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Bidirectional Resonant Converter with Half-Bridge Circuits: Analysis, Design, and Implementation

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Abstract: A soft-switching dc–dc converter with bidirectional power flow capability is studied in this paper for energy storage units or electric vehicle applications. The circuit topology of the studied converter is constructed using a dual half-bridge circuit with split capacitors. A series resonant tank with frequency control modulation is employed in the proposed circuit to realize the soft switching characteristics of active devices. Synchronous-rectifier-based Metal-Oxide-Semiconductor Field-Effect Transistor (MOSFETs) are used on the low-voltage side to lessen conduction loss on the power semiconductors. The benefits of the studied circuit are bidirectional power flow capability, high efficiency, low switching loss, and simple circuit structure. The achievability of the developed converter is demonstrated from test results with a scaled-down laboratory prototype.

Keywords: resonant circuit; bidirectional power flow; zero-voltage switching; half-bridge circuit

1. Introduction

Clean energy sources have been developed to lessen the demand of fossil fuel, reduce environmental air pollution, and decrease climate change. Therefore, power-electronics-based energy conversion has been researched and studied to increase energy conversion efficiency and reduce circuit size. For ac or dc microgrid systems [1–5], dc/dc converters, ac/dc converters, and dc/ac converters are needed to integrate the utility power, wind power, solar power, energy storage units, and residential loads. For electric vehicles [6–9], ac/dc converters are employed for the battery charger, dc/dc converters are adopted on the dc power distributing system, and dc/ac converters are used for ac motor drives. Bidirectional dc/dc converters [10–15] are widely used for battery-based energy storage units in dc microgrid systems and electric vehicles. Normally, bidirectional dc/dc converters are adopted to transfer electric energy between high voltage dc buses and energy storage units. Among the circuit topologies of bidirectional dc/dc converters, the half-bridge topologies and full-bridge topologies are most attractive solutions to realize bidirectional power flow capability. Phase-shift pulse-width modulation (PWM) is usually selected in full-bridge converters. However, high circulating current losses and high switching losses on light load are the main shortcomings. Resonant converters [16–19] with frequency modulation have been studied to achieve soft switching characteristics and bidirectional power flow capability. However, the circuit characteristics under forward and reverse power flow directions are different. Thus, the design procedures of bidirectional resonant converters are more difficult. In [16], the characteristic of a series resonant converter was achieved in forward power flow (or buck operation). However, the characteristic of a series resonant converter cannot be achieved at the whole load range in reverse power flow (or boost operation) due to the circuit topologies in the buck and boost operations not being symmetric circuits. In [17–19], the circuit topologies with half-bridge or full-bridge converters in both power flows are symmetric so that the power devices can be turned on under zero voltage and low switching losses. However,

one more inductor is connected to the switching leg so that high circulating current is introduced and conduction losses are increased. Therefore, the circuit efficiency is reduced.

A bidirectional dc/dc converter with a frequency modulation scheme is studied to realize the features of bidirectional power flow capability and zero-voltage switching. A dual half-bridge circuit is used in the studied circuit to realize bidirectional power flow. A series resonant tank is adopted on the high-voltage side to realize the soft switching characteristics for all active devices. Synchronous rectifiers are used on the low-voltage side to further reduce the conduction loss. In order to realize the same resonant characteristics under forward and reverse power flow directions, an ac power switch and parallel inductor are connected to the half-bridge leg on the high-voltage side. The proposed circuit and the conventional bidirectional resonant converters in [17–19] have the same resonant component counts. However, the proposed converter has less circulating current loss due to the fact that the parallel inductor is disconnected under forward power flow. The circuit topology of the studied circuit is simple and easy to implement. In Section 2, the circuit configuration, operating stages, and circuit characteristics of the proposed converter are discussed. In Section 3, the design procedures of the developed converter are provided. Experimental waveforms are demonstrated in Section 4, and conclusions are discussed in Section 5.

