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Analysis of Unidirectional Secondary Resonant Single Active Bridge DC–DC Converter

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Abstract: A compact and highly efficient unidirectional DC–DC converter is required as a battery charger for electrical vehicles, which will rapidly become widespread in the near future. The single active bridge (SAB) converter is proposed as a simple and high-frequency isolated unidirectional converter, which is comprised of an active H-bridge converter in the primary side, an isolated high frequency transformer, and a rectifying secondary diode bridge output circuit. This paper presents a novel, unidirectional, high-frequency isolated DC–DC converter called a Secondary Resonant Single Active Bridge (SR–SAB) DC–DC converter. The circuit topology of the SR–SAB converter is a resonant capacitor connected to each diode in parallel in order to construct the series resonant circuit in the secondary circuit. As a result, the SR–SAB converter achieves a higher total power factor at the high frequency transformer and a unity voltage conversion ratio under the unity transformer turns ratio. Small and nonsignificant overshoot values of current and voltage waveforms are observed. Soft-switching commutations of the primary H-bridge circuit and the soft recovery of secondary diode bridge are achieved. The operating philosophy and design method of the proposed converter are presented. Output power control using transformer frequency variation is proposed. The effectiveness of the SR–SAB converter was verified by experiments using a 1 kW, 48 VDC, and 20 kHz laboratory prototype.

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Keywords: secondary resonant single active bridge (SR–SAB) converter; battery charger; DC–DC converter; isolated converter; soft switching; unidirectional converter

1. Introduction

In 1991, the isolated bidirectional DC–DC converter, which adopts the dual-active-bridge (DAB) configuration and uses a soft-switching technique to achieve high efficiency, was proposed [1]. The transferred power of the DAB converter is regulated by the induced voltage drop on the inductor produced by the relative position of different phase shift angles between the primary and secondary transformer AC voltages. The soft-switching pattern of the DAB converter was resolved during this process [2]. The DAB converter achieved a high total power factor (TPF) at the high frequency transformer and a unity voltage conversion ratio under the unity transformer turns ratio. Many papers related to DAB converters have been published [3]. Papers on the soft-switching control method in the low power supply range [4,5] and the utilization of GaN FETs to construct high-efficiency converters [6] have also been published. In an additional study, a DAB converter was applied to an LED driver [7].

Several circuit topologies of unidirectional isolated DC–DC converters have been proposed for their application in battery chargers. One unidirectional topology based on the DAB converter is a configuration in which two switches in the secondary H-bridge converter are replaced with diodes. For wide-range operation, three operation modes of boost, buck-boost, and buck are proposed [8]. The light-load conversion efficiency was improved significantly by using the proposed control strategy [9].

The unidirectional configuration, with the diode rectifier circuit adding the choke coil in the secondary circuit, can obtain high efficiency [10]. The topology achieved a high TPF at the high-frequency transformer and a unity voltage conversion ratio under the unity transformer turns ratio. However, a voltage surge occurred, owing to the resonance between the transformer leakage inductance and the inherent parasitic capacitance associated with the secondary diodes. To suppress the voltage surge, an RCD snubber circuit was installed at the output of the diode bridge rectifier circuit [11]. A hybrid-switching converter was then proposed to achieve minimal voltage stress of the diode bridge rectifier circuit [12].

Another unidirectional topology of an SAB converter is a simpler configuration in which the secondary H-bridge of the DAB converter is replaced with a diode bridge rectifier circuit. When this was trialed, the SAB converter achieved soft switching in the same way as the DAB converter [13]. The partial resonant SAB converter, in which a capacitor is connected to each primary switch in parallel, was proposed to reduce the peak current and conduction loss [14]. By utilizing SiC switching devices in DC–DC circuits, high-efficiency performance was achieved [15]. Papers on the application of the SAB converter to the wind turbine converter were published [16,17]. However, the SAB converter does not operate at the unity voltage conversion ratio. Therefore, the DC voltage conversion ratio and the total power factor of the transformer in the SAB converter were low [1].

This paper presents a novel, unidirectional, high-frequency isolated DC–DC converter called a secondary resonant single active bridge (SR–SAB) DC–DC converter, as shown in Figure 1 [18]. In the SR–SAB DC–DC converter, a resonant capacitor C_r is connected in parallel to each diode in the secondary rectifying diode circuit of the SAB converter. A trapezoidal transformer current waveform, similar to that of a DAB converter, is obtained because the transformer current sign at the zero-crossing region is softly switched by the series resonance between the transformer leakage inductance L and the resonant capacitors C_r of the secondary diodes. As a result, the SR–SAB converter reaches the unity DC voltage conversion ratio and improves the total power factor of the transformer compared with the SAB converter. The behavior and design of the proposed SR–SAB converter will be explained. The transformer frequency which is equal to the switching frequency can be independently selected at a frequency lower than the resonant frequency. The leakage inductance can be designed to be sufficiently small enough to be incorporated into the high-frequency transformer. The output power can be adjusted over a wide range by controlling the transformer frequency.

