

Article

Analysis and Implementation of a Bidirectional Converter with Soft Switching Operation

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Abstract: This paper presents a soft switching direct current (DC) converter, with the benefits of bidirectional power conversion and wide-ranging voltage operation for battery charging and discharging capability. A series resonant circuit with variable switching frequency modulation is used to achieve the advantages of soft switching turn-on or turn-off of semiconductor devices. Therefore, the switching power losses in power devices can be reduced. A symmetric resonant circuit topology with a capacitor–inductor–inductor–capacitor (CLLC) structure is adopted to achieve a bidirectional power conversion capability for battery storage units in electric vehicle applications. Due to the symmetric circuit structure on both input and output sides, the converter has similar voltage gains for each power flow operation. In order to overcome the drawback of narrow voltage range operation in conventional resonant converters, a variable transformer turns ratio is adopted in the circuit, to achieve wide output voltage operation (150–450 V) for battery charging applications. To demonstrate the converter performance, a 1-kW laboratory prototype was constructed and tested. Experimental results are provided, to verify the effectiveness of the studied circuit.

Keywords: series resonant converter; bidirectional DC converter; wide voltage gain; frequency modulation



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1. Introduction

Power converters with bidirectional power flow have been widely utilized between AC and DC grid systems and battery storage systems. Three-phase AC/DC bidirectional power factor correctors (PFC) [1,2] and DC/DC bidirectional converters [3–5] are adopted to realize forward and reverse power flow operations between AC grid and DC grid systems. The same circuit structures can be used for electric vehicle (EV) systems, to achieve vehicle-to-grid (V2G) and grid-to-vehicle (G2V) operations. The main advantages of AC/DC PFC are high power factor and low current harmonics. The bidirectional DC/DC converters can realize forward/reverse power operations between the different DC voltage buses, such as battery storage systems or EV battery units [6]. For renewable energy applications [7,8], wide voltage DC/DC and AC/DC converters are required in solar power and wind power conversion, respectively, due to unstable solar intensity and wind speed. The basic circuit topologies of a bidirectional AC/DC converter are based on four or six three-leg active switches. These AC/DC converters can be controlled in PFC operation under forward power flow, or inverter operation under reverse power flow. The buck/boost DC/DC converter is the basic topology for accomplishing bidirectional power conversion. Buck (boost) operation is controlled under forward (reverse) power flow. However, a buck/boost converter has no electric isolation.

In [9], a pulse-width modulation converter was presented, to realize a bidirectional power flow capability with voltage step-down/step-up features during forward/backward power flow. However, a PWM converter cannot achieve both voltage step up/down demands under both power flow directions. The same drawback can be found in bidirectional inductor–inductor–capacitor (LLC) resonant converters with synchronous switches on the secondary side. In references [10,11], symmetric PWM converters with two full bridge

circuit structures were studied, to accomplish bidirectional power flow with PWM modulation and phase-shift angle control between two full bridge circuits with a load voltage range $V_o = 300\text{--}450$ V. However, two control variables (phase angle and duty cycle) are needed, so the control complexity is increased. In references [12–14], resonant converters with a capacitor–inductor–inductor–capacitor (CLLC) circuit structure were studied, to realize bidirectional power operation with pulse frequency modulation for electric EV or energy storage unit applications. The main advantage of this circuit topology is the symmetric circuit structure on the primary and secondary sides, to achieve voltage step up/down operations in both power flow directions. However, the CLLC converters has less voltage gain and the output voltage range is limited, such as to $V_{in} = 400\text{--}600$ V and $V_o = 300\text{--}450$ V. For the wide voltage variation conditions in solar power and wind power applications, the power converters are normally required to overcome wide solar intensity or wind speed variations. In references [15–19], two-stage converters and PWM or resonant converters with series-parallel connection were studied, to achieve wide voltage operation. However, these circuit topologies can only achieve forward power operation.

In order to overcome the drawbacks of the above circuit topologies, a new DC converter is provided, to implement soft switching operation, a wide voltage output capability (150~450 V), and bidirectional power flow for energy storage units, bidirectional DC nano-grid systems, and battery charge/discharge systems. A CLLC resonant converter is adopted in both input and output sides, to have bidirectional power conversion capability. The load voltage control is regulated by a frequency modulation scheme. Due to the resonant circuit characteristics, the power semiconductors are all turned on or turned off under zero voltage or zero current conditions, to achieve low switching loss. To extend the output voltage operating range, one additional power switch and a variable transformer turns ratio are adopted on the primary side. Thus, the proposed converter can be operated in two operation regions, with a low (high) transformer turns ratio for high (low) voltage output operation. Therefore, the studied circuit can overcome the narrow voltage range problem in conventional CLLC bidirectional converters. A laboratory prototype (1 kW power) was implemented to demonstrate the effectiveness of the proposed bidirectional resonant converter with the features of soft switching operation, bidirectional power flow, and wide voltage output. The circuit operations of basic bidirectional converters and the proposed converter are discussed and presented in Section 2. In Section 3, the circuit features and design example are provided. In Section 4, experiments are provided from on a laboratory prototype, to verify the theoretical discussions. A conclusion regarding the presented circuit is discussed in Section 5.

2. Circuit Diagram and the Principle of Operation

Figure 1 gives the circuit schematic of the bidirectional AC/DC circuit for electric vehicle charge/discharge. First, a bidirectional AC/DC converter with PFC function is required in the front stage, to achieve bidirectional power flow operation. The rear stage is a bidirectional DC/DC converter, to charge/discharge battery stacks. Therefore, G2V and V2G power flow operations can be achieved in a bidirectional AC/DC power converter (Figure 1). Due to high circuit-efficiency requirements, boost-type circuit topologies are normally adopted in PFC converters, to obtain low current harmonics and a high power factor (PF). Figure 2 shows the general four quadrature bidirectional PFC circuit topologies with a boost-type converter. In Figure 2a, a full bridge bridgeless PFC converter is used to achieve low current harmonics and high PF. Since a boost converter is adopted, the DC voltage $V_{dc} > \sqrt{2}V_{ac,rms}$ (peak voltage of AC source). A bidirectional half bridge bridgeless PFC converter is shown in Figure 2b. Two switches are employed to accomplish low current harmonics and high PF. However, the DC voltage $V_{dc} > 2\sqrt{2}V_{ac,rms}$, when AC current i_{ac} is in phase with V_{ac} . Power flow is controlled from the AC mains to DC bus voltage V_{dc} . If the line current i_{ac} is out of phase with V_{ac} , then the power flow is from V_{dc} terminal to V_{ac} terminal.