2. Proposed Converter

The circuit schematic of the studied dual half-bridge converter is provided in Figure 1 to realize bidirectional power flow. Two half-bridge circuits are used on both the high- and low-voltage sides and one resonant tank is adopted on the high-voltage side to achieve the soft switching characteristic under frequency control. If power is transferred from high-voltage V_1 to low-voltage V_2 as shown in Figure 2a, the ac switch S is open and the dual half-bridge converter is worked under buck mode operation. Q_1 and Q_2 are main power devices and Q_3 and Q_4 are operated as synchronous rectifiers. L_r , C_r , and L_{m1} are operated as an LLC resonant converter to have low switching losses and high circuit efficiency functions. If power is transferred from low-voltage V_2 to high-voltage V_1 as shown in Figure 2b, the ac switch S is closed and the dual half-bridge converter is worked under boost mode operation. Under this condition, Q_3 and Q_4 are the main power devices, Q_1 and Q_2 are off, and the antiparallel diodes are worked as the full-wave rectifier. L_r , C_r , and L_{m2} are operated as an LLC resonant converter to give low switching losses and high circuit efficiency functions.

Figure 1. Circuit schematic of the proposed dual half-bridge converter.

Figure 2. Circuit operation: (a) power flow from V_1 to V_2 ; (b) power flow from V_2 to V_1 .

2.1. Forward Power Flow

The studied dual half-bridge converter can achieve bidirectional power flow capability. When power is transferred from high-voltage V_1 to low-voltage V_2 , battery banks or super capacitors can be adopted as energy storage units on the low-voltage terminal. The circuit configuration of the adopted dual half-bridge converter under forward power flow is shown in Figure 2a. The ac switch S is off under forward power flow. Q_1 and Q_2 are worked as the main switches to generate a square wave voltage on voltage v_{ab} . Q_3 and Q_4 are operated as synchronous rectifiers to lessen conducting losses on the low-voltage side. According to the polarity of the secondary side current, a square wave voltage is generated on voltage $v_{cd'}$. L_{m1} is the magnetizing inductance of the transformer; L_r and C_r are the resonant inductor and capacitor. Q_1 and Q_2 are controlled by using frequency modulation to regulate load current I_2 or voltage V_2 . Due to the resonant operation by L_r , C_r , and L_{m1} , the soft switching operation of Q_1 and Q_2 can be realized. Figure 3a demonstrates the basic PWM signals of the studied converter under forward power flow operation. If the series resonant frequency is higher (or lower) than the operating frequency, there are six (or four) operation stages per switching cycle as shown in Figure 3b–g. The circuit operations of the studied circuit under forward power flow are presented in the following.

Figure 3. Pulse-width modulation (PWM) waveforms, stage operations, and gain curve under forward power flow: (a) PWM waveforms; (b) Stage 1; (c) Stage 2; (d) Stage 3; (e) Stage 4; (f) Stage 5; (g) Stage 6; (h) gain curve of the studied circuit.

Stage 1 ($t_0 \sim t_1$): The output voltage of C_{Q1} is decreased to zero at t_0 . Therefore, the soft switching characteristic of Q_1 can be achieved after t_0 due to $i_{Q1}(t_0) = i_{Lr}(t_0) < 0$ and D_{Q1} conducts. Since $i_{Lr} > i_{Lm1}$ and $i_{Q3} < 0$, Q_3 is turned on and the secondary side current flows through MOSFET Q_3 turn-on resistor R_{on} to charge capacitor C_3 . Due to low turn-on resistance R_{on} , the conduction losses on the low-voltage side can be reduced compared to using a fast recovery diode. The ac side voltages of the two half-bridge circuits are $v_{ab} = V_{C1} = V_1/2$ and $v_{cd} = V_{C3} = V_2/2$. Thus, the primary magnetizing voltage $v_{Lm1} = v_{cd} = nV_2/2$ and the magnetizing current i_{Lm1} increase with the current slope $nV_2/(2L_{m1})$. The magnetizing current variation in Stage 1 is about $\Delta i_{Lm1,1} = nV_2\Delta t_{10}/(2L_{m1})$ where Δt_{10} is the time interval in Stage 1. C_r and L_r are resonant in Stage 1 under $v_{ab} = V_1/2$, $v_{Lm1} = nV_2/2$ and $f_r = 1/2\pi\sqrt{C_r L_r}$. If $f_r > f_{sw}$, then $T_r/2 < T_{sw}/2$. Therefore, the transformer secondary current will decrease to zero before Q_1 turns off. The circuit operation goes to Stage 2. If $f_r < f_{sw}$, then $T_r/2 > T_{sw}/2$. When Q_1 turns off, the secondary winding current is still positive. Then, the circuit operation goes to Stage 3.

