12)} \]

This mode ends when current \( i_{L_s} \) reaches 0 A.

Mode 9 (\( t_9 - t_0 \)): Switch \( S_1 \) is in the turn-off state, and the sum of the input current and magnetizing current is being transferred to the secondary. Current \( i_{D1} \) is equal to \( (I_i + I_{L_m})/n \). This mode ends when switch \( S_1 \) is turned on.

The average current of magnetizing inductor \( L_m \) is equal to the average current of snubber inductor \( L_s \), since \( I_{L_s,\text{avg}} = I_{D2,\text{avg}} \) and \( I_{D2,\text{avg}} = I_{L_m,\text{avg}} \). Therefore, it should be noted that transformer core volume of the proposed converter is much smaller compared to that of the flyback-based converter since \( I_{L_s,\text{avg}} = I_{L_m,\text{avg}} \) can be designed to be small.

### B. Voltage Gain Expression

To obtain voltage gain of the proposed converter, it is assumed that voltage across \( C_r \) is constant and magnetizing current is ignored during the switching period \( T_s \).

1) Below-Resonance Operation (\( DT_s > 0.5T_r \)): Since the average current of diode \( D_2 \) is identical to the average load current in the steady state, the following equation is obtained:
\[ I_{D2,\text{avg}} = \frac{V_o}{R_o} = \frac{1}{T_s} \int_{t_1}^{t_4} i_{D2}(t)\,dt \]
\[ = \frac{1}{T_s} \int_{t_1}^{t_4} -i_{L_r}(t)\,dt. \]  
\[ \text{(13)} \]

From (4) and (13), the minimum voltage of the resonant capacitor \( V_{C_r,\text{min}} \) can be obtained by
\[ V_{C_r,\text{min}} = nV_{C_i} - \frac{V_o}{2C_r f_s R_o}. \]  
\[ \text{(14)} \]

From (5) and (14), the maximum voltage of the resonant capacitor \( V_{C_r,\text{max}} \) can be obtained by
\[ V_{C_r,\text{max}} = nV_{C_i} + \frac{V_o}{2C_r f_s R_o}. \]  
\[ \text{(15)} \]

The time interval from \( t_7 \) to \( t_9 \) in Fig. 3 can be obtained from (12) and (15) by
\[ t_9 - t_7 = \frac{nV_o}{I_i f_s R_o}. \]  
\[ \text{(16)} \]

The time interval from \( t_5 \) to \( t_6 \) in Fig. 3 can be obtained from (7) by
\[ t_6 - t_5 = \frac{C_s}{I_i} \left( \frac{V_o - V_{C_r,\text{max}}}{n} + V_{C_i} \right). \]  
\[ \text{(17)} \]

The time interval from \( t_6 \) to \( t_7 \) in Fig. 3 is quarter the \( L_r - C_s \) resonant frequency and determined by
\[ t_7 - t_6 = \frac{\pi}{2\omega_{r,3}}. \]  
\[ \text{(18)} \]

By applying (16)–(18) to (1–D)\( T_s \), the voltage gain can be obtained by
\[ \frac{V_o}{V_i} = \frac{n + B}{1 - D - A}, \]  
\[ \text{(19)} \]

where \( A = \frac{\pi L_r}{2\omega_{r,3}} \) and \( B = \frac{C_r/(2C_i f_s R_o - I_i)}{2nC_r} \).

2) Above-Resonance Operation (\( DT_s < 0.5T_r \)): Fig. 6 shows key waveforms of the proposed converter in the above-resonance operation. The operating principles of the above resonance are the same as that of the below resonance except time interval from \( t_3 \) to \( t_4 \). Assuming that \( C_s << n^2 C_r \), an equivalent circuit of time interval from \( t_3 \) to \( t_4 \) is shown in Fig. 7(b). The time interval from \( t_7 \) to \( t_6 \) can be approximated as \( DT_s \). Since the average current of diode \( D_2 \) is identical to the average load current, it can be approximated by
\[ I_{D2,\text{avg}} = \frac{V_o}{R_o} \approx \frac{1}{T_s} \int_0^{DT_s} i_{D2}(t)\,dt \]
\[ \approx \frac{1}{T_s} \int_0^{DT_s} -i_{L_r}(t)\,dt. \]  
\[ \text{(20)} \]
From (4) and (20), minimum voltage of the resonant capacitor $V_{Cr,\text{min}}$ can be obtained by