The effectiveness of the proposed converter is verified through experiments using a 1 kW, 48 V, and 20 kHz rating laboratory prototype.

Figure 1. Proposed unidirectional secondary resonant single active bridge (SR–SAB) DC–DC converter circuit.

2. Conventional SAB DC–DC Converter

Figure 2 shows the conventional SAB DC–DC converter. In the circuit configuration of Figure 2a, the SAB converter is comprised of a primary H-bridge converter, an isolated high-frequency transformer, and a secondary rectifying diode bridge circuit. The

primary H-bridge converter is constructed from four switches, $R^+ - S^-$, with the small soft-switching capacitor C_s in parallel. The leakage inductance L of the transformer is expressed as the equivalent leakage inductance converted to the secondary side. By denoting the number of turns of the transformer's primary coil with N_1 , and the number of turns of the transformer's secondary coil with N_2 , the turns ratio can be expressed as $a = N_1/N_2$. Figure 2b shows the secondary equivalent circuit. The primary voltage component $v'_1 (= v_1/a)$, which transfers the primary voltage v_1 to the secondary side, is represented as the voltage across the secondary magnetizing inductance of the transformer.

Figure 2. Conventional SAB DC-DC converter.

Figure 3 shows the voltage and current operation waveforms of the transformer in the steady state of the conventional unidirectional SAB converter. The conditions of the waveforms are presented at a unity turns ratio of $a = 1$ for the isolated transformer while neglecting the transformer magnetizing current. The primary H-bridge converter generates a square AC voltage $v_1 (= v'_1)$ with the amplitude of the input DC voltage V_{in} , and frequency $f_s (= 1/2T_s)$ by controlling the switches $R^+ - S^-$ with a 50 percent duty cycle. If the amplitude V_{in} of the primary voltage v'_1 is equal to the output DC voltage V_{out} in Figure 3, the secondary diodes $U^+ - V^-$ cannot be turned on, and no power is sent to the secondary side in the SAB converter. Therefore, the SAB converter does not work at a unity DC voltage conversion ratio $M_v = V_{out}/V_{in} = 1$. In the voltage waveforms in Figure 3, the amplitude V_{in} of the primary voltage v'_1 is higher than the output DC voltage V_{out} . The transformer currents i_1 and i_2 flow because of the difference in voltage between the primary voltage v'_1 and the output DC voltage V_{out} .

Figure 3. Voltage and current operation waveforms of conventional SAB DC-DC converter.

The analytical transformer current i_2 in the positive half period T_s of primary voltage v_1' is derived. From Figure 2b, the following voltage equation is obtained:

$$v_1' = L \frac{di_2}{dt} + v_2 \quad (1)$$

In the duration T_a ($t_a \leq t \leq t_b$), the primary voltage v_1' in Figure 3 is larger than the output DC voltage V_{out} and the negative transformer current i_2 flows. The diodes U^- and V^+ are in the on-state. Substituting $v_1' = V_{in}$, $v_2 = -V_{out}$, and the initial current $i_2(t_a) = -I_0$ into (1), the secondary current $i_2(t)$ in the duration T_a is obtained as follows:

$$i_2(t) = \frac{V_{in} + V_{out}}{L} (t - t_a) - I_0 \quad (t_a \leq t \leq t_b) \quad (2)$$

Because the secondary current $i_2(t_b) = 0$ in (2), the duration T_a is obtained in (3):

$$T_a = t_b - t_a = L \frac{I_0}{V_{in} + V_{out}} \quad (3)$$

In the duration T_b ($t_b \leq t \leq t_a + T_s$), the positive transformer current i_2 flows and the diodes U^+ and V^- are in the on-state. Substituting $v_1' = V_{in}$ and $v_2 = V_{out}$ into (1), the secondary current $i_2(t)$ in the duration T_b is obtained as follows:

$$i_2(t) = \frac{V_{in} - V_{out}}{L} (t - t_b) \quad (t_b \leq t \leq t_a + T_s) \quad (4)$$