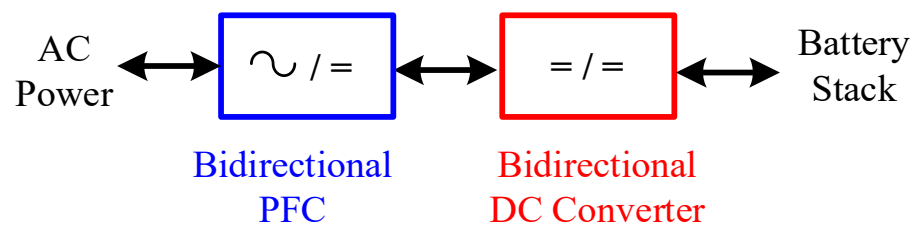


Figure 1. Bidirectional converter for battery a charger/discharger in electric vehicle applications.

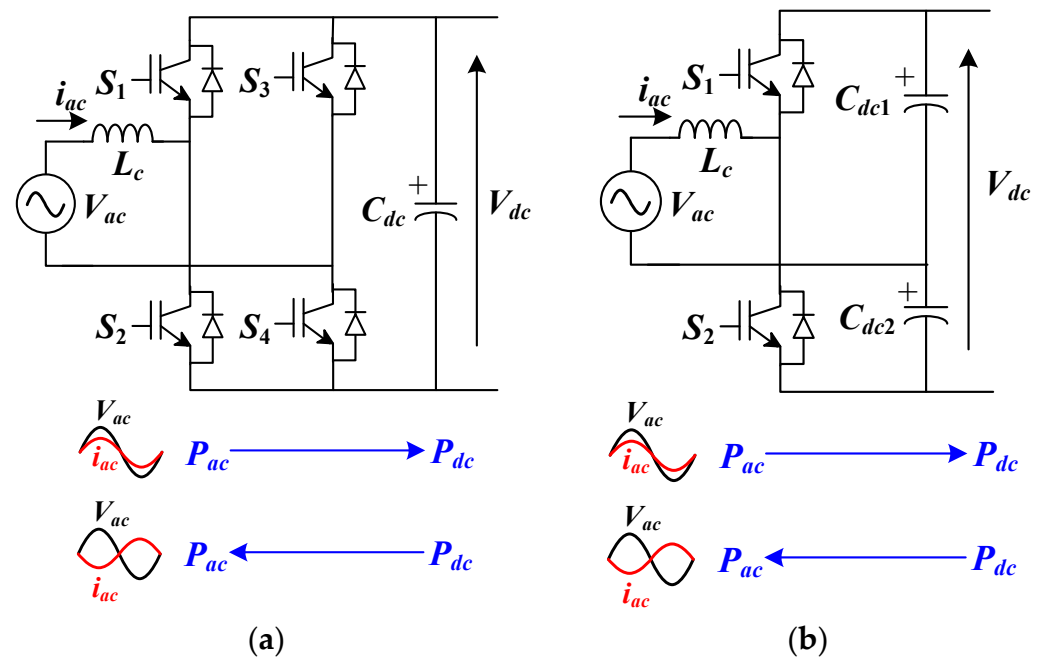


Figure 2. Circuit topologies of bidirectional single-phase bridgeless PFC converters with power factor correction: (a) full bridge structure (b) half bridge structure.

Figure 3 gives the conventional circuit topologies of DC bidirectional converters for battery a charger/discharger. The conventional non-isolate converter with bidirectional power conversion is shown in Figure 3a. For forward power conversion, the circuit functions at voltage step-down, and power flow is from the V_{in} terminal to V_{Bat} terminal. If the circuit is used for reverse power conversion, voltage step-up operation is realized in this DC converter. The simple circuit structure and control scheme are the main advantages of this circuit topology. However, the lack of electric isolation is the main drawback of this DC converter. A full bridge resonant converter is provided in Figure 3b for battery charging and discharging operations. Compared to a conventional unidirectional resonant circuit, this circuit topology can achieve voltage step-up and step-down operations for power flow control. In Figure 3c, the symmetric half bridge structure has less circuit components and also achieves forward and backward power operation. The power switches in Figure 3b,c can turn on at zero voltage, to lessen the electromagnetic interference and switching loss. However, the disadvantage of the resonant converters in Figure 3b,c is low voltage gain (narrow voltage range) if a wide voltage output range is required for a battery charging station for universal electric vehicle charging applications.

