Article

Bidirectional DC Converter with Frequency Control: Analysis and Implementation

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Abstract: In this paper, a direct current (dc) converter with the abilities of bidirectional power transfer and soft switching characteristics is studied and implemented. The circuit schematic of the developed dc converter is built by a half-bridge converter and a center-tapped rectifier with synchronous rectifier. Under forward power transfer, a half-bridge circuit is controlled to regulate the low-voltage side at a stable value. For backward power transfer, a center-tapped rectifier with synchronous rectifier is regulated to control the high-voltage side at the desired voltage value, and the half-bridge circuit is operated as a voltage doubler rectifier. Active power devices are operated at zero-voltage switching using a series resonant technique on the high-voltage side with frequency modulation and inductive load operation. The practicability of the developed converter is established from experiments with a laboratory prototype circuit.

Keywords: dc–dc converter; bidirectional power control; soft switching

1. Introduction

To reduce air pollution and climate change, renewable energy sources with power electronic conversion have been demanded and developed in recent years. Bidirectional power converters [1–18] have been developed for alternating current (ac) to direct current (dc) conversion and dc–dc conversion. In references [1–7], ac–dc bidirectional power converters are employed between an ac power source and dc bus voltage, and dc–ac bidirectional power converters are widely employed in ac drive systems. Direct current to dc bidirectional power converters [8–13] are demanded for battery-based storage systems. Bidirectional power flow dc converters with dual full-bridge circuit topology or half-bridge circuit topology were proposed to deliver power between high-voltage side dc bus and low-voltage side storage devices. Duty cycle control is normally used in dual full-bridge converters to control load voltage and attain soft switching characteristics. The main drawbacks of dual full-bridge converters are complicated control scheme and high conduction losses under low effective duty cycle case with high circulating current. Resonant techniques with a frequency-control scheme have been developed in modern power converters with high circuit efficiency due to wide load range of soft switching characteristics. The dual full-bridge resonant converter in reference [14] achieves soft switching characteristics in forward power transfer. However, the soft switching characteristics were lost in this circuit topology in backward power transfer due to the non-symmetric circuits in both power flow operations. The circuit topologies of dual full-bridge resonant converters in references [15–17] are symmetric for both forward and backward power transfer. Therefore, active switches can achieve soft switching at turn-on instant to eliminate switching loss. However, one parallel inductor is used on the high-voltage side. The circulating current in this parallel inductor will result in additional power loss when the circuit is operated under forward power transfer.

A soft switching resonant converter is studied and implemented in this paper to have the benefits of bidirectional power transfer ability, soft switching characteristics and less power devices.

The proposed circuit is constructed by a half-bridge circuit on the primary-side and a center-tapped rectifier on the secondary-side. The LLC resonant circuit with frequency control is used on high-voltage side to control secondary-side voltage, reach soft switching turn-on for main power devices. To achieve same LLC resonant characteristics for both forward and backward power transfer, a parallel inductor is connected between half-bridge leg and split input capacitors on the high-voltage side. Under forward power transfer, power switches on half-bridge circuit are controlled by frequency-modulation to generate a square voltage input to the resonant tank and regulate low-side voltage V_L . Power switches on a center-tapped rectifier with synchronous switches are operated as synchronous rectifiers for reducing the conduction losses on the low-voltage side. Under forward power flow, the parallel inductor on the high-voltage side is disconnected (ac power switch is off) to reduce the power loss on this parallel inductor. When the proposed circuit is worked under backward power transfer, the parallel inductor is used on the high-voltage side to achieve LLC resonant characteristics. Power switches on the center-tapped rectifier are controlled to create a square voltage waveform on the primary-side of the isolated transformer and regulate high-side voltage V_H . Finally, experimental results are presented to verify the feasibility of the proposed circuit.