Stage 2 ($t_1 \sim t_2$): If $f_r > f_{sw}$, the transformer secondary current decreases to zero before Q_1 turns off. When i_{Q3} equals zero at t_1 , Q_3 turns off. i_{Lr} flows through Q_1 , C_r , L_r , L_{m1} , and C_1 . Since $C_1 \gg C_r$, components C_r , L_r , and L_{m1} are resonant under $v_{ab} = V_1/2$ and resonant frequency $f_p = 1/2\pi\sqrt{C_r(L_r + L_{m1})}$. When $t = T_{sw}/2$, switch Q_1 turns off and the circuit operation of Stage 2 is completed.

Stage 3 ($t_2 \sim t_3$): When $t = t_2 = T_{sw}/2$, Q_1 turns off, and C_{Q1} (C_{Q2}) is charged (discharged) by i_{Lr} . In order to ensure that Q_2 turns on at zero voltage, the zero-voltage condition of Q_2 is calculated using

$$i_{Lm1,peak} \geq \frac{V_1}{2} \sqrt{\frac{2C_Q}{L_{m1} + L_r}} \quad (1)$$

where $C_Q = C_{Q1} = C_{Q2}$. The maximum magnetizing current is estimated using Equation (2).

$$i_{Lm1,peak} = \frac{\Delta i_{Lm}}{2} \approx \frac{nV_2}{8L_{m1}f_{sw}} \quad (2)$$

The time interval in Stage 3 can be estimated as follows when C_{Q2} is discharged to zero voltage:

$$\Delta t_{23} = \frac{2V_1C_Q}{i_{Lr}(t_2)} \approx \frac{2V_1C_Q}{i_{Lm1,peak}} = \frac{16L_{m1}f_{sw}V_1C_Q}{nV_2} \leq t_{dead} \quad (3)$$

where t_{dead} is the dead time between Q_1 and Q_2 . Based on Equation (3), the maximum magnetizing inductance L_{m1} can be calculated using Equation (4).

$$L_{m1} \leq \frac{t_{dead}nV_2}{16f_{sw}V_1C_Q} \quad (4)$$

Stage 4 ($t_3 \sim t_4$): The voltage of C_{Q2} is decreased to zero at t_3 . The body diode D_{Q2} conducts due to $i_{Lr}(t_3) > 0$ (or $i_{Q2}(t_3) < 0$). After time t_3 , the soft switching characteristic of Q_2 can be achieved. Since the secondary side current i_{Q4} is negative, Q_4 is forced to turn on to decrease conduction loss. Then, the ac side voltage $v_{cd} = -V_2/2$, the magnetizing voltage $v_{Lm1} = -nV_2/2$, and i_{Lm1} decreases. The magnetizing current variation in Stage 2 is about $\Delta i_{Lm1,4} = nV_2\Delta t_{34}/(2L_{m1})$ where Δt_{34} is the time interval in Stage 4. In Stage 4, C_r and L_r are resonant under $v_{ab} = -V_1/2$, $v_{Lm1} = -nV_2/2$ and $f_r = 1/2\pi\sqrt{C_r L_r}$. If $f_r > f_{sw}$, the circuit operation goes to Stage 5. If $f_r < f_{sw}$, the circuit operation goes to Stage 6.

Stage 5 ($t_4 \sim t_5$): If $f_r > f_{sw}$, i_{Q4} will go to zero before Q_2 turns off. When i_{Q4} equals zero at t_4 , Q_4 is forced to turn off. In Stage 5, i_{Lr} flows through Q_2 , C_r , L_r , L_{m1} , and C_2 . Components C_r , L_r , and L_{m1} are resonant under $v_{ab} = -V_1/2$ with resonant frequency $f_p = 1/2\pi\sqrt{C_r(L_r + L_{m1})}$. When Q_2 turns off, the circuit operation goes to Stage 6.

Stage 6 ($t_5 \sim T_{sw} + t_0$): When Q_2 turns off at t_5 , C_{Q1} (C_{Q2}) is discharged (charged) by i_{Lr} (t_5) < 0 . To ensure that Q_1 turns on at zero voltage, the zero-voltage condition of Q_1 is the same as that of Q_2 . The time interval in this stage is short enough to be neglected in state–state analysis. At time $T_{sw} + t_0$, the voltage of C_{Q1} decreases to zero voltage and the circuit operation of this stage is completed.