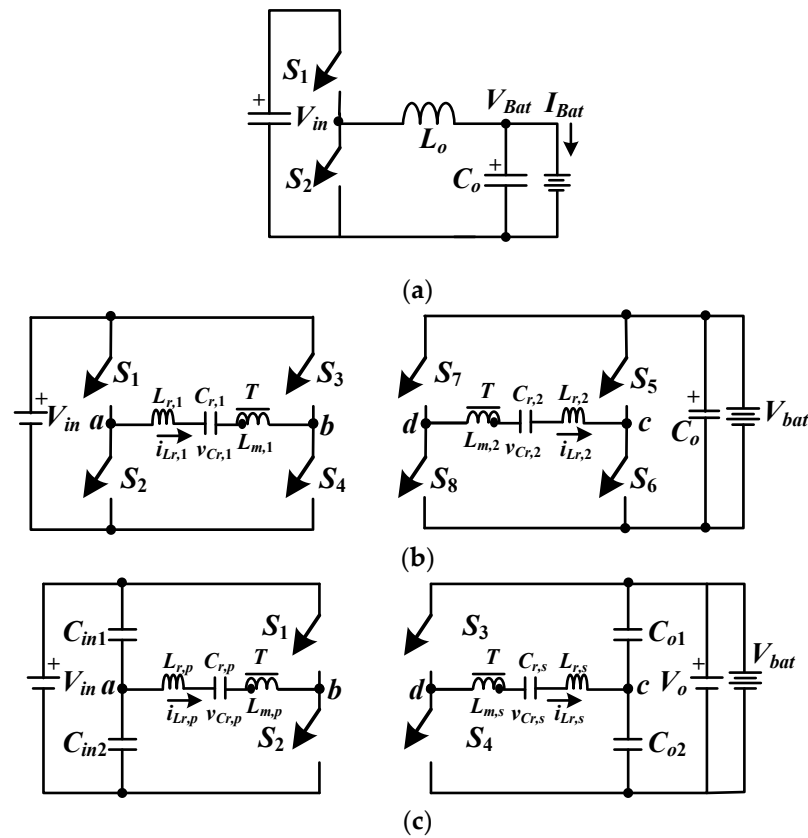


Figure 3. Circuit topologies of a bidirectional converter: (a) buck/boost converter, (b) isolated resonant converter with symmetric full bridge circuit, (c) isolated resonant converter with symmetric half bridge circuit.

Figure 4a gives the circuit structure of the proposed converter. The proposed circuit not only can achieve bidirectional power conversion, but also an extended output voltage range from 150–450 V for battery stacks. Four active switches S_1 – S_4 and one ac switch Q are used on the input side, and two switches S_5 and S_6 are employed on the output side. S_5 and S_6 are inactive when the circuit is used for forward power conversion. The forward power conversion is controlled by switches S_1 – S_4 and Q , to charge battery stacks. However, S_1 – S_4 are off when reverse power conversion is required in the proposed circuit. The reverse power is controlled by switches S_5 and S_6 , to realize battery discharging operations. The power semiconductors are all turned on or turned off at zero voltage or zero current conditions, for both power conversions. Therefore, the electromagnetic interference and switching loss can be reduced. The main drawback of a conventional bidirectional resonant converter is the limited input or output voltage operation. To implement a bidirectional converter with a wide voltage operation in universal battery charger/discharger units, two half bridge resonant circuit structures, (S_1 , S_2 , T , $C_{r,1,1}$, $L_{r,1,1}$, C_1 and C_2) and (S_3 , S_4 , T , $C_{r,1,2}$, $L_{r,1,2}$, Q , $C_{r,1,1}$, $L_{r,1,1}$, C_1 and C_2), are employed on the input side, to extend output voltage range. Figure 4b shows a circuit schematic operated for forward power conversion, with low voltage output range. The ac switch Q is on, and active switches S_1 and S_2 are turned off. Only S_3 and S_4 are activated with a frequency modulation scheme, to control the load voltage. The resonant components on the left-hand side include $C_{r,1,2}$, $L_{r,1,2}$, $C_{r,1,1}$, and $L_{r,1,1}$, and the turns ratio of transformer T in Figure 4b is $2n_p/n_s$. When forward power conversion and high voltage output range are demanded, the high voltage gain is needed in the converter. Therefore, switches Q , S_3 , and S_4 are turned off in Figure 4c. The equivalent resonant components on the left-hand side become $C_{r,1,1}$ and $L_{r,1,1}$, with the transformer turns ratio n_p/n_s , to achieve a higher voltage gain than the equivalent circuit in Figure 4b. In order to realize the same series resonant frequencies as in Figure 4b,c, the

resonant components are selected as $L_{r1,1} = L_{r1,2}$ and $C_{r1,1} = C_{r1,2}$. For battery discharging operations (Figure 4d), the reverse power flow is controlled by switches S_5 and S_6 . If the battery stack is in the full power condition, to transfer power from the V_{bat} side to V_{in} side, the ac switch Q is turned off and the half bridge diode rectifier with D_{S1} and D_{S2} is operated on the left-hand side. Therefore, the proposed converter can be controlled in three operating modes, to accomplish soft switching operation, a wide voltage output, and bidirectional power flow control. A comparison of the performance of the proposed converter and related past works in the literature is provided in Table 1. The proposed bidirectional converter has a wider voltage operation capability compared to the other circuit topologies.