$$V_{Cr,\text{min}} = nV_Cc - \frac{V_o}{C_r f_s R_o \left(1 - \cos(\omega_1 DT_s)\right)}.$$  \hfill (21)

From (5) and (21), maximum voltage of the resonant capacitor $V_{Cr,\text{max}}$ can be obtained by

$$V_{Cr,\text{max}} = nV_Cc - \frac{V_o \cos(\omega_1 DT_s)}{C_r f_s R_o \left(1 - \cos(\omega_1 DT_s)\right)}.$$  \hfill (22)

The time interval from $t_0$ to $t_1$ in Fig. 6 can be obtained from (12) and (22) by

$$t_1 - t_0 = \frac{nV_o}{I_s f_s R_o}.$$  \hfill (23)

Assuming that $dV_{Cc}/dt \approx I_s/C_c$ during the time interval from $t_0$ to $t_3$ in Fig. 6, this time interval can be obtained from (7) by

$$t_3 - t_0 = \frac{C_c}{I_s} \left(\frac{V_o}{n} + V_{Cc}\right).$$  \hfill (24)

The time interval from $t_3$ to $t_6$ in Fig. 6 is quarter the $L_s, C_s$ resonant frequency and determined by

$$t_6 - t_3 = \frac{\pi}{2\omega_{g1}}.$$  \hfill (25)

By applying (23)–(25) to (1–$D$) $I_s$, the voltage gain can be obtained by

$$\frac{V_o}{V_i} = \frac{n + C}{1 - D - A}$$  \hfill (26)

where $C = C_c \frac{[C_r f_s R_o (1 - \cos(\omega_1 DT_s)))]}{[C_r f_s R_o (1 - \cos(\omega_1 DT_s))]}$.

Using (19) and (26), the voltage gain of the proposed converter is plotted as shown in Fig. 8.

C. Design Procedure

In this section, a design procedure of the proposed converter is presented with an example. A specification for the design example is given as follows: output power $P_o = 250$ W, output voltage $V_o = 380$, input voltage $V_i = 28$–38 V, and switching frequency $f_s = 100$ kHz.

1) Choose Average Value of Snubber Inductor Current

$I_{L_s,\text{avg}}$: $I_{L_s,\text{avg}}$ should be as small as possible to in order to minimize conduction loss of the snubber components and magnitude of the magnetizing current. It is seen from (2) and (10) that $I_{L_s,\text{avg}}$ is proportional to the snubber capacitance $C_s$. However, if $C_s$ is chosen to be small to reduce conduction loss of the snubber components, the voltage rating of the switch increases, as shown in (9), resulting in high conduction loss of the switch. Therefore, considering tradeoff between conduction losses of the switch and snubber components, $I_{L_s,\text{avg}}$ is chosen to be around 3% of average input current, which is expressed as

$$I_{L_s,\text{avg}} = 0.03 I_s, \text{avg} = 0.27 \text{ A.}$$  \hfill (27)

2) Determine Values of $n$, $L_s$, and $C_s$:

In order to simplify the design procedure, the voltage gain can be approximated as

$$\frac{V_o}{V_i} \approx \frac{n}{1 - D}.$$  \hfill (28)

As mentioned earlier, the below-resonance operation is chosen for the proposed converter due to smaller switch turn-off current and diode $d_i d_i$. From Fig. 2, the minimum duty cycle for the below-resonance operation can be obtained by

$$D_{\text{min}} = \frac{\pi f_s \sqrt{L_r C_r}}{2},$$  \hfill (29)

Since resonant inductance $L_r$ should be designed to minimize reverse-recovery effects of diode $D_1$, the time interval from $t_0$ to $t_1$ in Fig. 3 should be greater than $3\tau_{rr1}$, which is expressed as

$$t_1 - t_0 = 3\tau_{rr1} = \frac{(I_s + I_{Lm}) L_r}{n V_o (1 + 1/2C_r f_s R_o)}$$  \hfill (30)

where $\tau_{rr1}$ is reverse-recovery time of diode $D_1$.