2. Circuit Diagram

Figure 1a shows the circuit schematic of a conventional resonant converter with a center-tapped rectifier on the secondary side. The main drawback in Figure 1a is high conduction loss on secondary-side rectifier diodes for high current output. In order to reduce conduction loss on secondary-side diodes, synchronous switches instead of fast recovery diodes are used on the secondary-side as shown in Figure 1b. For achieving bidirectional power control, Figure 2a demonstrates the circuit structure of the developed bidirectional dc converter. The circuit differences between the proposed converter and conventional unidirectional half-bridge resonant converter shown in Figure 1b are one ac switch S and inductor L_{m1} used in the proposed circuit to achieve LLC resonant behavior for both forward and backward power transfer. Therefore, the advantages of a conventional unidirectional LLC resonant converter all exist in the proposed circuit. Half-bridge circuit is employed on the high voltage side to clamp voltage ratings of power semiconductor devices Q_1 and Q_2 at V_H . A center-tapped circuit topology with synchronous rectifier is employed on the secondary-side to reduce the conduction loss on the low-voltage side. An ac switch S is adopted on the primary-side to achieve resonant behavior when the adopted circuit is operated at backward power transfer. Figure 2b demonstrates the equivalent circuit of the developed circuit operated at forward power transfer from V_H to V_L . AC switch S is turned off. C_r , L_r and L_{m2} are operated to achieve soft switching turn-on of Q_1 and Q_2 . Power devices Q_3 and Q_4 on the low-voltage side operate as the synchronous switches. Therefore, the conduction losses on Q_3 and Q_4 are reduced at the low-voltage and high-current sides. Figure 2c illustrates circuit structure of the studied circuit operated at backward power transfer from V_L to V_H . At backward power transfer condition, power switch S is closed and Q_1 and Q_2 are turned off. L_{m1} , C_r and L_r are resonant. Thus, main switches Q_3 and Q_4 are turned on under zero-voltage switching. Table 1 gives the basic comparison between the proposed circuit and the conventional bidirectional dc–dc converters. It can be observed that the proposed converter has less component counts.

Figure 1. Circuit diagram of conventional resonant converter with (a) center-tapped rectifier on the secondary-side (b) with synchronous rectifier on the secondary-side.

Figure 2. Proposed circuit (a) circuit diagram (b) forward power transfer (c) reverse power transfer.

Table 1. Circuit topologies and component counts between the proposed converter and the other bidirectional dc–dc converters.

	Primary-Side	Secondary-Side	Component Counts	Control Scheme
Proposed converter	Half-bridge circuit	Center-tapped circuit	5 switches, 3 magnetic components, 1 resonant capacitor	Frequency control
Circuit topology in Reference [8]	Full-bridge circuit	Center-tapped circuit	6 switches, 2 magnetic components	Duty cycle control
Circuit topology in Reference [9]	Full-bridge circuit	Full-bridge circuit	8 switches, 2 magnetic components	Duty cycle control
Circuit topology in Reference [10]	Half-bridge circuit	Half-bridge circuit	6 switches, 3 magnetic components, 1 resonant capacitor	Frequency control
Circuit topology in Reference [14]	Full-bridge circuit	Full-bridge circuit	8 switches, 2 magnetic components, 1 resonant capacitor	Frequency control
Circuit topology in Reference [16]	Full-bridge circuit	Full-bridge circuit	8 switches, 3 magnetic components, 2 resonant capacitors	Frequency control

3. Principles of Operation

3.1. Forward Power Flow

If the converter is operated at forward power transfer as shown in Figure 2a, the power flow is from V_H terminal to V_L terminal. Power device S is controlled at the OFF state. Power switches Q_1 and Q_2 are regulated with frequency control to create a square voltage v_{ab} on the high-voltage side. Power switches Q_3 and Q_4 are operated as the synchronous switches to decrease conduction loss compared to the rectifier diodes with high voltage drop when diodes are conducting. The ON/OFF state of the synchronous switches are based on the current direction on the low-voltage side. Then a square voltage waveform can be created on the primary-side of the isolated transformer. L_{m2} , L_r and C_r are the magnetizing inductance, series resonant inductance and series resonant capacitance, respectively. Due to the inductive impedance of the resonant tank on the high-voltage side, as shown in Figure 3a, the soft switching characteristics of main switches Q_1 and Q_2 are achieved. The basic pulse-width

modulation signals for forward power transfer operation are illustrated in Figure 3a. There are six operation modes under f_r (series resonant frequency) $> f_{sw}$ (switching frequency). On the other hand, four operation modes per switching cycle can be observed if $f_{sw} > f_r$. Figure 4 illustrates the equivalent topological circuits for each mode of operation under forward power transfer. The principles of operation are discussed in what follows.

Figure 3. Resonant tank of the studied circuit (a) forward power transfer from V_H to V_L (b) backward power transfer from V_L to V_H .

Mode 1 [$t_0 \sim t_1$]: At t_0 , $v_{CQ1} = 0$. Since $i_{Q1}(t_0) < 0$, D_{Q1} conducts. Thus, Q_1 is turned on under zero voltage after t_0 . In mode 1, $i_{Lr} > i_{Lm2}$ so that $i_{Q3} < 0$ and D_{Q3} conducts. Thus, Q_3 is forced to turn on. Due to low turn-on resistance R_{on} , the conduction losses on Q_3 are reduced. In this mode, the voltage $v_{ab} = V_{CH1} = V_H/2$ and $v_{Lm2} = nV_L$. The magnetizing current i_{Lm2} will increase with the current slope nV_L/L_{m2} . In mode 1, the current variation on L_{m2} is calculated as $\Delta i_{Lm2,1} = nV_L \Delta t_{10}/L_{m2}$ where Δt_{10} is time duration in this mode. C_r and L_r are resonant under $v_{ab} = V_H/2$, $v_{Lm2} = nV_L$ and the series resonant frequency $f_r = 1/2\pi\sqrt{C_r L_r}$. If $f_r > f_{sw}$, then i_{DQ3} will fall to zero. Thus, the next mode will go to mode 2. On the other hand, the circuit operation will go to mode 3 if $f_r < f_{sw}$.