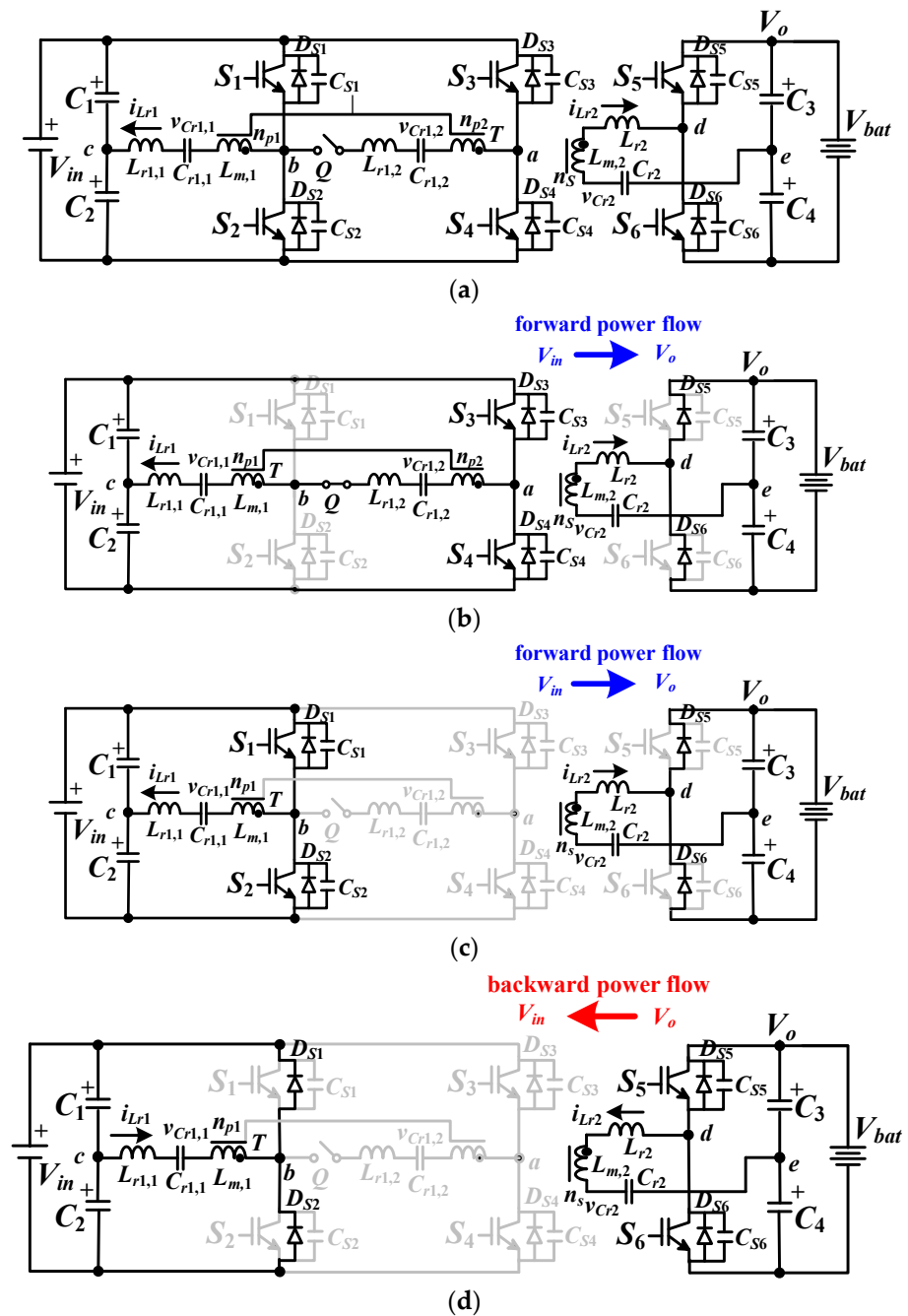


Figure 4. Proposed bidirectional converter with wide output voltage capability: (a) circuit diagram, (b) forward power conversion and low voltage output, (c) forward power conversion and high voltage output, (d) reverse power conversion.

Table 1. Comparison between the presented circuit and the other bidirectional converters.

	Primary-Side	Secondary-Side	Input/Output Voltages	Power Flow Direction	Control Scheme
Proposed circuit	Half-bridge circuit	Half-bridge circuit	$V_{in} = 400$ V $V_o = 150$ – 450 V	Voltage step-up/step-down for bidirectional power flow	Frequency control
Circuit structure in [9]	Half-bridge circuit	Half-bridge circuit	$V_{in} = 48$ V $V_o = 24$ – 30 V	Voltage step-down for forward power flow and voltage step-up for reverse power flow	Phase shift + PWM control
Circuit structure in [11]	Full-bridge circuit	Full-bridge circuit	$V_{in} = 400$ V $V_o = 400$ V	Voltage step-up/step-down for bidirectional power flow	Phase shift + PWM control
Circuit structure in [13]	Full-bridge circuit	Full-bridge circuit	$V_{in} = 382$ – 408 V $V_o = 400$ V	Voltage step-up/step-down for bidirectional power flow	Frequency control
Circuit structure in [15]	Cascade half-bridge circuit	Parallel center-tapped rectifier	$V_{in} = 750$ – 800 V $V_o = 24$ V	Voltage step-up/step-down for forward power flow	Frequency control

2.1. Forward Power Conversion and Low Voltage Output

When a battery stack has a low capacity or depleted condition, the converter is controlled in low voltage output mode (Figure 4b). Q turns on and S_1 , S_2 , S_5 , and S_6 are off under low voltage output mode. S_3 and S_4 are operated with frequency modulation, to control the battery voltage or battery charging current. The transformer turns ratio in Figure 4b is $n_1 = (n_{p1} + n_{p2})/n_s = 2n_p/n_s = 2n$, where $n_{p1} = n_{p2} = n_p$ and $n = n_p/n_s$. Figure 5a shows the main PWM signals of the proposed converter under a f_{sw} (switching frequency) $< f_r$ (resonant frequency) condition. The circuit parameters are assumed as $L_{r1,1} = L_{r1,2} = n^2 L_{r2}$ and $C_{r1,1} = C_{r1,2} = C_{r2}/n^2$. Based on the above assumption, the resonant frequencies under forward and reverse power flow operations are identical. Figure 5b–i show the operating circuits for eight operating states.

- State 1 [$t_0 \leq t < t_1$]: For $t = t_0$, the output capacitor voltage v_{CS3} is decreased to zero voltage and the diode D_{S3} becomes forward biased, owing to $i_{Lr1}(t_0)$ being negative. At this time, S_1 is turned on, to have zero voltage switching. Since i_{Lr2} is positive, the antiparallel diodes D_{S5} and D_{S6} are conducting.
- State 2 [$t_1 \leq t < t_2$]: The primary current i_{Lr1} becomes positive value after t_1 . Therefore, i_{Lr1} will flow through switch S_3 instead of diode D_{S3} . In this state, the forward power flow is from C_1 to C_3 . The leg voltages are $V_{ac} = V_{C1} = V_{in}/2$ and $V_{de} = V_{C3} = V_o/2$. The series resonant frequency is $f_r = 1/(2\pi\sqrt{(L_{r1,1} + L_{r1,2})(C_{r1,1}C_{r1,2})/(C_{r1,1} + C_{r1,2})}) = 1/(2\pi\sqrt{L_{r1,1}C_{r1,1}})$. If $f_{sw} < (or >) f_r$, then the next state operation of the proposed circuit will go to state 3 (or state 4).
- State 3 [$t_2 \leq t < t_3$]: At time t_2 , i_{Lr2} becomes zero at time t_2 and D_{S5} becomes off. i_{Lr1} flows through S_3 , $L_{m,1}$, $C_{r1,1}$, $L_{r1,1}$, $C_{r1,2}$, $L_{r1,2}$, and C_1 .
- State 4 [$t_3 \leq t < t_4$]: At time t_3 , S_3 turns off. i_{Lr1} charges C_{S3} and discharges C_{S4} . $V_{CS4} = 0$ at t_4 .
- State 5 [$t_4 \leq t < t_5$]: $v_{CS4} = 0$ at t_4 and D_{S4} becomes forward biased due to $i_{Lr1}(t_4) > 0$. At this moment, S_4 turns on under zero voltage. In state 5, $i_{Lr2} < 0$ and D_{S6} is conducting to charge C_4 .
- State 6 [$t_5 \leq t < t_6$]: Since $V_{ac} = -V_{C2} = -V_{in}/2$, i_{Lr1} is decreased and becomes negative at time t_5 . Then, the primary current i_{Lr1} will flow through S_4 instead of D_{S4} . Forward power flow is from C_2 to charge C_4 .
- State 7 [$t_6 \leq t < t_7$]: $i_{Lr2} = 0$ at $t = t_6$ and D_{S6} turns off. The primary side current i_{Lr1} flows through C_2 , $L_{r1,1}$, $C_{r1,1}$, $L_{m,1}$, $L_{r1,2}$, $C_{r1,2}$, and S_4 .
- State 8 [$t_7 \leq t < T_{sw} + t_0$]: S_4 turns off at $t = t_7$. Then, C_{S3} is discharged and C_{S4} is charged by current i_{Lr1} . This state is terminated at $T_{sw} + t_0$.