Based on the operating principles, the RMS current and turn-on voltage of switch $S_1$ can be obtained, respectively, as follows:

$$I_{S1,\text{rms}} \approx \sqrt{2} I_s + \frac{n\pi I_s}{2\sqrt{2} D_{\text{min}}}, \quad (D_{\text{min}} < D < D_{\text{max}})$$  \hfill (31)

$$V_{S1,\text{on}} = V_{S1}(t_0) = \frac{V_o - V_{Cr,\text{min}}}{n} + V_{Cc}.$$  \hfill (32)

Fig. 9 shows RMS current and turn-on voltage of switch $S_1$ based on (28)–(32) with different values of $n$. In this example, the turn ratio of the transformer $n$ is chosen to be 5 considering tradeoff of conduction loss and switching loss of switch $S_1$. Resonant values $L_r$ and $C_r$ are determined by 5 $\mu$H and 560 nF, respectively, using (28)–(30).

3) Determine Value of $C_s$:

$I_{L_s,\text{avg}}$ can be obtained by, using (2), (6), and (10)

$$I_{L_s,\text{avg}} = \frac{C_r}{T_s} \left[3V_{Cc}(t_0) + 3V_{Cc} - 2V_{Cs,\text{max}} + \frac{2(V_o - V_{Cr,\text{max}})}{n} + 0.5 v_{C_s}(t_0) \sin \left(\cos^{-1} \left(\frac{-V_{Cc}}{v_{C_s}(t_0)}\right) \cos^{-1} \left(\frac{-V_{Cc}}{v_{C_s}(t_0)}\right)\right)\right]$$  \hfill (33)
where
\[ v_{Cs}(t_0) = \frac{2(V_C + V_o - V_{Cr,max})}{n} - V_{Cs,max} \]
and
\[ V_{Cs,max} = \frac{L_i + L_m}{n} \sqrt{\frac{V_o}{n} + V_o - V_{Cr,max}}. \]

By applying \( n, L_r, C_r, \) and (27) to (33), snubber capacitance \( C_s \) can be calculated as 16 nF.

4) Determine Value of \( L_s \): Snubber inductance \( L_s \) should be designed to minimize reverse-recovery effects of snubber diodes \( D_{s1} \) and \( D_{s2} \). Therefore, the time interval from \( t_2 \) to \( t_3 \) in Fig. 3 should be greater than \( 3t_{rr2} \), which is expressed as
\[ t_3 - t_2 = 3t_{rr2} = \frac{v_{Cs}(t_0)L_s \sin^{-1}\left(-\frac{V_i}{v_{Cs}(t_0)}\right)}{V_i} \sqrt{\frac{C_s}{L_s}}, \tag{34} \]
where \( t_{rr2} \) is reverse-recovery time of diodes \( D_{s1} \) and \( D_{s2} \).

According to (34), snubber inductance \( L_s \) is calculated as 5 \( \mu \)H.

5) Select Semiconductor Devices: Semiconductor devices of the proposed converter are selected based on the previous design procedure and operating principles. It can be seen from Fig. 3 that output diodes \( D_1 \) and \( D_2 \) have a maximum voltage stress of \( V_o \). Peak current of output diode \( D_2 \) is \( 0.5i_p/D_{min} \) from (13). Maximum voltage stress across the switch \( S_1 \) is determined by
\[ V_{S1,max}(t) = \frac{I_s + I_{Lm}}{n} \sqrt{\frac{L_i}{C_s}} + \frac{V_o - V_{Cr,max}}{n} + V_C. \tag{35} \]

Current stress of switch \( S_1 \) is determined by (31). As shown in Figs. 3 and 4, maximum voltage stresses across snubber diodes \( D_{s1} \) and \( D_{s2} \) are \( v_{Cs}(t_0) + V_C \) and \( V_{Cr} \), respectively. The peak currents of snubber diodes \( D_{s1} \) and \( D_{s2} \) are \( v_{Cs}(t_0)\sqrt{C_s/L_s} \) from (2) and \( L_i \), respectively. Selected devices and component ratings according to the previous design procedure are shown in Table I.