To derive the transfer function of the studied converter under forward power flow, a fundamental frequency approach is adopted to derive the ac voltage gain. According to the on/off states of $Q_1 \sim Q_4$, the square voltage waveforms are obtained on voltages v_{ab} and v_{cd} . Since the resonant tank of L_r , C_r , and L_{m1} is worked as a band pass filter, the harmonic frequency on v_{ab} and v_{cd} can be neglected to simplify the characteristic analysis. The root mean square (rms) values of v_{ab} and v_{cd} at the switching frequency are $\sqrt{2}V_1/\pi$ and $\sqrt{2}V_2n/\pi$, respectively. The equivalent resistance at the primary side of transformer can be calculated as $R_{ac2} = 2n^2R_o/\pi^2$. The equivalent circuit of the adopted resonant tank is shown in Figure 2a. The effectively fundamental sinusoidal voltage is $v_{ab,f}$ with rms voltage value $\sqrt{2}V_1/\pi$ and the resonant tank consists of L_r , C_r , L_{m1} , and R_{ac2} . The voltage transfer function between low-voltage V_2 and high-voltage V_1 is derived in Equation (5).

$$G_{1,2}(s) = \frac{v_{R_{ac2}}(s)}{v_{ab,f}(s)} = \frac{\frac{sL_{m1}R_{ac2}}{sL_{m1} + R_{ac2}}}{\frac{1}{sC_r} + sL_r + \frac{sL_{m1}R_{ac2}}{sL_{m1} + R_{ac2}}} \quad (5)$$

The amplitude of the voltage transfer function in Equation (5) can be obtained using Equation (6):

$$|G_{1,2}(f_{sw})| = \frac{kF^2}{\sqrt{[(k+1)F^2 - 1]^2 + [QkF(F^2 - 1)]^2}} \quad (6)$$

where $f_r = 1/(2\pi\sqrt{L_r C_r})$, $Q = \sqrt{L_r/C_r}/R_{ac2}$, $k = L_{m1}/L_r$, and $F = f_{sw}/f_r$. Based on the input voltage V_1 , output voltage V_2 , inductor ratio k , and load resistance R_{ac2} , the necessary switching frequency can be calculated from Equation (6). The gain curve of the proposed converter under forward power flow is plotted in Figure 3h under inductor ratio $k = L_{m1}/L_r = 7$.

2.2. Reverse Power Flow

The power flow of the studied dual half-bridge circuit can also be delivered from low-voltage V_2 to high-voltage V_1 as shown in Figure 2b. Under this condition, the ac switch S is turned on, Q_3 and Q_4 are operated as the main power switch to generate a square voltage waveform on voltage v_{cd} , and the primary voltage $v_{Lm1} \approx nv_{cd} = v_{cd'}$. Q_1 and Q_2 are off and the antiparallel diodes D_{Q1} and D_{Q2} are operated as the full-wave rectifiers. According to the polarity of i_{Q1} and i_{Q2} , a square wave voltage is generated on voltage v_{ab} . L_r , C_r , and L_{m2} are activated as the resonant tank to realize an inductive load operation. Therefore, the soft switching characteristic of main power devices Q_3 and Q_4 can be achieved. Q_3 and Q_4 are controlled with frequency modulation to regulate current I_1 or voltage V_1 . Figure 4 demonstrates the PWM waveforms and the equivalent circuit of the studied converter under reverse power flow.

Stage 1 ($t_0 \sim t_1$): C_{Q3} is discharged to zero voltage at t_0 , $i_{Lr} > 0$, and $i_{Q3} < 0$. The soft switching characteristic of Q_3 is realized when Q_3 is turned on after t_0 . Since $-i_{Lr} > i_{Lm2}$ and $i_{Q1} < 0$, D_{Q1} conducts and C_1 is charged by $-i_{Q1}$. In Stage 1, the ac side voltages are $v_{ab} = V_{C1} = V_1/2$ and $v_{cd} = V_{C3} = V_2/2$. i_{Lm2} increases with the current slope $V_1/(2L_{m2})$. C_r and L_r are resonant in Stage 1 under $v_{ab} = V_1/2$, $v_{Lm1} = nV_2/2$, and $f_r = 1/2\pi\sqrt{C_r L_r}$. If $f_r > f_{sw}$, the circuit operation goes to the next stage. If $f_r < f_{sw}$, the circuit operation goes to Stage 3. Power is transferred in this stage from V_2 to V_1 through Q_3 , T , L_r , C_r , and D_{Q1} .