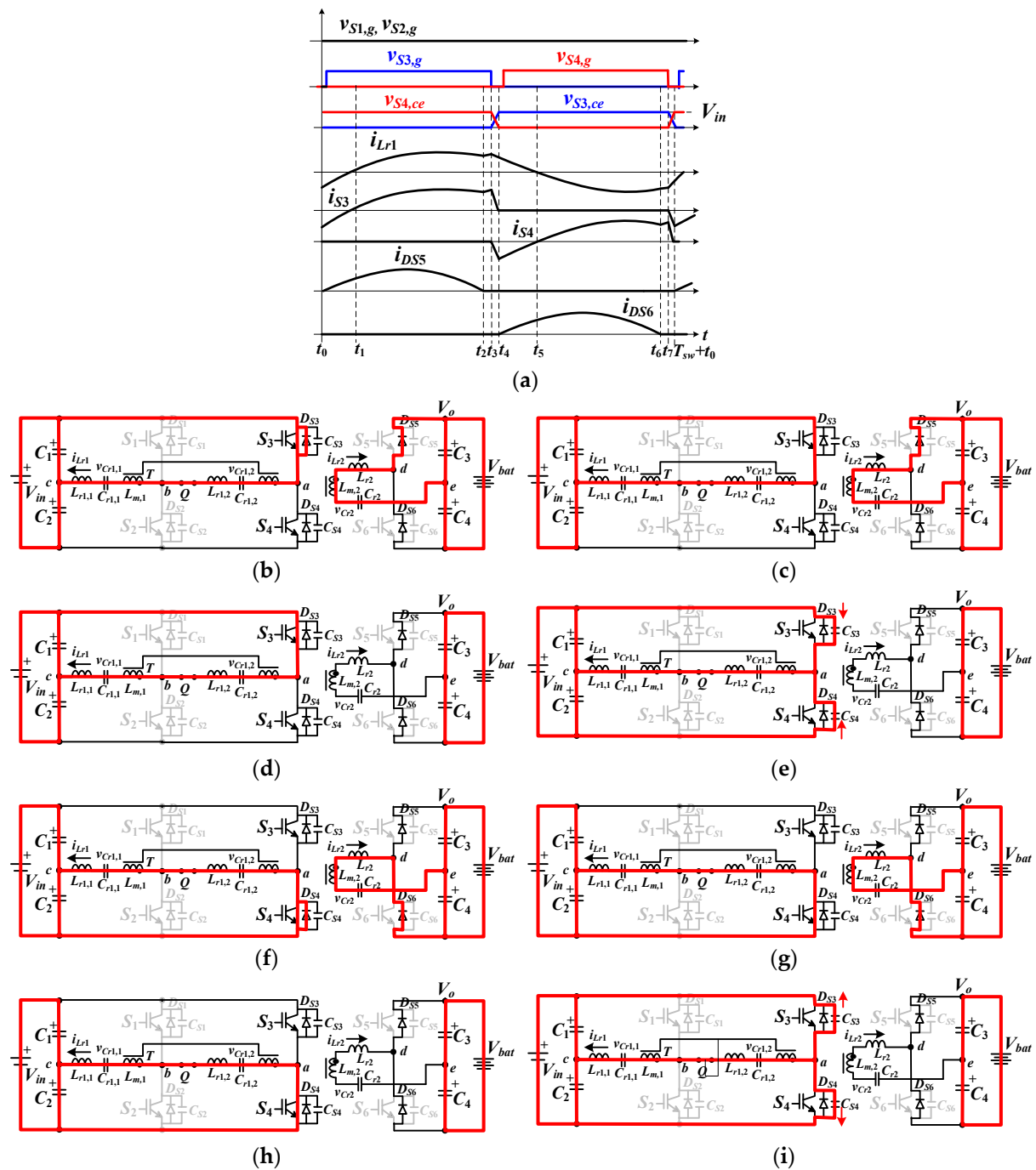


Figure 5. Circuit waveforms and state operations for low voltage output range: (a) PWM waveforms, (b) state 1, (c) state 2, (d) state 3, (e) state 4, (f) state 5, (g) state 6, (h) state 7, (i) state 8.

2.2. Forward Power Conversion and High Voltage Output

When the battery stack is in a high capacity condition, the proposed converter is operated at high voltage output mode (Figure 4c). To achieve high voltage output, switches S_3 – S_6 and Q are turned off. S_1 and S_2 are triggered by variable frequency control. In Figure 4c, the circuit has a turns ratio $n_2 = n_{p1}/n_s = n$. The key PWM waveforms and state operations in every PWM period are shown in Figure 6. Due to the similar circuit operations in Figures 5 and 6, the state operations in this mode are neglected in this section.