The duration T_b is expressed using (3) in (5):

$$T_b = T_s - T_a = T_s - L \frac{I_0}{V_{in} + V_{out}} \quad (5)$$

Because the secondary current $i_2(t_a + T_s) = I_o$ in (4), the peak current I_o is solved as follows:

$$i_2(t_a + T_s) = \frac{V_{in} - V_{out}}{L} \left(T - L \frac{I_o}{V_{in} + V_{out}} \right) = I_o$$

$$I_o = \frac{V_{in}^2 - V_{out}^2}{2LV_{in}} T_s \quad (6)$$

Substituting I_o in Figure 3 into Equations (3) and (5), the durations T_a and T_b are obtained as follows:

$$T_a = \frac{V_{in} - V_{out}}{2V_{in}} T_s \quad (7)$$

$$T_b = \frac{V_{in} + V_{out}}{2V_{in}} T_s \quad (8)$$

The total power factor TPF of the primary side is defined in (9) by using the output power P_{out} and the effective values of voltage V_1 and current I_1 :

$$TPF = \frac{P_{out}}{V_1 I_1} \quad (9)$$

The output power P_{out} can be calculated using the following equation:

$$P_{out} = \frac{1}{T_s} \int_{t_a}^{t_a + T_s} V_{out} i_2 dt = \frac{(V_{in}^2 - V_{out}^2) V_{out} T_s}{4LV_{in}}. \quad (10)$$

The effective value of primary current I_1 is calculated as follows:

$$I_1 = \frac{1}{\sqrt{3}} I_o = \frac{1}{\sqrt{3}} \frac{V_{in}^2 - V_{out}^2}{2LV_{in}} T_s \quad (11)$$

Substituting (10), (11), and the effective values of primary voltage $V_1 = V_{in}$ into (9), the total power factor TPF is obtained as follows:

$$TPF = \frac{\sqrt{3} V_{out}}{2V_{in}} \quad (12)$$

A design example of the conventional SAB and proposed SR-SAB converters are explained. Table 1 shows the common specifications of the converter set at $P_{out} = 1.0$ kW, $V_{out} = 48$ V, and $f_s = 20$ kHz. Table 2 shows the comparisons of calculated characteristics between the conventional and proposed converters. In the conventional converter, the input DC voltage $V_{in} = 62.5$ V, the voltage conversion ratio $M_v = V_{out}/V_{in} = 1.30$, the peak current $I_o = 41.7$ A, and the total power factor $TPF = 0.66$. The characteristics of the proposed SR-SAB converter in Table 2, which are derived from the theory in Section 4, improves significantly compared with those of the conventional converter.

Table 1. Common specifications for conventional SAB and proposed SR-SAB converters on 1 kW output power.

Output power P_{out}	1.0 kW
Output voltage V_{out}	48 V
Turns ratio a	1
Leakage inductance L	7.7 μ H
Transformer frequency f_s	20 kHz

Table 2. Comparisons of characteristics between conventional SAB and proposed SR–SAB converters on 1 kW output power.

Converter	SAB	SR–SAB
Input voltage V_{in}	62.5 V	48.0 V
Voltage conversion ratio M_v	1.30	1.00
Peak current	41.7 A	25.0 A
Effective current value I_1	24.1 A	23.5 A
Total power factor TPF	0.66	0.89
Resonant capacitor C_r	–	0.52 μ F

3. Unidirectional Secondary Resonant Single Active Bridge (SR–SAB) DC–DC Converter

3.1. Main Circuit Configuration and Behavior

The circuit configuration of the proposed unidirectional SR–SAB DC–DC converter in Figure 1 adds the resonant capacitor C_r in parallel with the secondary diodes $U^+ - V^-$ in the conventional SAB converter shown in Figure 2. Figure 4 shows the voltage and current operation waveforms of the transformer as it corresponds to the proposed SR–SAB converter given the condition that the unity turns ratio $a = 1$ for the high-frequency transformer while neglecting the transformer magnetizing current. The SR–SAB converter achieves the unity input and output voltage conversion ratio $M_v = 1$ ($V_{in} = V_{out}$).

Figure 4. Voltage and current operation waveforms of the proposed SR–SAB DC–DC converter.

The primary H-bridge converter produces the rectangular AC voltage v_1 with the amplitude V_{in} and switching frequency $f_s (= 1/2T_s)$ by alternately switching the switches

$R^+ - S^-$ with a 50 percent duty cycle. The secondary voltage v_2 with the rectangular waveform lagged by the primary voltage v_1 is obtained. The magnetizing current is very small and neglected, leading to the production of the identical trapezoidal current waveforms for the primary current as well as the secondary current. The rectified current i_{out} flows according to the conditions of the on-state diodes.

3.2. Soft-Switching Commutation of Primary Converter

The primary converter achieves soft switching in all commutations. Figure 5 shows the soft-switching modes of the primary H-bridge converter from R^- and S^+ switches to R^+ and S^- switches.

Figure 5. Soft-switching operation modes of the primary H-bridge switches circuit.