### III. EXPERIMENTAL RESULT

A 250 W laboratory prototype of the proposed converter has been built and tested to verify the proposed concept. Component ratings and selected devices of the proposed converter are listed in Table I. We can see from Table I that current ratings of the snubber components are much lower than those of main components. Leakage inductance of the transformer is used as the resonant inductance. Figs. 10 and 11 show experimental waveforms at full-load and half-load conditions when input voltage is 28 V, respectively. Figs. 10(a) and 11(a) show that switch \( S_1 \) is turned on with ZCS at both full- and half-load conditions. Figs. 10(b) and 11(b) show the experimental waveforms of switch \( S_1 \) at turn-off. In theory, \( S_1 \) is turned off with ZVS. However, it is seen from Figs. 10(b) and 11(b) that \( S_1 \) generates small amounts of losses due to ringing caused by parasitic resonance between MOSFET’s output capacitance and parasitic inductances in prototype circuit. This loss could be reduced with professional manufacturing techniques.

Figs. 10(c) and 11(c) show that diode \( D_1 \) is turned off with ZCS at both conditions. Figs. 10(d) and 11(d) show that diode \( D_2 \) is turned off with ZCS at both conditions. Figs. 10(e) and 11(e) show waveforms of \( I_{Lr} \) and \( V_{Cr} \), which are in close agreement with the analytical waveforms shown in Fig. 3. Voltages \( V_{Cr,max} \) at both conditions are measured to be 143.1 and 142.5 V, respectively. They are close to analytical values of 145.8 and 143 V obtained from (14) and (15). Figs. 10(f) and 11(f) show waveform of \( v_{Cs} \).

Fig. 12 shows the theoretical and experimental voltage gains of the proposed converter under \( V_i = 28 \) V, \( R_o = 577 \) \( \Omega \). The experimental voltage gain is in close agreement with the theoretical voltage gain.