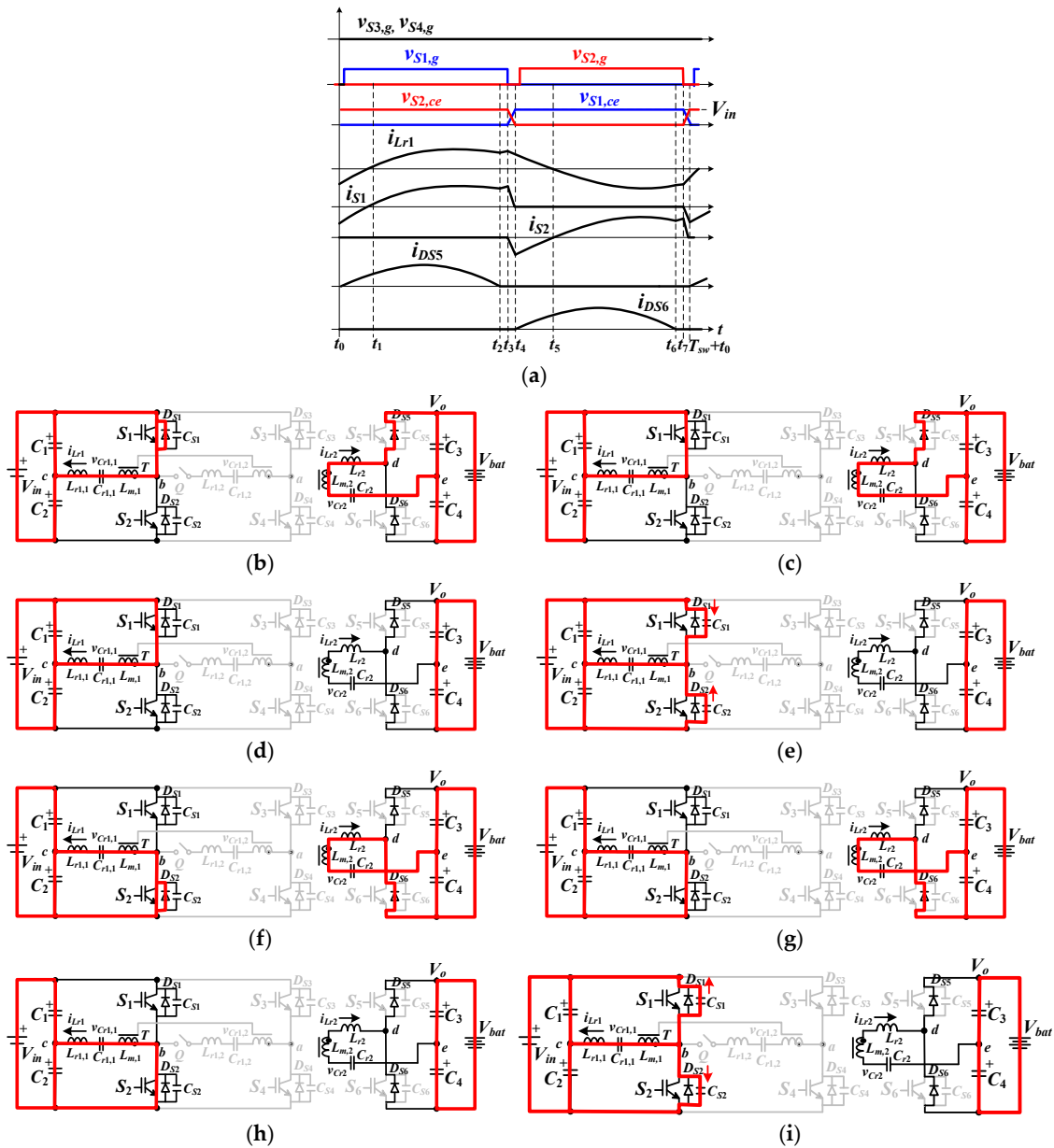


Figure 6. Circuit waveforms and state operations for high voltage output range: (a) PWM waveforms, (b) state 1, (c) state 2, (d) state 3, (e) state 4, (f) state 5, (g) state 6, (h) state 7, (i) state 8.

2.3. Reverse Power Conversion

If the battery stack is discharged, the proposed circuit is operated in reverse power conversion mode. The V_{Bat} terminal transfers power to V_{in} terminal (Figure 4d). All switches on the left-hand side of the converter are turned off. Only switches S_5 and S_6 are triggered by the frequency modulation scheme. The PWM signals and the operating circuits in every PWM period are provided in Figure 7.

State 1 [$t_0 \leq t < t_1$]: The capacitor voltage v_{CS5} is decreased to zero at t_5 . Then, D_{S5} becomes forward biased due to $i_{Lr2}(t_0)$ being negative. At this moment, the zero voltage turn-on of S_5 can be achieved. Since i_{Lr1} is positive, D_{S1} is conducting to charge C_1 .

State 2 [$t_1 \leq t < t_2$]: Since $V_{de} > 0$ in this state, i_{Lr2} is increased and has a positive value after time t_1 . Therefore, i_{Lr2} flows through S_5 instead of D_{S5} . Reverse power flow is from V_{C3} to V_{C1} .

State 3 [$t_2 \leq t < t_3$]: $i_{Lr1} = 0$ at time t_2 and D_{S1} turns off. Then, i_{Lr2} is equal to the magnetizing current i_{Lm2} .

State 4 [$t_3 \leq t < t_4$]: S_5 turns off at $t = t_3$ and $i_{Lr2}(t_3) > 0$. Thus, C_{S5} (C_{S6}) is charged (discharged).

State 5 [$t_4 \leq t < t_5$]: The capacitor voltage v_{CS6} is decreased to zero at $t = t_4$. Then, D_{S6} becomes forward biased due to $i_{Lr2}(t_4)$ being positive. After time t_4 , the zero voltage turn-on of S_6 can be realized. In state 5, $i_{Lr1}(t)$ is negative and D_{S2} becomes forward biased. Since the leg voltage $V_{de} = -V_{C4}$, i_{Lr2} decreases in this state.

State 6 [$t_5 \leq t < t_6$]: After time t_5 , $i_{Lr2} < 0$. Then, i_{Lr2} flows through switch S_6 instead of diode D_{S6} . Reverse power conversion is achieved from the V_{C4} terminal to V_{C2} terminal.

State 7 [$t_6 \leq t < t_7$]: The left-hand side current $i_{Lr1} = 0$ at time t_6 and D_{S2} becomes reverse biased.