In Mode 1–1, R^- and S^+ switches are in the turn-on state, generating the primary voltage v_1 of $-V_{in}$. Primary current i_1 with negative value $-I_n$ is conducted through R^- and S^+ switches. The voltages on the parallel capacitors of R^- and S^+ switches are kept at zero. When R^- and S^+ switches are changed to the turn-off state, zero voltage switching (ZVS) of R^- and S^+ switches is obtained because the voltage of the parallel capacitors are zero. After that, the circuit changes to Mode 1–2.

In Mode 1–2, the primary current i_1 with the negative value $-I_n$ starts to flow through the four capacitors. The voltage of the parallel capacitors C_s of R^- and S^+ switches increases, and the voltage of the parallel capacitors C_s of R^+ and S^- switches decreases. The primary voltage v_1 interchanges from $-V_{in}$ to V_{in} . When the voltages on parallel capacitor C_s of R^+ and S^- switches is reduced to zero, the anti-parallel diodes on R^+ and S^- switches are switched to the turn-on state and start conducting. After that, the circuit changes to Mode 1–3.

In Mode 1–3, the switching-on gate signals for R^+ and S^- switches are given, leading the primary current i_1 to increase continuously toward zero. When the sign of the primary current i_1 changes to positive, the circuit changes to Mode 1–4.

In Mode 1–4, R^+ and S^- switches are changed to the on-state by zero-voltage switching because of the zero voltage of the parallel capacitor.

3.3. Operation Analysis of Secondary Diode Rectifier Circuit

Figure 6 shows the commutation modes in the secondary side of the diode rectifying circuit, swapping between the pair of diodes U^- and V^+ and the pair of diodes U^+ and V^- under the unity input and output voltage conversion ratio $M_v = 1$ ($V_{in} = V_{out}$).

Figure 6. Commutations of the secondary diode rectifier circuit.

In Mode 2-1, the primary voltage $v'_1 (= V_1/a)$ of $-V_{in} (= -V_{out})$ is generated, and the secondary current i_2 with negative value $-I_n$ flows through the pair of diodes U^- and V^+ . The voltages on parallel capacitors C_r of diodes U^- and V^+ are both zero, and the voltages on parallel capacitors C_r of diodes U^+ and V^- are fully charged to the value V_{out} of output voltage. When the value of primary voltage v'_1 changes from $-V_{in}$ to $V_{in} (= V_{out})$ at the time $t = t_2$, the circuit starts to enter Mode 2-2.

In Mode 2-2, within the duration $t_2 \leq t \leq t_3$, the primary voltage v'_1 changes to the positive value $V_{in} (= V_{out})$. The diodes U^- and V^+ remain on-state, and the negative transformer current i_2 flows. Substituting $v'_1 = V_{out}$, $v_2 = -V_{out}$ and the initial current $i_2(t_2) = -I_n$ into formula (1), the equation regarding the secondary current $i_2(t)$ in Mode 2-2 can be derived as follows:

$$i_2(t) = \frac{2V_{out}}{L}(t - t_2) - I_n \quad (t_2 \leq t \leq t_3) \quad (13)$$

The secondary current $i_2(t)$ increases linearly toward zero value, which is shown in Figure 4. Upon the timing $t = t_3$, the secondary current $i_2(t)$ reaches zero, and Mode 2-2 is completed. The duration $T_2 = t_3 - t_2$ of Mode 2-2, is obtained from (13) by the following equation:

$$T_2 = t_3 - t_2 = \frac{LI_n}{2V_{out}} \quad (14)$$

At $i_2(t_3) = 0$, all four diodes of the secondary rectifier circuit change to turn-off state, and then Mode 2-3 begins.

In Mode 2-3, the voltages on parallel capacitors C_r of diodes U^- and V^+ are both zero, and the voltages on parallel capacitors C_r of diodes U^+ and V^- are both charged to the output voltage V_{out} at $t = t_3$. The inductor L , together with four capacitors C_r , forms

an LC resonating circuit and the resonance occurs. Due to the symmetry property of the circuit, the secondary current flow through each pair of capacitors C_r with the half value $i_2/2$. Therefore, in Mode 2–3, the voltage equation is formulated as follows:

$$v'_1 = L \frac{di_2}{dt} + \frac{2}{C_r} \int_{t_3}^t \frac{i_2}{2} dt - V_{out} \quad (15)$$

Substituting $v'_1 = V_{out}$ and $i_2(t_3) = 0$ into (15), the secondary current $i_2(t)$ in the duration $t_3 \leq t \leq t_4$ of Mode 2–3 is obtained as follows:

$$i_2(t) = 2V_{out} \sqrt{\frac{C_r}{L}} \sin \frac{t-t_3}{\sqrt{LC_r}} \quad (t_3 \leq t \leq t_4) \quad (16)$$