Mode 2 [$t_1 \sim t_2$]: At time t_1 , i_{Q3} falls to zero. Then Q_3 is forced to turn off. i_{Lr} will flow through Q_1 , C_r , L_r , L_{m2} and C_{H1} . Due to $C_{H1} > C_r$, C_r and L_{m2} are resonant in mode 2 with $v_{ab} = V_H/2$ and resonant frequency $f_p = 1/2\pi\sqrt{C_r(L_r + L_{m2})}$. Mode 2 ends at the half of the switching frequency $t = T_{sw}/2$.

Mode 3 [$t_2 \sim t_3$]: At time t_2 , Q_1 turns off and $i_{Lr}(t_2) > 0$. After time t_2 , v_{CQ1} (v_{CQ2}) is increased (decreased) because C_{Q1} (C_{Q2}) is charged (discharged) by the resonant current i_{Lr} . The zero-voltage switching condition of power device Q_2 is derived as.

$$i_{Lm2,peak} \geq \frac{V_H}{2} \sqrt{\frac{2C_{eq}}{L_{m2} + L_r}}, \quad (1)$$

where $C_{eq} = C_{Q1} = C_{Q2}$. Based on the switching frequency, magnetizing inductance and turn-ratio, the maximum magnetizing current can be calculated as:

$$i_{Lm2,peak} = \frac{\Delta i_{Lm2}}{2} \approx \frac{nV_L}{4L_{m2}f_{sw}}, \quad (2)$$

From the given dead time, switching frequency, and input and output voltages, the magnetizing inductance L_{m2} is obtained as:

$$L_{m2} \leq \frac{t_d n V_L}{8 f_{sw} V_H C_{eq}}, \quad (3)$$

Mode 4 [$t_3 \sim t_4$]: At t_3 , $v_{CQ2} = 0$. Since $i_{Lr}(t_3) > 0$, the antiparallel diode D_{Q2} is forward biased. Thus, Q_2 will turn on at zero voltage switching. Due to $i_{Q4}(t_3) < 0$, Q_4 turns on to reduce the conduction loss on Q_4 . In this mode, $v_{ab} = -V_H/2$, $v_{Lm2} = -nV_L/2$, and i_{Lm2} decreases. The current variation on L_{m2} is $\Delta i_{Lm2,4} = nV_L \Delta t_{34} / L_{m2}$ where Δt_{34} is the time duration in mode 4. L_r and C_r are resonant with $v_{ab} = -V_H/2$, $v_{Lm2} = -nV_L$ and $f_r = 1/2\pi\sqrt{C_r L_r}$. If $f_r > f_{sw}$, the next step of the circuit operation will go to mode 5. Otherwise, the circuit operation will go to mode 6.

Mode 5 [$t_4 \sim t_5$]: At time t_4 , $i_{Q4} = 0$ and synchronous switch Q_4 turns off. The resonant inductor current i_{Lr} will flow through Q_2 , C_r , L_r , L_{m2} and C_{H2} . L_r , C_r and L_{m2} are resonant with input voltage $v_{ab} = -V_H/2$.

Mode 6 [$t_5 \sim T_{sw} + t_0$]: At t_5 , power device Q_2 turns off. Due to $i_{Lr}(t_5) < 0$, capacitor C_{Q1} (C_{Q2}) is discharged (charged). The soft switching condition of power device Q_1 is the same as power device Q_2 . The charge (discharge) time of C_{Q2} (C_{Q1}) is sufficiently quick and can be ignored. At time $T_{sw} + t_0$, $v_{CQ1} = 0$ and the circuit operations of the developed converter are completed.

Figure 4. Cont.

Figure 4. Equivalent topological circuits for each mode operation under forward power transfer (a) mode 1 (b) mode 2 (c) mode 3 (d) mode 4 (e) mode 5 (f) mode 6.