Stage 2 ($t_1 \sim t_2$): When i_{DQ1} is decreased to zero at t_1 , D_{Q1} is reverse biased. i_{Lr} flows through S , C_r , L_r , T , and L_{m2} . C_r , L_r , and L_{m2} are resonant under $v_{Lm1} = nV_2/2$ and resonant frequency $f_p = 1/2\pi\sqrt{C_r(L_r + L_{m2})}$. When $t = t_2$, Q_3 turns off and the circuit operation of Stage 2 is completed.

Figure 4. PWM waveforms, stage operations, and gain curve under reverse power flow: (a) PWM waveforms; (b) Stage 1; (c) Stage 2; (d) Stage 3; (e) Stage 4; (f) Stage 5; (g) Stage 6; (h) gain curve of the studied circuit.

Stage 3 ($t_2 \sim t_3$): Q_3 turns off at time t_2 . Since $i_{Lr} < 0$, C_{Q3} (C_{Q4}) is charged (discharged). In order to ensure that Q_4 turns on at zero voltage, the zero-voltage condition of Q_4 is calculated in Equation (7):

$$i_{Lm2,peak} \geq \frac{V_2}{2} \sqrt{\frac{2C_{Q,s}}{L_{m2} + L_r}} \quad (7)$$

where $C_{Q,s} = C_{Q3} = C_{Q4}$. The peak values of i_{Lm1} and i_{Lm2} can be calculated as

$$i_{Lm1,peak} = \frac{\Delta i_{Lm1}}{2} \approx \frac{V_2}{8L_{m1}f_{sw}} \quad (8)$$

$$i_{Lm2,peak} = \frac{\Delta i_{Lm2}}{2} \approx \frac{V_1}{8L_{m2}f_{sw}}. \quad (9)$$

The time interval in Stage 3 can be estimated when C_{Q4} is discharged to zero voltage:

$$\Delta t_{23} = \frac{2V_2C_{Q,s}}{n[i_{Lm1}(t_2) + i_{Lm2}(t_2)]} \approx \frac{2V_2C_{Q,s}}{n[i_{Lm1,peak} + i_{Lm2,peak}]} = \frac{16L_{m1}L_{m2}f_{sw}V_2C_{Q,s}}{n(L_{m2}V_2 + L_{m1}V_1)} \leq t_{dead} \quad (10)$$

where t_{dead} is the dead time between Q_3 and Q_4 .

Stage 4 ($t_3 \sim t_4$): C_{Q4} is discharged to zero at t_3 . The body diode D_{Q4} of Q_4 conducts due to the fact that $i_{D4}(t_3) < 0$. Therefore, the soft switching characteristic of Q_4 can be achieved after time t_3 . In Stage 4, i_{Q2} is negative and D_{Q2} conducts. Therefore, $v_{ab} = -V_1/2$ and $v_{cd} = -V_2/2$. C_r and L_r are resonant with resonant frequency $f_r = 1/2\pi\sqrt{C_r L_r}$. If $f_r > f_{sw}$, the circuit operation goes to Stage 5. If $f_r < f_{sw}$, the circuit operation goes to Stage 6.

Stage 5 ($t_4 \sim t_5$): If $f_r > f_{sw}$, i_{Q2} will go to zero before Q_4 is turned off. When i_{DQ2} equals zero at t_4 , D_{Q2} is reverse biased. In Stage 5, i_{Lr} flows through C_r , L_r , T , and L_{m2} . Components C_r , L_r , and L_{m2} are resonant under $v_{cd} = -V_2/2$ with resonant frequency $f_p = 1/2\pi\sqrt{C_r(L_r + L_{m2})}$. When Q_4 turns off at t_5 , the circuit operation goes to Stage 6.

Stage 6 ($t_5 \sim T_{sw} + t_0$): When Q_4 turns off at t_5 , C_{Q3} (C_{Q4}) is discharged (charged). The time interval in Stage 6 can be neglected in state–state analysis. At time $T_{sw} + t_0$, the voltage of C_{Q3} is decreased to zero and the circuit operation of Stage 6 is completed.