State 8 [$t_7 \leq t < T_{sw} + t_0$]: S_6 turns off at time t_7 and i_{Lr2} charges (discharges) C_{S6} (C_{S5}). At time $T_{sw} + t_0$, the switching cycle is completed.

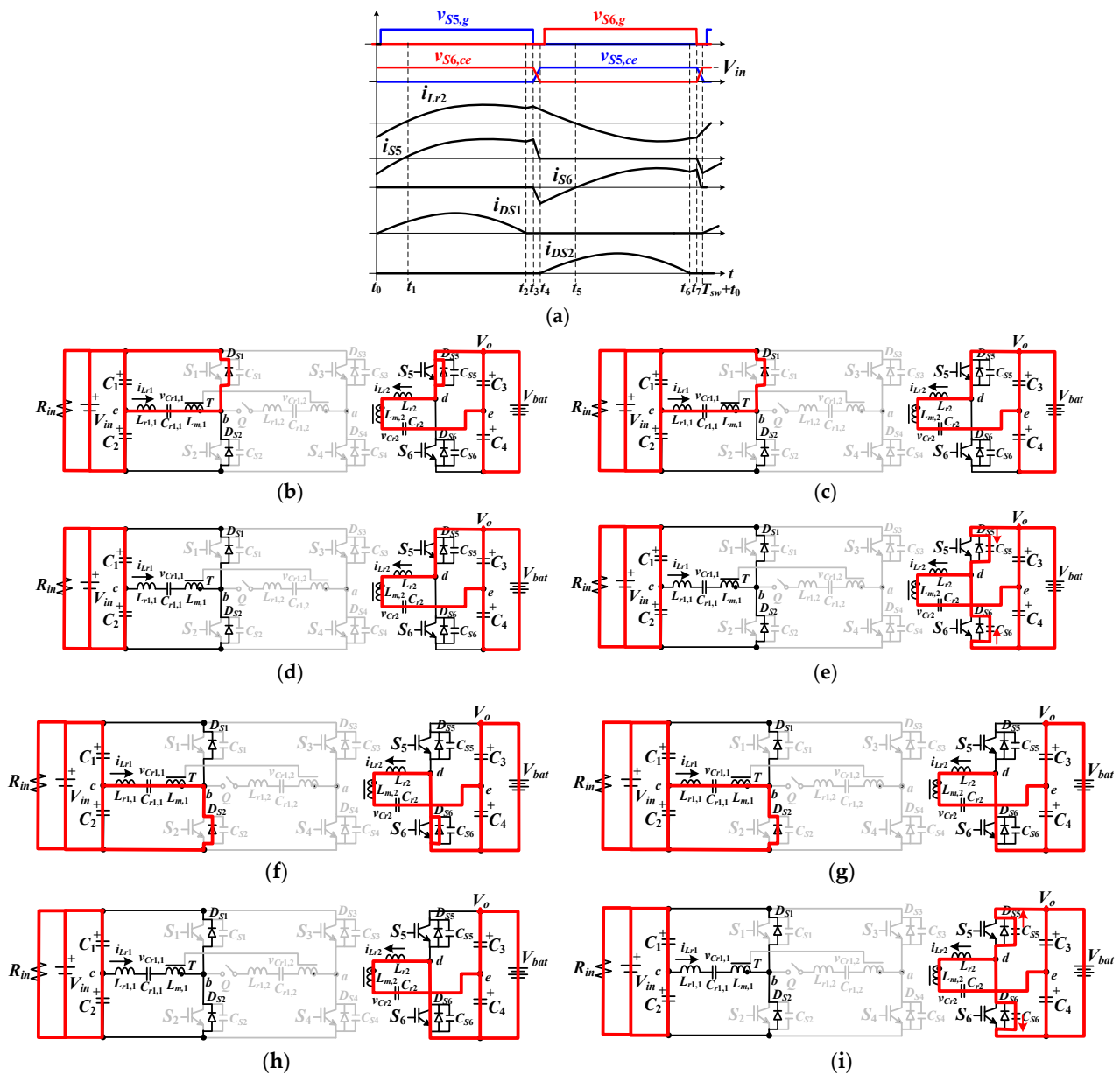


Figure 7. Circuit waveforms and state operations under reverse power operation: (a) PWM waveforms, (b) state 1, (c) state 2, (d) state 3, (e) state 4, (f) state 5, (g) state 6, (h) state 7, (i) state 8.

3. Circuit Properties and Design Example

The proposed bidirectional converter has three operating modes (Figure 4b–d). The output voltage is controlled using a variable switching frequency. The fundamental switching frequency approach [20] is selected to control the voltage gain of the resonant circuit. According to the on-off states of the power semiconductors, the leg voltages are obtained as V_{ac} (or V_{bc}) = $V_{in}/2$ or $-V_{in}/2$ and V_{de} = $V_o/2$ or $-V_o/2$. The equivalent resonant tank circuits of Figure 4b–d can be obtained in Figure 8. In Figure 4b, the turns ratio is $n_1 = 2n$. However, the effective transformer turns ratio in Figure 4c,d is $n_2 = n$. The voltage gain of the proposed circuit can be obtained from the root mean square (*rms*) values of the input and output voltages at fundamental switching frequency. The *rms* leg voltages of input and output terminals are calculated as $V_{ac,f} = \sqrt{2}V_{in}/\pi$ in Figure 4b, $V_{bc,f} = \sqrt{2}V_{in}/\pi$ in Figure 4c,d, and $V_{de,f} = \sqrt{2}V_o/\pi$. The equivalent resonant tanks at fundamental switching frequency are provided in Figure 9. For forward power operation, the equivalent load resistance $R_{o,e} = \frac{2n_1^2}{\pi^2}R_o$ for low voltage output, $R_{o,e} = \frac{2n_2^2}{\pi^2}R_o$ for high voltage output, and $R_{in,e} = \frac{2R_{in}}{\pi^2 n_2^2}$ for reverse power operation.