### TABLE I

<table>
<thead>
<tr>
<th>Component</th>
<th>Rating</th>
<th>Selected devices</th>
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<tbody>
<tr>
<td>Switch ( S_1 )</td>
<td>( V_{ph} ) 110 V</td>
<td>IRFP4568</td>
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<tr>
<td></td>
<td>( I_{max} ) 11.8 A</td>
<td></td>
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<tr>
<td>Snubber diodes ( D_{s1}, D_{s2} )</td>
<td>( V_{ph} ) 94 V</td>
<td>UG8DT</td>
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<td></td>
<td>( I_{max} ) 0.27 A</td>
<td></td>
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<tr>
<td>Output diodes ( D_1 ), ( D_2 )</td>
<td>( V_{ph} ) 380 V</td>
<td>VS-HFA04TB60</td>
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<td></td>
<td>( I_{max} ) 0.65 A</td>
<td></td>
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<td>Filter inductor ( L_i )</td>
<td>Inductance 100 ( \mu )H 8.85 A</td>
<td>Ferrite core PQ32/30</td>
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<tr>
<td>Snubber inductor ( L_s )</td>
<td>Inductance 5 ( \mu )H 0.8 A</td>
<td>Ferrite core EF16</td>
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<tr>
<td>Snubber capacitor ( C_s )</td>
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</tr>
<tr>
<td></td>
<td>( V_{ph} ) 82 V</td>
<td></td>
</tr>
<tr>
<td></td>
<td>( I_{max} ) 1.4 A</td>
<td></td>
</tr>
<tr>
<td>Clamp capacitor ( C_r )</td>
<td>Capacitance 82 ( \mu )F</td>
<td>FFB44D0826K</td>
</tr>
<tr>
<td></td>
<td>( V_{ph} ) 38 V</td>
<td></td>
</tr>
<tr>
<td></td>
<td>( I_{max} ) 7.6 A</td>
<td></td>
</tr>
<tr>
<td>Transformer ( T_i )</td>
<td>Leakage inductance 5 ( \mu )H</td>
<td>Ferrite core PQ32/30</td>
</tr>
<tr>
<td></td>
<td>Magnetic inductance 93 ( \mu )H</td>
<td></td>
</tr>
<tr>
<td></td>
<td>Turn ratio 1:5</td>
<td></td>
</tr>
<tr>
<td></td>
<td>VA 273V/3A</td>
<td></td>
</tr>
<tr>
<td>Resonant capacitor ( C_i )</td>
<td>Capacitance 560 nF</td>
<td>ECW-FD2W564K4</td>
</tr>
<tr>
<td></td>
<td>( V_{ph} ) 196 V</td>
<td></td>
</tr>
<tr>
<td></td>
<td>( I_{max} ) 1.6 A</td>
<td></td>
</tr>
<tr>
<td>Output capacitor ( C_o )</td>
<td>Capacitance 1 ( \mu )F</td>
<td>ECQ-E2W105KH</td>
</tr>
<tr>
<td></td>
<td>( V_{ph} ) 380 V</td>
<td></td>
</tr>
<tr>
<td></td>
<td>( I_{max} ) 0.87 A</td>
<td></td>
</tr>
</tbody>
</table>
Fig. 10. Experimental waveforms at $V_i = 28$ V and $P_o = 250$ W: (a) switch $S_1$ at turn-on, (b) switch $S_1$ at turn-off, (c) diode $D_1$ at turn-off, (d) diode $D_2$ at turn-off, (e) current $i_{Lr}$ and voltage $v_{Cr}$, and (f) voltage $v_{Cs}$.

Fig. 11. Experimental waveforms at $V_i = 28$ V and $P_o = 125$ W: (a) switch $S_1$ at turn-on, (b) switch $S_1$ at turn-off, (c) diode $D_1$ at turn-off, (d) diode $D_2$ at turn-off, (e) current $i_{Lr}$ and voltage $v_{Cr}$, and (f) voltage $v_{Cs}$.

Fig. 12. Theoretical and experimental voltage gain of the proposed converter under $V_i = 28$ V and $R_o = 577$ Ω.

Fig. 13. Measured efficiency of the proposed converter.

Fig. 14. Loss analysis of the proposed converter at full load ($V_i = 38$ V).

The efficiency of the proposed converter is measured by YOKOGAWA WT3000 and shown in Fig. 13. The maximum measured efficiency of the proposed converter is 97.0% at 200 W when input voltage is 38 V. The measured full-load efficiencies according to input voltage variation are 96.0% at $V_i = 28$ V and 96.9% at $V_i = 38$ V, respectively. The European efficiency of the proposed converter is 96.26%.

Fig. 14 shows loss analysis of the proposed converter at full load when input voltage is 38 V. The total loss of the proposed converter is 7.77 W. The large portion of the losses comes from losses of the transformer and conduction losses of diode $D_1$ and $D_2$, which are 47.6% and 22.0% of the total loss, respectively. Fig. 15 shows the photograph of the proposed prototype.

IV. CONCLUSION

In this paper, a soft-switched single switch isolated converter was proposed for step-up application such as MIC, portable fuel cell systems, and vehicle inverters. Improved features such
as fully soft-switched characteristics of switch and diode, low-rated lossless snubber, and reduced transformer volume make the proposed converter achieve lower cost and higher power density compared to the conventional flyback-based converter. Experimental results on a 100 kHz, 250 W prototype are provided to validate the proposed concept. The maximum measured efficiency of 97.0% was obtained at 200 W.

REFERENCES