Although there are several approaches [18–21] to generate square voltage waveforms on the ac side of the converter leg, fundamental frequency analysis [22] is the most useful approach to obtain voltage gain of the proposed circuit under different switching frequency. Based on the switching states of power devices, two square voltage signals are observed on high-voltage side v_{ab} and v_{Lm2} . C_r , L_r and L_{m2} work as a circuit filter to eliminate harmonic signals. Therefore, v_{ab} and v_{Lm2} can be treated as ac voltage's only fundamental frequency to simplify the circuit analysis. The fundamental root mean square (rms) voltages $v_{ab,rms}$ and $v_{Lm2,rms}$ are calculated as $\sqrt{2}V_H/\pi$ and $2\sqrt{2}V_L n/\pi$. The primary-side fundamental resistance is obtained as $R_{ac2} = 8n^2 R_{o,L}/\pi^2$ where $R_{o,L}$ is load resistance on the low-voltage side. The equivalent circuit at the fundamental switching frequency under forward power transfer is shown in Figure 3a. Based on a resonant tank consisting of C_r , L_r , L_{m2} and R_{ac2} , the voltage gain $G_{H2L}(s)$ and $|G_{H2L}(s)|$ under different switching frequency are obtained as:

$$G_{H2L}(s) = \frac{v_{Lm2,rms}(s)}{v_{ab,rms}(s)} = \frac{\frac{sL_{m2}R_{ac2}}{sL_{m2} + R_{ac2}}}{sL_r + \frac{1}{\frac{s}{C_r} + \frac{sL_{m2}R_{ac2}}{sL_{m2} + R_{ac2}}}}, \quad (4)$$

$$|G_{H2L}(F)| = \frac{K_1 F^2}{\sqrt{[(K_1 + 1)F^2 - 1]^2 + [Q_1 K_1 F(F^2 - 1)]^2}}, \quad (5)$$

where $Q_1 = \sqrt{L_r/C_r}/R_{ac2}$, $K_1 = L_{m2}/L_r$, $F = f_{sw}/f_r$ and $f_r = 1/(2\pi\sqrt{L_r C_r})$. Based on (5), the switching frequency is derived with the given high voltage value V_H , low voltage value V_L , inductor ratio $K_1 = L_{m2}/L_r$ and equivalent load resistance R_{ac2} .

3.2. Reverse Power Flow

When the proposed converter is worked as backward power transfer to transfer power from V_L to V_H (Figure 3b), power devices Q_3 and Q_4 on the low-voltage side work as active switches to control primary-side voltage V_H . For having the circuit characteristics of LLC resonant converter, ac switch S is ON so that L_{m1} is connected between points a and b . Based on the amplitude of $|i_{Lr}|$ and $|i_{Lm1}|$, the anti-parallel diodes D_{Q1} and D_{Q2} are ON or OFF and the quasi-square voltage waveform is generated on voltage v_{ab} . Likewise, a square waveform is also generated on the primary-side voltage v_{Lm2} based on the ON/OFF state of Q_3 and Q_4 . Under reverse power transfer operation, L_r , C_r and L_{m1} are resonant and the input impedance from the low-voltage side should be an inductive load. Then, Q_3 and Q_4 can be operated under zero-voltage condition such that the switching loss is lessened. The key pulse-width modulation signals under backward power transfer operation are demonstrated in Figure 3b. The equivalent topological circuits corresponding to mode operations are demonstrated in Figure 5.

Mode 1 [$t_0 \sim t_1$]: The voltage on C_{Q3} is decreased to zero voltage at time t_0 . Since $i_{Q3} < 0$ and $i_{Lr} > 0$, the body diode D_{Q3} of MOSFET Q_3 conducts so that MOSFET Q_3 turns on after t_0 to realize soft switching. In this mode, $i_{Lr}(t_0) + i_{Lm1}(t_0) < 0$, D_{Q1} is conducting, C_{H1} is charged by i_{DQ1} , $v_{ab} = V_{CH1} = V_H/2$, $v_{Lm2} = nV_L$, i_{Lm2} increases with the slope nV_L/L_{m2} and i_{Lm1} increases with the slope $V_H/(2L_{m1})$. L_r and C_r are resonant. Power transfer is from V_L to V_H in this mode through components Q_3 , T , L_r , C_r and D_{Q1} . If $f_{sw} < f_r$, the circuit action will go to mode 2. However, the circuit action will go to mode 3.

Figure 5. Equivalent topological circuits at reverse power transfer (a) mode 1 (b) mode 2 (c) mode 3 (d) mode 4 (e) mode 5 (f) mode 6.

Mode 2 [$t_1 \sim t_2$]: If $i_{DQ1}(t_1) = 0$, then diode D_{Q1} is off. The resonant current i_{Lr} flows through components S , C_r , L_r , T and L_{m2} , the components C_r , L_r and L_{m1} are resonant with $v_{Lm2} = nV_L$, and the resonant frequency is $f_p = 1/2\pi\sqrt{C_r(L_r + L_{m1})} < f_r$.