The transfer function of the studied circuit under reverse power flow is similar to that under forward power flow operation. Q_3 and Q_4 are the main active devices and D_{Q1} and D_{Q2} are worked as the full-wave diode rectifier. The circuit characteristic of L_r , C_r , and L_{m2} (instead of L_{m1}) is like a band pass filter to filter the harmonic current. Figure 2b gives the equivalent circuit of the adopted resonant tank. The effectively fundamental sinusoidal voltage is $v_{cd,f}$ with rms voltage value $\sqrt{2}V_2n/\pi$, and the resonant tank consists of L_r , C_r , L_{m2} , and R_{ac1} . The equivalent resistance at the high-voltage side is calculated as $R_{ac1} = 2R_o/\pi^2$. The voltage transfer function between high-voltage V_1 and low-voltage V_2 is derived in Equation (11).

$$G_{2,1}(s) = \frac{v_{R_{ac1}}(s)}{v_{cd,f}(s)} = \frac{\frac{sL_{m2}R_{ac1}}{sL_{m2} + R_{ac1}}}{\frac{1}{sC_r} + sL_r + \frac{sL_{m2}R_{ac1}}{sL_{m2} + R_{ac1}}} \quad (11)$$

Likewise, the amplitude of the voltage transfer function in Equation (11) can be calculated using Equation (12):

$$|G_{2,1}(f_{sw})| = \frac{kF^2}{\sqrt{[(k+1)F^2 - 1]^2 + [QkF(F^2 - 1)]^2}} \quad (12)$$

where $f_r = 1/(2\pi\sqrt{L_r C_r})$, $Q = \sqrt{L_r/C_r}/R_{ac1}$, $k = L_{m2}/L_r$, and $F = f_{sw}/f_r$. Based on V_1 , V_2 , k , and R_{ac1} , the necessary switching frequency can be obtained from Equation (12). The gain curve of the proposed converter under reverse power flow is plotted in Figure 4h under inductor ratio $k = L_{m2}/L_r = 7$.

3. Design Procedure of the Studied Converter

The design procedure with a laboratory prototype is discussed and presented in this section to demonstrate the practicability of the studied bidirectional dual half-bridge converter. The voltage range at high voltage is from 350 V to 400 V and the voltage range at the low-voltage side is from 38 V to 52 V. The rated power of the prototype circuit is 480 W. When the proposed circuit is worked under forward power flow, V_2 is controlled at 48 V. If the proposed circuit is worked under reverse power flow, V_1 is controlled at 400 V. Since the circuit characteristics of the proposed converter under forward

and reverse power flow are similar, only the design procedure of the prototype circuit under forward power flow is discussed in the following. The transformer turn ratio is calculated using Equation (13) under maximum input and output voltages and $G_{1,2} = 1$.

$$n = |G_{1,2}| \times \frac{V_{1,\max}}{V_{2,\max}} = 1 \times \frac{400}{52} \approx 7.7 \quad (13)$$

The actual primary and secondary winding turns are 24 turns and 3 turns, respectively. Therefore, the minimum and maximum dc gains are determined using Equations (14) and (15), respectively, under nominal output voltage.

$$|G_{1,2}|_{dc,\min} = n \times \frac{V_{2,\text{nom}}}{V_{1,\max}} = \frac{24}{3} \times \frac{48}{400} \approx 0.96 \quad (14)$$

$$|G_{1,2}|_{dc,\max} = n \times \frac{V_{2,\text{nom}}}{V_{1,\min}} = \frac{24}{3} \times \frac{48}{350} \approx 1.097 \quad (15)$$

The select inductor ratio k and quality factor Q are based on the necessary maximum dc gain $|G_{1,2}|_{dc,\max} = 1.097$. The maximum gain of the proposed converter under the selected $k = 7$ and $Q = 0.6$ is about 1.35, which is greater than $|G_{1,2}|_{dc,\max} = 1.097$. Based on the selected turn ratio, the equivalent resistance R_{ac2} at full load is determined from Equation (16).

$$R_{ac2} = \frac{2n^2}{\pi^2} R_o = \frac{2 \times (24/3)^2}{3.14159^2} \times \frac{48}{10} \approx 62.25 \, \Omega \quad (16)$$

The resonant parameters C_r and L_r are calculated from Equations (17) and (18) under the series resonant frequency $f_r = 100$ kHz and full load.

$$C_r = \frac{1}{2\pi Q f_r R_{ac2}} = \frac{1}{2\pi \times 0.6 \times 100 \times 10^3 \times 62.25} \approx 42.6 \, \text{nF} \quad (17)$$

$$L_r = \frac{1}{(2\pi f_r)^2 C_r} = \frac{1}{(2\pi \times 100 \times 10^3)^2 \times 42.6 \times 10^{-9}} \approx 59.5 \, \mu\text{H} \quad (18)$$