The secondary voltage v_2 is derived utilizing the secondary current $i_2(t)$ in Equation (16) as follows:

$$v_2(t) = v'_1(t) - L \frac{di_2}{dt} = V_{out} \left(1 - 2 \cos \frac{t-t_3}{\sqrt{LC_r}} \right) \quad (17)$$

The secondary current i_2 and voltage v_2 in (16) and (17) are both sinusoidal waveforms, which are shown in Figure 4. When the secondary current $i_2(t)$ becomes I_n and voltage $v_2(t)$ becomes $v_2(t) = V_{out}$ at $t = t_4$, simultaneously, Mode 2–3 is completed. Because at $t = t_4$, the phase angle is $\pi/2$ as in Figure 4, the amplitude value I_n of the secondary current i_2 and the duration T_3 in Mode 2–3 are obtained from (16) and (17) by the following equations:

$$I_n = 2V_{out} \sqrt{\frac{C_r}{L}} \quad (18)$$

$$T_3 = t_4 - t_3 = \frac{\pi}{2} \sqrt{LC_r} \quad (19)$$

The secondary current i_2 in Mode 2–2 in (13) and duration T_2 in (14) is applied to Equation (18), and is rewritten in the following equations:

$$i_2(t) = \frac{2V_{out}}{L} (t - t_2 - \sqrt{LC_r}) \quad (t_2 \leq t \leq t_3) \quad (20)$$

$$T_2 = \sqrt{LC_r} \quad (21)$$

At $t = t_4$, the voltages on the parallel capacitors of the diodes U^+ and V^- reduce to zero, then the diodes U^+ and V^- start conducting, and the circuit shifts to Mode 2–4 in Figure 6.

In Mode 2–4, substituting the primary voltage $v'_1 = V_{out}$, the secondary voltage $v_2 = V_{out}$, and the initial value of the secondary current $i_2(t_4) = I_n$ into (1), in Mode 2–4, the secondary current $i_2(t)$ is given by the following equation:

$$i_2(t) = I_n = 2V_{out} \sqrt{\frac{C_r}{L}} \quad (t_4 \leq t) \quad (22)$$

The output power P_{out} is calculated by the average transferring power for the half period T_s of the isolated high-frequency transformer using the rectified current i_{out} in Figure 4:

$$P_{out} = \frac{1}{T_s} \int_{t_2}^{t_2+T_s} V_{out} i_{out} dt = 2V_{out}^2 \sqrt{\frac{C_r}{L}} \left(1 - \frac{1+\pi}{2T_s} \sqrt{LC_r} \right) \quad (23)$$

The effective value of primary current I_1 is calculated as follows:

$$I_1 = \sqrt{\frac{1}{T_s} \int_{t_2}^{t_2+T_s} i_2^2 dt} = 2V_{out} \sqrt{\frac{C_r}{L} \left\{ 1 - \left(\frac{2}{3} + \frac{\pi}{4} \right) \frac{\sqrt{LC_r}}{T_s} \right\}} \quad (24)$$

Substituting (23), (24), and $V_1 = V_{out}$, into (9), the total power factor TPF is obtained as follows:

$$TPF = \frac{1 - \frac{1+\pi}{2T_s} \sqrt{LC_r}}{\sqrt{1 - \left(\frac{2}{3} + \frac{\pi}{4}\right) \frac{\sqrt{LC_r}}{T_s}}} \quad (25)$$

3.4. Output Power Control by Frequency Variation

Under the condition of the unity input and output voltage conversion ratio $M_v = 1$ ($V_{in} = V_{out}$), the output power P_{out} in (23) is a function of the half period T_s of the high-frequency transformer. If the half period T_s is shorter than that in Figure 4, the duration of Mode 2–4 is shorter and the output power P_{out} decreases. The output power P_{out} can be controlled by the transformer frequency f_s given by the following equation:

$$f_s = \frac{1}{2T_s} \quad (26)$$

The resonant frequency f_o in the circuit is given in (27):

$$f_o = \frac{1}{2\pi\sqrt{LC_r}} \quad (27)$$

Substituting (26) and (27) into (23), the output power P_{out} is rewritten as a function of the frequency ratio f_s/f_o as follows:

$$P_{out} = 2V_{out}^2 \sqrt{\frac{C_r}{L}} \left\{ 1 - \left(\frac{1}{2} + \frac{1}{2\pi} \right) \frac{f_s}{f_o} \right\} \quad (28)$$

Figure 7 shows the characteristics of the output power P_{out} with respect to the frequency ratio f_s/f_o based on (28). The resonant frequency is determined by the circuit parameters and is a constant value.