Mode 3 [$t_2 \sim t_3$]: At t_2 , power device Q_3 turns off. Since $i_{Q3} > 0$ and $i_{Q4} < 0$, C_{Q3} is charged and C_{Q4} is discharged. The zero-voltage switching condition of Q_4 is obtained as:

$$(L_{m1} + L_r)i_{Lm1,peak}^2 + L_{m2}i_{Lm2,peak}^2 \geq 2C_{Q,s}V_L^2, \quad (6)$$

where $C_{Q,s} = C_{Q3} = C_{Q4}$, $i_{Lm2,peak} \approx nV_L/(4L_{m2}f_{sw})$ and $i_{Lm1,peak} \approx V_H/(8L_{m1}f_{sw})$. The time duration in this mode is calculated when v_{CQ4} is decreased to zero voltage.

$$\Delta t_{23} \approx \frac{2V_L C_{Q,s}}{n[i_{Lm1,peak} + i_{Lm2,peak}]} = \frac{16L_{m1}L_{m2}f_{sw}V_L C_{Q,s}}{n(L_{m2}V_H + 2nL_{m1}V_L)} \leq t_d, \quad (7)$$

where t_d is a dead time.

Mode 4 [$t_3 \sim t_4$]: At time t_3 , $v_{CQ4} = 0$. Then, D_{Q4} becomes forward biased. Since $i_{Q4}(t_3)$ is negative, Q_4 can be turned on at this instant without turn-on switching loss. In this mode, D_{Q2} is forward biased, Q_4 is turned on, $v_{ab} = v_{Lm1} = -V_H/2$, $v_{Lm2} = -nV_L$, and C_r and L_r are resonant. If $f_{sw} < f_r$, the next operating mode will go to mode 5. Otherwise, the next operating mode is mode 6.

Mode 5 [$t_4 \sim t_5$]: If $f_{sw} < f_r$, $i_{DQ2} = 0$ before active device Q_4 turns off. At time t_4 , $i_{DQ2} = 0$ and D_{Q2} is reverse biased. Therefore, the resonant inductor current i_{Lr} flows through L_{m1} , C_r , L_r and T . L_{m1} , C_r and L_r are resonant under $v_{Lm2} = -nV_L$.

Mode 6 [$t_5 \sim T_{sw} + t_0$]: Q_4 is turned off at t_5 . Then, C_{Q3} and C_{Q4} are discharged and charged, respectively. At time $T_{sw} + t_0$, $v_{CQ3} = 0$. Then the circuit operation will go to mode 1 for next switching period.

The operation of the proposed converter at reverse power transfer is like the forward power transfer. Q_3 and Q_4 are controlled by frequency-mode operation. The resonant tank included L_r , C_r and L_{m1} and is operated as a band-pass filter to eliminate harmonics. In Figure 3b, the input fundamental *rms* voltage $v_{Lm2,rms} = 2\sqrt{2}nV_L/\pi$. The resonant tank includes R_{ac1} , L_{m1} , C_r and L_r . The equivalent resistance $R_{ac1} = 2R_{o,H}/\pi^2$. The voltage gains $G_{L2H}(s)$ and $|G_{L2H}(s)|$ under reverse power transfer are calculated as:

$$G_{L2H}(s) = \frac{v_{Lm1,rms}(s)}{v_{Lm2,rms}(s)} = \frac{\frac{sL_{m1}R_{ac1}}{sL_{m1}+R_{ac1}}}{\frac{1}{sC_r} + sL_r + \frac{sL_{m1}R_{ac1}}{sL_{m1}+R_{ac1}}}, \quad (8)$$

$$|G_{L2H}(F)| = \frac{K_2 F^2}{\sqrt{[(K_2 + 1)F^2 - 1]^2 + [Q_2 K_2 F(F^2 - 1)]^2}}, \quad (9)$$

where $f_r = 1/(2\pi\sqrt{L_r C_r})$, $Q_2 = \sqrt{L_r/C_r}/R_{ac1}$, $K_2 = L_{m1}/L_r$ and $F = f_{sw}/f_r$. Based on Equation (9), the necessary switching frequency f_{sw} is calculated from the given parameters V_H , V_L , $K_2 = L_{m1}/L_r$ and R_{ac1} .