The actual resonant parameters used in the prototype are $C_r = 42$ nF and $L_r = 60$ μH . The magnetizing inductance L_{m1} is determined from Equation (19).

$$L_{m1} = kL_r = 7 \times 60 = 420 \, \mu\text{H} \quad (19)$$

The primary side rms load current is calculated in Equation (20).

$$I_{rms,p} = \frac{\pi I_o}{\sqrt{2}n} = \frac{\pi \times 10}{\sqrt{2} \times (24/3)} = 2.777 \, \text{A} \quad (20)$$

The minimum switching frequency is determined in Equation (21).

$$f_{sw,\min} = f_p = \frac{1}{2\pi \sqrt{C_r(L_r + L_{m1})}} = \frac{1}{2\pi \sqrt{42 \times 10^{-9} \times (60 + 420) \times 10^{-6}}} \approx 35.45 \, \text{kHz} \quad (21)$$

The rms magnetizing current of transformer T at $f_{sw,\min}$ under minimum switching frequency is calculated in Equation (22).

$$I_{Lm1,rms} = \frac{1}{2\sqrt{3}} \frac{nV_2}{4f_{sw,\min} L_{m1}} = \frac{1}{2\sqrt{3}} \times \frac{(24/3) \times 48}{4 \times 35.45 \times 10^3 \times 420 \times 10^{-6}} \approx 1.86 \, \text{A} \quad (22)$$

The rms values of the resonant inductor current $I_{Lr,rms}$ and the primary winding current $I_{pri,T}$ under minimum switching frequency and full load are calculated and expressed in Equation (23).

$$I_{pri,T} = I_{Lr,rms} = \sqrt{I_{Lm1,rms}^2 + I_{rms,p}^2} = \sqrt{1.86^2 + 2.77^2} \approx 3.34 \text{ A} \quad (23)$$

The rms value of the secondary winding current $I_{sec,T}$ is given as

$$I_{sec,T} = n i_{rms,p} = (24/3) \times 2.777 = 22.21 \text{ A}. \quad (24)$$

The peak voltage of the resonant capacitor C_r under minimum switching frequency and full load is calculated in Equation (25).

$$v_{Cr,peak} = \sqrt{2} I_{Lr,rms} X_{Cr} = \frac{\sqrt{2} I_{Lr,rms}}{2\pi f_{sw,min} C_r} = \frac{\sqrt{2} \times 3.34}{2\pi \times 35.45 \times 10^3 \times 42 \times 10^{-9}} \approx 505 \text{ V} \quad (25)$$

The voltage stresses of Q_1 and Q_2 are equal to $V_1 = 400 \text{ V}$, and the voltage stresses of Q_3 and Q_4 are equal to $V_{2,max} = 52 \text{ V}$. The current stresses of $Q_1 \sim Q_4$ are calculated in Equations (26) and (27).

$$I_{Q1,rms} = I_{Q2,rms} = I_{Lr,rms} / \sqrt{2} = 3.34 / \sqrt{2} \approx 2.36 \text{ A} \quad (26)$$

$$I_{Q3,rms} = I_{Q4,rms} = I_{sec,T} / \sqrt{2} = 22.21 / \sqrt{2} \approx 15.7 \text{ A} \quad (27)$$

The insulated gate bipolar transistor (IGBT) IRG4PC40W with 600 V/20 A rating with less switching losses and higher switching frequency operation were adopted for switches Q_1 and Q_2 and MOSFETs IRFB4321PbF with 150 V/85 A rating were adopted for switches Q_3 and Q_4 . The ac switch S was implemented using two MOSFETs SiHG20N50C with 500 V/20 A rating with back-to-back connection. The selected inductor $L_{m2} = 180 \mu\text{H}$ gives inductor ratio $k = L_{m2}/L_r = 3$ under reverse power flow operation from V_2 to V_1 . The adopted input and output capacitances are $C_1 = C_2 = 180 \mu\text{F}/400 \text{ V}$ and $C_3 = C_4 = 2200 \mu\text{F}/100 \text{ V}$.