Figure 7. Output power characteristic with respect to the switching frequency of the transformer.

By increasing the transformer frequency f_s , the output power P_{out} can be reduced. In Figure 7, the output power is notarized as $P_{out} = 1$ at the frequency ratio $(f_s/f_o) = 1/4$. The maximum frequency ratio $(f_s/f_o)_{max}$ is the condition in which the duration of Mode 2–4 in

Figure 4 is zero. Because it means $T_s = T_2 + T_3$, the maximum frequency ratio $(f_s/f_o)_{max}$ is given by the following equation:

$$\left(\frac{f_s}{f_o}\right)_{max} = \frac{2\pi\sqrt{LC_r}}{2(T_2 + T_3)} = \frac{2\pi}{2 + \pi} = 1.22 \quad (29)$$

The highest transformer frequency satisfying the output power P_{out} of (28) is $(f_s/f_o)_{max} = 1.22$. In Figure 7, the output power P_{out} at $(f_s/f_o)_{max} = 1.22$ can be reduced to 0.23 of the rated output power $P_{out} = 1$. When the frequency ratio is higher than $(f_s/f_o)_{max} = 1.22$, the output power P_{out} can be reduced further than that in (28).

4. Design of Circuit Parameters

The design and calculation method for the circuit parameters of the proposed SR-SAB converter are introduced here. Table 3 shows the specifications and designed parameters of the proposed SR-SAB converter in a laboratory experimental system. The SR-SAB converter is designed with the rated specifications of output power $P_{out} = 1$ kW, output voltage $V_{out} = 48$ V, input voltage $V_{in} = 265$ V, transformer frequency $f_s = 20$ kHz, and frequency ratio $f_s/f_o = 1/4$. The turn ratio of the high-frequency transformer a is equal to the ratio of the input and output voltages, as seen below:

$$a = \frac{V_{in}}{V_{out}} = 5.5 \quad (30)$$

Table 3. Specifications and designed parameters of proposed SR-SAB converter.

(A) Rated specifications.		
Output power P_{out}		1 kW
Output voltage V_{out}		48 V
Input voltage V_{in}		265 V
Transformer frequency f_s		20 kHz
Frequency ratio f_s/f_o		1/4
(B) Designed parameters.		
Turns ratio a		5.5
Leakage inductance L		7.7 μ H
Resonant capacitor C_r		520 nF
Soft-switching capacitor C_s		1 nF
DC capacitors C_1, C_2		1500 μ F

Because transformer frequency is selected as $f_s = 20$ kHz, the half-cycle period T_s , resonant frequency f_o , and resonant angular frequency $1/\sqrt{LC_r}$ are given in the following equations:

$$T_s = \frac{1}{2f_s} = 25 \mu\text{s} \quad (31)$$

$$f_o = 4f_s = 80 \text{ kHz} \quad (32)$$

$$\frac{1}{\sqrt{LC_r}} = 2\pi f_o \simeq 5.0 \times 10^5 \text{ rad/s} \quad (33)$$

The duration T_2 of Mode 2–2 in (21) and the duration T_3 of Mode 2–3 in (19) are calculated as (34) and (35), respectively.

$$T_2 = \sqrt{LC_r} = 2.0 \mu\text{s} \quad (34)$$

$$T_3 = \frac{\pi}{2} \sqrt{LC_r} = 3.14 \mu\text{s} \quad (35)$$

The amplitude I_n of the secondary current i_2 is obtained from (18) and (23) as follows:

$$I_n = 2V_{out} \sqrt{\frac{C_r}{L}} = \frac{P_{out}}{V_{out} \left(1 - \frac{1+\pi}{2T_s} \sqrt{LC_r}\right)} = 25 \text{ A} \quad (36)$$

From (36), the characteristic impedance $\sqrt{L/C_r}$ is obtained the following equation:

$$\sqrt{\frac{L}{C_r}} = \frac{2V_{out}}{I_n} = 3.84 \text{ } \Omega \quad (37)$$

From (33) and (37), the leakage inductance L and resonant capacitor C_r are determined:

$$L = \sqrt{LC_r} \sqrt{\frac{L}{C_r}} = 7.7 \text{ } \mu\text{H} \quad (38)$$

$$C_r = \sqrt{LC_r} \sqrt{\frac{C_r}{L}} = 0.52 \text{ } \mu\text{F} \quad (39)$$