4. Experimental Waveforms

In this section, a design procedure of the studied circuit is discussed and the experiments are demonstrated. For forward power operation, the high-side voltage $V_H = 350$ V ~ 400 V and the low-side voltage $V_L = 48$ V. For reverse power transfer, the low-voltage V_L varies from 38 V to 52 V and $V_H = 400$ V. The power rating $P_{o, rated} = 480$ W. Based on Equations (5) and (9), the voltage transfer functions for both forward power operation and reverse power operation are alike. To simplify the circuit design, only the forward power transfer is discussed. The voltage gain $|G_{H2L}(s)|$ under $V_H = 400$ V is set to one. Then, the turn-ratio n of power transformer T is calculated as $n = |G_{H2L}| \times \frac{V_{H,max}}{2V_{L,max}} = 1 \times \frac{400}{2 \times 52} \approx 3.85$. In this prototype, the selected primary turns and secondary turns are 23 and 6, respectively. According to turn-ratio n , the dc gains are derived under 48 V output voltage.

$$|G_{H2L}|_{dc,min} = \frac{2nV_{L,nom}}{V_{H,max}} = \frac{2 \times (23/6) \times 48}{400} \approx 0.92, \quad (10)$$

$$|G_{H2L}|_{dc,max} = \frac{2nV_{L,nom}}{V_{H,min}} = \frac{2 \times (23/6) \times 48}{350} \approx 1.05, \quad (11)$$

Since the maximum gain of $|G_{H2L}(s)|$ is 1.05, the select inductor ratio K_1 and quality factor Q_1 should be selected under the full load and minimum input voltage case. The larger K_1 can reduce circulating current loss and the smaller K_1 can increase voltage gain. Since the maximum voltage gain of $|G_{H2L}(s)|$ is 1.05, $K_1 = 13$ is selected in the prototype to lessen the circulating current loss. The maximum gain of $|G_{H2L}(s)|$ under $K_1 = 13$ and $Q_1 = 0.35$ is close to 1.18. Since $n = 23/6$, R_{ac2} at rated power can be obtained as:

$$R_{ac2} = \frac{8n^2}{\pi^2} R_{o,L} = \frac{8 \times (23/6)^2}{3.14159^2} \times \frac{48}{10} \approx 57.17 \Omega, \quad (12)$$

Since f_r is selected as 100 kHz, C_r and L_r are calculated as $C_r = 1/2\pi Q_1 f_r R_{ac2} \approx 79.5$ nF and $L_r = 1/(2\pi f_r)^2 C_r \approx 31.86$ μ H. The selected components are $C_r = 78$ nF, $L_r = 32$ μ H and $L_{m2} = K_1 L_r = 13 \times 32 = 416$ μ H. The secondary-side *rms* current reflected to the primary-side is obtained as $I_{rms,p} = \frac{\pi I_o}{2\sqrt{2}n} \approx 2.9$ A. The minimum switching frequency of the prototype circuit is obtained as $f_{sw,min} = 1/2\pi\sqrt{C_r(L_r + L_{m2})} \approx 27$ kHz. The maximum *rms* magnetizing current occurs at minimum switching frequency $f_{sw,min}$ and can be expressed as $I_{Lm2,rms} = \frac{1}{2\sqrt{3}} \frac{nV_L}{4f_{sw,min}L_{m2}} \approx 1.18$ A. The *rms* current of L_r is obtained as $I_{Lr,rms} = \sqrt{I_{Lm2,rms}^2 + I_{rms,p}^2} \approx 3.13$ A. The maximum peak

voltage on capacitor C_r is derived as $v_{Cr,peak} = \frac{\sqrt{2}I_{Lr,rms}}{2\pi f_{sw,min}C_r} \approx 334.5V$. The theoretical voltage ratings of $Q_1 \sim Q_4$ are $v_{Q1,stress} = v_{Q2,stress} = 400V$ and $v_{Q3,stress} = v_{Q4,stress} = 104V$. The *rms* switch currents are $I_{Q1,rms} = I_{Q2,rms} = I_{Lr,rms}/\sqrt{2} \approx 2.21A$ and $I_{Q3,rms} = I_{Q4,rms} = nI_{sec,p}/\sqrt{2} \approx 7.85A$. Power devices Q_1 and Q_2 are implemented using IRG4PC40W with 600 V/20 A rating, and power devices Q_3 and Q_4 are implemented using IRFB4321PbF with 150 V/85 A rating. Two SiHG20N50C with 500 V/20 A rating with back-to-back connection is adopted to implement ac switch S . The selected inductor $L_{m1} = 224 \mu H$ means the inductor ratio $K_2 = L_{m1}/L_r = 7$ at reverse power transfer condition. In the prototype circuit, the input and output capacitances are $C_{H1} = C_{H2} = 180 \mu F/400V$ and $C_L = 2200 \mu F/100V$.