4. Experimental Results

Based on the circuit parameters derived from the previous section, a 480 W prototype circuit was tested to validate the practicability of the studied converter. Figures 5–8 demonstrate the test results under forward power flow operation. Figure 5 shows the test results of Q_1 at 20% load and full load. The test results show that Q_1 turns on at zero voltage for both 20% and 100% loads. Likewise, Figure 6 demonstrates test results of Q_2 under different load conditions. It is observed that the soft switching turn-on characteristics of Q_2 are also achieved. Figure 7 demonstrates the measured waveforms of the resonant voltage and resonant current. The test results show that the switching frequency is close to the series resonant frequency under 400 V input. Thus, the resonant current is close to a sinusoidal waveform. On the other hand, the series resonant frequency is greater than the switching frequency under the 350 V input case. Figure 8 demonstrates the test results of the secondary side currents and capacitor voltages. The output capacitor voltages V_{C3} and V_{C4} are balanced and the switch currents $-i_{Q3}$ and $-i_{Q4}$ are also balanced. Figures 9–11 demonstrates test results of the studied converter under reverse power flow. The soft switching characteristics of Q_3 and Q_4 are given in Figures 9 and 10 and the resonant voltage and current are demonstrated in Figure 11. Figure 12 shows the measured circuit efficiencies of the studied circuit. The efficiencies of the proposed converter under forward power flow and $V_1 = 400 \text{ V}$ are 95.2%, 95.7%, and 93.6% under 20%, 50%, and 100% loads, respectively. The efficiencies of the proposed converter under reverse power flow and $V_2 = 52 \text{ V}$ are 89.7%, 91.5%, and 91.2% under 20%, 50%, and 100% loads, respectively. The proposed converter has better circuit efficiency under forward power flow due to the fact that MOSFETs are used on the low-voltage side to reduce the conduction losses. The measured switching frequencies of the proposed converter under forward power flow are 138 kHz (or 84 kHz), 114 kHz (or 79 kHz), and 99 kHz (or 71 kHz) under

400 V (or 350 V) input and 20% load, 50% load, and 100% load output, respectively. Lower input voltage results in low switching frequency and the switching frequency is increased when the dc load is decreased. Figure 13 gives a picture of the experimental setup and a photo of the proposed converter in the laboratory test.

Figure 5. Test results of switch Q_1 under $V_1 = 400$ V with (a) 96 W (20%) output and (b) 480 W (100%) output.

Figure 6. Test results of switch Q_2 under $V_1 = 400$ V with (a) 96 W (20%) output and (b) 480 W (100%) output.

Figure 7. Test results of v_{ab} , $-v_{Cr}$, and i_{Lr} under 480 W output with (a) $V_1 = 350$ V and (b) $V_1 = 400$ V.

Figure 8. Test results of $-i_{Q3}$, $-i_{Q4}$, V_{C3} , and V_{C4} under 480 W output with (a) $V_1 = 350$ V and (b) $V_1 = 400$ V.

Figure 9. Test results of Q_3 and Q_4 under reverse power flow and 96W (20% load) output with (a) Q_3 at $V_2 = 38$ V, (b) Q_3 at $V_2 = 52$ V, (c) Q_4 at $V_2 = 38$ V, and (d) Q_4 at $V_2 = 52$ V.

Figure 10. Test results of Q_3 and Q_4 under reverse power flow and 480 W (100% load) output with (a) Q_3 at $V_2 = 38$ V, (b) Q_3 at $V_2 = 52$ V, (c) Q_4 at $V_2 = 38$ V, and (d) Q_4 at $V_2 = 52$ V.

Figure 11. Test results of $-v_{Cr}$, i_{Lr} , i_{Lm2} , i_{DQ1} , and i_{DQ2} under reverse power flow and 480 W output with (a) $V_2 = 38$ V and (b) $V_2 = 52$ V.

Figure 12. Measured circuit efficiencies of the studied converter under (a) forward power flow and (b) reverse power flow.

Figure 13. Pictures of the studied converter in the laboratory: (a) experimental setup, (b) prototype circuit.

5. Conclusions

This paper studied a resonant converter constructed using a dual half-bridge circuit to realize bidirectional power flow. Two half-bridge circuits were used on the high- and low-voltage sides to clamp the voltage rating of the active devices at input and output voltages. A series resonant tank was adopted on the primary side of the isolated transformer to realize soft switching characteristics of the power devices. A frequency control scheme was adopted to control the switching frequency so that the voltage gain of the resonant tank is adjustable and the load voltage can be controlled. A laboratory prototype of the design was built and the test results demonstrated the feasibility of the studied circuit.

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