Regarding to the primary H-bridge converter of Mode 1–2 in Figure 5, each soft-switching capacitor C_s is charged between 0 and voltage V_{in} . Half of the primary current $i_1 (= I_n/a)$ flows to the capacitor. Therefore, the duration T_{12} of Mode 1–2 is obtained as follows:

$$T_{12} = \frac{C_s V_{in}}{I_n/2a} \quad (40)$$

By giving the duration $T_{12} = 0.1 \text{ } \mu\text{s}$ of Mode 1–2, the soft switching capacitor C_s is designed as follows:

$$C_s = \frac{I_n T_{12}}{2a V_{in}} = 0.8 \times 10^{-9} \simeq 1 \text{ nF} \quad (41)$$

In order to compare them with the characteristics in Table 2, the characteristics of the proposed SR–SAB converter at unity turn ratio $a = 1$ are calculated. The effective values of the primary current $I_1 = 23.5 \text{ A}$ in (24) and the total power factor $TPF = 0.89$ in (25) are obtained using $V_{out} = 48 \text{ V}$, the leakage inductance L in (38), and the resonant capacitor C_r in (39). As mentioned in Section 2, the proposed SR–SAB converter can operate at a voltage conversion ratio $M_v = 1.0$, and the peak current and total power factor TPF in Table 2 show significant improvement compared with those of the conventional SAB converter.

5. Experimental Results

The operation and behavior of the 1 kW, 265 V/48 V SR–SAB converter described in Table 3 is verified by experimental waveforms. The primary H-bridge is composed of four Rohm SCT3022AL SiC MOSFETs, and the secondary diode bridge is composed of four VS–100BGQ100 Schottky diodes. The inductor value $L = 7.7 \text{ } \mu\text{H}$ is derived from the leakage inductance of the high-frequency transformer. Furthermore, the resonant capacitor C_r is composed of one 330 nF and two 100 nF film capacitors in parallel. The dead time $T_{dead} = 0.2 \text{ } \mu\text{s}$ of the primary switches is used in the experimental circuit.

Figure 8a shows the experimental set up of the proposed SR–SAB DC–DC converter. The DC input source $V_{in} = 265 \text{ V}$ is obtained from the three-phase full-bridge diode rectifier circuit from the three-phase supply voltages of 60 Hz and 200 V, while the 48 V DC load is the regenerative dc power supplier, pCUBE. The controller is the PE–Expert 4 system using DSP TMS320C6657, and the DL850 Yokogawa is used for measuring the voltage and current waveforms. Finally, Xviewer displaying software is used for displaying the voltage and current waveforms directly from the DL850.

Figure 8b shows the electrical diagram of the proposed SR–SAB DC–DC converter with the detailed descriptions of the circuit that has been presented above in Figure 8a.

(a)

(b)

Figure 8. (a) Proposed SR-SAB DC-DC 265VDC/48VDC converter experimental setup; (b) Proposed SR-SAB DC-DC 265VDC/48VDC converter electrical diagram.

Figure 9 shows the experimental voltage and current waveforms during two periods of the high-frequency transformer at $f_s = 20$ kHz. The sequential waveforms are the gate driving signal v_g of the switch R^+ , primary H-bridge voltage v_1 , transformer secondary side voltages v_2 , output DC voltage V_{out} , secondary side current i_2 , and output DC current i_{dc} . The gate signal v_g and the half period $T_s = 25 \mu s$ generates the primary H-bridge voltage v_1 with the amplitude $V_{in} = 265$ V and switching frequency $f_s = 20$ kHz. The secondary rectangular waveform voltage v_2 is also observed. The secondary current i_2 flows in the circuit according to the difference in the phase-shift angle between v'_1 and v_2 . The amplitude I_n of the secondary current i_2 is at the designed value of 25 A. The duration $T_2 = 2 \mu s$ of Mode 2–2 in (34) and $T_3 = 3.14 \mu s$ of Mode 2–3 in (35) are obtained. The average output DC current i_{dc} is 21 A. All the experimental waveform results are similar to the theoretical results.

Figure 9. Experimental voltage and current waveforms at $f_s = 20$ kHz.

Figure 10 shows the experimental voltage and current waveforms of the isolated high-frequency transformer at $f_s = 80$ kHz. The order of the waveforms shown in Figure 10 is the same as that in Figure 9. The gate driving signal v_g with the half period $T_s = 6.25 \mu s$ generates the primary H-bridge voltage v_1 with the amplitude $V_{in} = 265$ V and the switching frequency $f_s = 80$ kHz. The amplitude I_n of the secondary current $i_2 = 25$ A, while the duration $T_2 = 2 \mu s$ of Mode 2–2 in (34) and $T_3 = 3.14 \mu s$ of Mode 2–3 in (35) are the same as in Figure 9. The duration of Mode 2–4 is shorter than that in Figure 9. As a result, the average output DC current i_{dc} is reduced to 8.75 A.

Figure 10. Experimental voltage and current waveforms at $f_s = 80$ kHz.