Based on the circuit values calculated in design procedure, a prototype circuit was implemented, and the experiments are provided to verify the effectiveness of the developed circuit. Figure 6 gives the photograph of the proposed prototype. The test results of the developed converter operated at forward power transfer are shown in Figures 7–10. Likewise, Figures 11–14 show the test results under backward power transfer. Figure 7 gives the test results of the developed circuit under 100% load and $V_H = 350V$ case. Since f_{sw} is less than f_r , the secondary-side currents are decreased to zero before active device is turned off. Similarly, Figure 8 illustrates the experimental results under 100% rated power and $V_H = 400V$ case. It is clear that f_{sw} (switching frequency) is close to f_r (resonant frequency) so that i_{Lr} and v_{Cr} are close to sinusoidal waveforms. Figures 9 and 10 demonstrate the test waveforms of the power devices Q_1 and Q_2 under different input voltage and load situations. It is observed that zero-voltage turn-on switching of Q_1 and Q_2 are achieved from 20 to 100% of rated power. The switching frequency at $V_H = 350V$ case is less than the switching frequency at $V_H = 400V$ case. For forward power transfer, the circuit efficiencies are 94.73% (90%), 94.8% (93.2%) and 93.9% (92.3%) at 20%, 50% and 100% rated power, respectively, under 400 V (350 V) input. The measured switching frequencies are 151.3 kHz (67.4 kHz), 114.3 kHz (60 kHz) and 97.6 kHz (52.85 kHz) at 20%, 50% and 100% rated power, respectively, under 400 V (350 V) input. Figures 11 and 12 give the test results of main voltage and current waveforms under backward power transfer at input voltage $V_L = 38V$ and 52 V cases. Based on test results, the switching frequencies f_{sw} under these cases are all less than f_r (resonant frequency). Therefore, the reverse recovery current loss on D_{Q1} and D_{Q2} is improved. Figures 13 and 14 demonstrate the measured waveforms of power devices Q_3 and Q_4 at different load and voltage cases. The soft switching characteristics of Q_3 and Q_4 are realized from 20% of rated power. Based on test results shown in Figures 7–14, the experimental waveforms are agreed with the theoretical waveforms in the system analysis. For backward power transfer, the circuit efficiencies are 88.2% (87.1%), 91.1% (90.4%) and 90.2% (90.1%) at 20%, 50% and 100% rated power, respectively, under 52V (38V) input. The measured switching frequencies are 72.5 kHz (62.8 kHz), 60.2 kHz (53.4 kHz) and 53.5 kHz (48.8 kHz) at 20%, 50% and 100% rated power, respectively, under 52 V (38 V) input. Because of low voltage and high current input at the low voltage side under backward power transfer, more conduction loss will be introduced on power switches at the low voltage side. Thus, the circuit efficiency under backward power operation is less than the circuit efficiency under forward power operation.

Figure 6. Photograph of the proposed prototype.

Figure 7. Test results under forward power transfer and full load with $V_H = 350$ V (a) $v_{Q1,g}$, $v_{Q2,g}$ and i_{Lr} [$v_{Q1,g}$, $v_{Q2,g}$: 10 V/div; i_{Lr} : 5 A/div; time: 4 μ s/div] (b) $v_{Q1,g}$, v_{Cr} and v_{ab} [$v_{Q1,g}$: 10 V/div; v_{Cr} , v_{ab} : 200 V/div; time: 4 μ s/div] (c) $v_{Q1,g}$, $v_{Q2,g}$, $-i_{Q3}$ and $-i_{Q4}$ [$v_{Q1,g}$, $v_{Q2,g}$: 10V/div; $-i_{Q3}$, $-i_{Q4,g}$: 25 A/div; time: 4 μ s/div].

Figure 8. Test results under forward power transfer and full load with $V_H = 400\text{V}$ (a) $v_{Q1,g}$, $v_{Q2,g}$ and i_{Lr} [$v_{Q1,g}$, $v_{Q2,g}$: 10 V/div; i_{Lr} : 5 A/div; time: 4 μs /div] (b) $v_{Q1,g}$, v_{Cr} and v_{ab} [$v_{Q1,g}$: 10 V/div; v_{Cr} , v_{ab} : 200 V/div; time: 4 μs /div] (c) $v_{Q1,g}$, $v_{Q2,g}$, $-i_{Q3}$ and $-i_{Q4}$ [$v_{Q1,g}$, $v_{Q2,g}$: 10 V/div; $-i_{Q3}$, $-i_{Q4}$: 25 A/div; time: 4 μs /div].

Figure 9. Test results of Q_1 under (a) $V_H = 350\text{V}$ and 20% load [$v_{Q1,g}$: 10 V/div; $v_{Q1,d}$: 200 V/div; i_{Q1} : 2 A/div; time: 4 μs /div] (b) $V_H = 350\text{V}$ and 100% load [$v_{Q1,g}$: 10 V/div; $v_{Q1,d}$: 200 V/div; i_{Q1} : 5 A/div; time: 4 μs /div] (c) $V_H = 400\text{V}$ and 20% load [$v_{Q1,g}$: 10 V/div; $v_{Q1,d}$: 200 V/div; i_{Q1} : 2 A/div; time: 4 μs /div] (d) $V_H = 400\text{V}$ and 100% load [$v_{Q1,g}$: 10 V/div; $v_{Q1,d}$: 200 V/div; i_{Q1} : 5 A/div; time: 4 μs /div].