Figure 11 shows the experimental voltage and current waveforms of the isolated high-frequency transformer at $f_s = 100$ kHz. The order of the waveforms shown in Figure 11 is the same as that in Figure 9. The gate driving signal v_g with the half period $T_s = 5 \mu\text{s}$ generates the primary H-bridge voltage v_1 with the amplitude $V_{in} = 265$ V and the switching frequency $f_s = 100$ kHz. The amplitude I_n of secondary current $i_2 = 25$ A, while the duration $T_2 = 2 \mu\text{s}$ of Mode 2–2 in (34) and also $T_3 = 3.14 \mu\text{s}$ of Mode 2–3 in (35) are the same as in Figure 10. The duration of Mode 2–4 is shorter than that in Figure 10. As a result, the average output DC current i_{dc} is reduced to 4.6 A.

Figure 11. Experimental voltage and current waveforms at $f_s = 100$ kHz.

Figure 12 shows the experimental voltage and current waveforms during two and a half periods of the high-frequency transformer at $f_s = 5$ kHz. The order of the waveforms shown in Figure 12 is the same as that in Figure 9. The gate signal v_g with the half period $T_s = 100$ μ s generates the primary H-bridge voltage v_1 with the amplitude $V_{in} = 265$ V and switching frequency $f_s = 5$ kHz. The secondary rectangular waveform voltage v_2 is observed. The secondary current i_2 flows in the circuit according to the difference in the phase-shift angle between v_1' and v_2 . The amplitude I_n of the secondary current i_2 is at the designed value of 25 A. The total duration $T_2 + T_3 = 5.14$ μ s is obtained, while the duration $T_2 = 2$ μ s of Mode 2–2 in (34) and $T_3 = 3.14$ μ s of Mode 2–3 in (35) are achieved in the same way as in Figure 9. The average output DC current i_{dc} is 24.7 A. As can be seen from Figure 12, a very high TPF experimental value can be achieved in this case for SR–SAB compared to the calculated theoretical value ($TPF = 0.97$).

Figure 12. Experimental voltage and current waveforms at $f_s = 5$ kHz.

With all the experimental voltage and current waveforms, no high peak stress on the primary and secondary sides is observed as it is in some cases of the normal LC resonant circuit. The overshoot values of primary and secondary voltage in all cases are within the range of 5~10 percent, which is normal, and which is in fact another advantage of this topology because switching devices and capacitors can be selected that are small enough to reduce the overall size.

Figure 13 shows the experimental characteristics of the output power P_{out} versus the frequency ratio f_s/f_0 . When the transformer frequency f_s varies from 5 kHz to 100 kHz, the output power P_{out} gradually decreases from 1185 W to 220 W, and the characteristics coincide with the theoretical linear line in (28). The output power P_{out} can be regulated over a wide range by adjusting the frequency ratio f_s/f_0 from 0 to 1.22.

Figure 13. Characteristics of output power P_{out} versus switching frequency ratio f_s/f_o .

Figure 14 shows the efficiency of the converter for the output power range. The input as well as output electrical powers are both measured using a power analyzer (WT1800, YOKOGAWA). A maximum efficiency of 97.33% is achieved at $P_{out} = 1185$ W with the transformer frequency $f_s = 5$ kHz. At the rated output power $P_{out} = 1$ kW and the transformer frequency $f_s = 20$ kHz, the measured efficiency of the circuit is 95.8%. At a low output power $P_{out} = 220$ W at $f_s = 100$ kHz, the efficiency is 90.2%. When the transformer frequency f_s increases from 5 kHz to 100 kHz, the efficiency decreases owing to the reduction in the output power and the increase in the switching loss and iron loss of the transformer with a higher transformer frequency.

Figure 14. Efficiency versus output power characteristic.

6. Conclusions

This paper presented a unidirectional SR-SAB DC-DC converter, in which the resonant capacitor C_r connected to each diode in parallel in the secondary circuit of the SAB converter. The SR-SAB DC-DC converter achieves the unity DC voltage conversion ratio and improves the total power factor (TPF) of the transformer and the series resonance between inductor L and resonant capacitor C_r compared with the conventional SAB converter.

The overshoot value of the primary and secondary voltage is not significant, meaning that it is an advantage to select smaller rating devices. The soft-switching commutations in the primary H-bridge and secondary diode bridge are achieved. The behavior analysis and circuit parameter design of the proposed SR-SAB DC-DC converter were clarified. A control method for the output power using the transformer frequency was proposed. The characteristics of the proposed converter were verified by theory and experimental results using a laboratory prototype with an output power of 1 kW, transformer frequency of 20 kHz, and output DC voltage of 48 V.

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