Figure 10. Test results of Q_2 under (a) $V_H = 350\text{V}$ and 20% load [$v_{Q2,g}$: 10 V/div; $v_{Q2,d}$: 200 V/div; i_{Q2} : 2 A/div; time: 4 μs /div] (b) $V_H = 350\text{V}$ and 100% load [$v_{Q2,g}$: 10 V/div; $v_{Q2,d}$: 200 V/div; i_{Q2} : 5 A/div; time: 4 μs /div] (c) $V_H = 400\text{V}$ and 20% load [$v_{Q2,g}$: 10 V/div; $v_{Q2,d}$: 200 V/div; i_{Q2} : 2 A/div; time: 4 μs /div] (d) $V_H = 400\text{V}$ and 100% load [$v_{Q2,g}$: 10 V/div; $v_{Q2,d}$: 200 V/div; i_{Q2} : 5 A/div; time: 4 μs /div].

Figure 11. Test results under backward power transfer and full load with $V_L = 38\text{V}$ (a) $v_{Q3,g}$, $v_{Q4,g}$ and i_{Lm1} [$v_{Q3,g}$, $v_{Q4,g}$: 10 V/div; i_{Lm1} : 5 A/div; time: 4 μs /div] (b) $v_{Q3,g}$, V_H , v_{Cr} and $-i_{Lr}$ [$v_{Q3,g}$: 10 V/div; V_H , v_{Cr} : 500 V/div; $-i_{Lr}$: 10 A/div; time: 4 μs /div] (c) $v_{Q3,g}$, i_{DQ1} and i_{DQ2} [$v_{Q3,g}$: 10 V/div; i_{DQ1} , i_{DQ2} : 5 A/div; time: 4 μs /div].

Figure 12. Test results under backward power transfer and full load with $V_L = 52$ V (a) $v_{Q3,g}$, $v_{Q4,g}$ and i_{Lm1} [$v_{Q3,g}$, $v_{Q4,g}$: 10 V/div; i_{Lm1} : 5 A/div; time: 4 μ s/div] (b) $v_{Q3,g}$, V_H , v_{Cr} and $-i_{Lr}$ [$v_{Q3,g}$: 10 V/div; V_H , v_{Cr} : 500 V/div; $-i_{Lr}$: 10 A/div; time: 4 μ s/div] (c) $v_{Q3,g}$, i_{DQ1} and i_{DQ2} [$v_{Q3,g}$: 10 V/div; i_{DQ1} , i_{DQ2} : 5 A/div; time: 4 μ s/div].

Figure 13. Test results of the power device Q_3 under backward power transfer (a) $V_L = 38$ V and 20% load (b) $V_L = 380$ V and 100% load (c) $V_L = 52$ V and 20% load (d) $V_L = 52$ V and 100% load [$v_{Q3,g}$: 10 V/div; $v_{Q3,d}$: 100 V; i_{Q3} : 25 A/div; time: 4 μ s/div].

Figure 14. Test results of the power device Q_4 under backward power transfer (a) $V_L = 38$ V and 20% load (b) $V_L = 38$ V and 100% load (c) $V_L = 52$ V and 20% load (d) $V_L = 52$ V and 100% load [$v_{Q4,d}$: 10 V/div; $v_{Q4,g}$: 100 V; i_{Q4} : 25 A/div; time: 4 μ s/div].

5. Conclusions

A bidirectional resonant converter is studied and implemented to achieve bidirectional power transfer capability. To achieve the dual symmetric resonant behavior for both forward power operation and backward power operation and reduce the circulating current loss under forward power transfer, an ac switch and a parallel inductor is used on the high voltage side. The input impedance of the proposed circuit is regulated with the frequency-modulation scheme. The soft switching of active devices is realized. The forward and backward power transfers have the same LLC circuit characteristics. The parallel inductor on the primary-side is disconnected under forward power transfer in order to improve the circulating current loss. Finally, experimental waveforms are provided to demonstrate the achievability of the developed converter. The main drawback of the proposed converter is more conduction loss on magnetizing inductor during the backward power transfer, because the low-side voltage will introduce a magnetizing current, and this is an additional power loss on low-side power devices and transformers. Therefore, the circuit efficiency at the backward power transfer is less than forward power transfer. In order to overcome low circuit efficiency at backward power flow in the proposed converter, the symmetric resonant circuit, such as CLLC or CLLC, instead of using half-bridge or full-bridge circuit topology, may be researched and studied in future work.

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