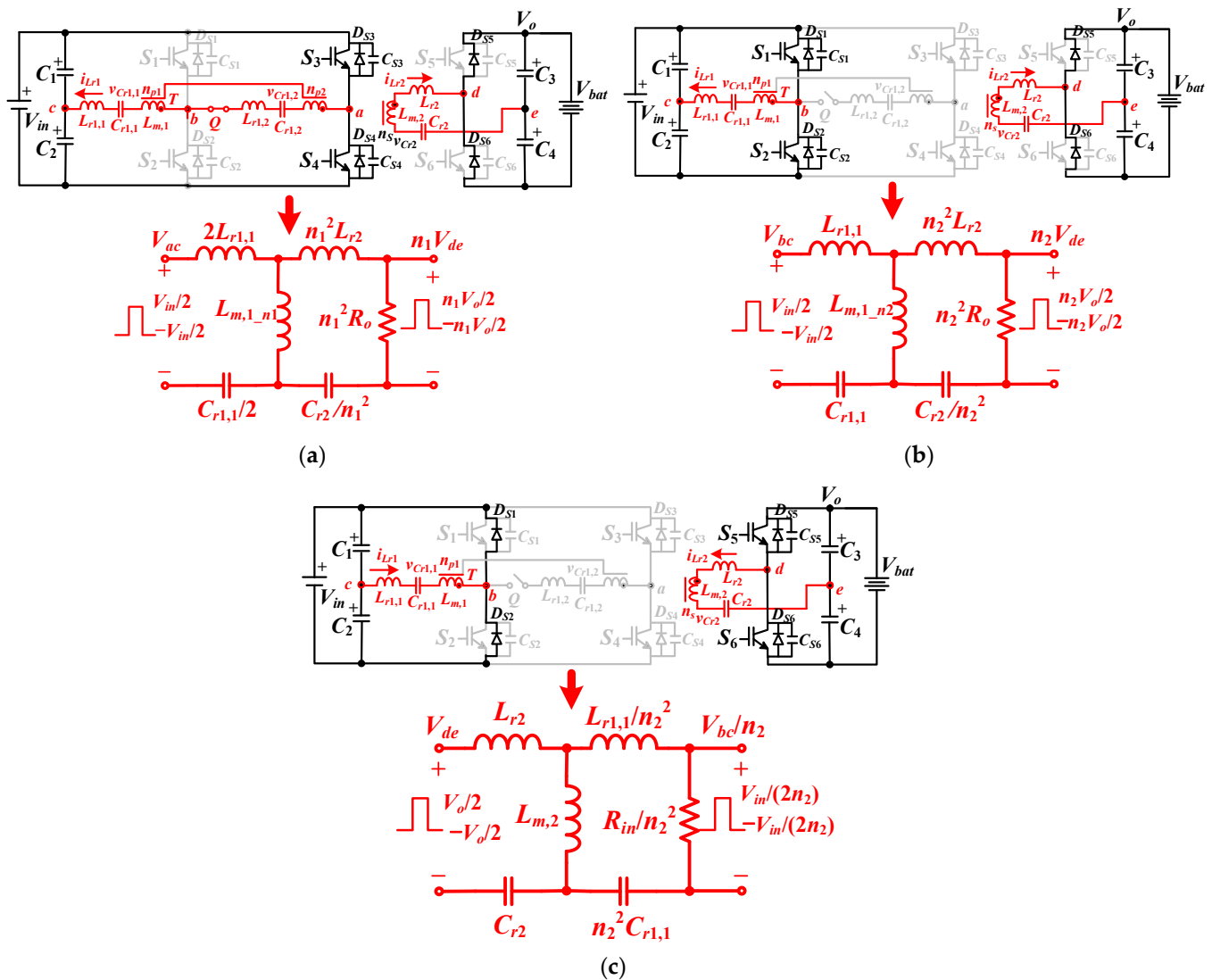


Figure 8. Resonant tank circuits at (a) forward power conversion and low output voltage range, (b) forward power conversion and high output voltage range, (c) reverse power conversion.

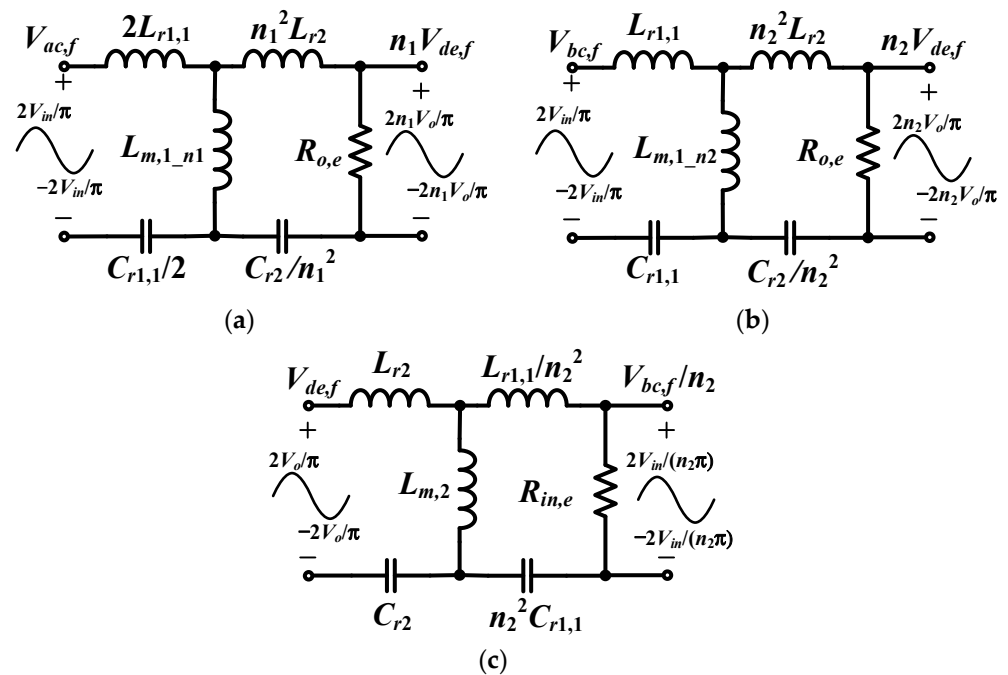


Figure 9. Resonant circuits at fundamental switching frequency at (a) forward power conversion and low output voltage range, (b) forward power conversion and high output voltage range, (c) reverse power conversion.

Based on the equivalent resonant circuit in Figure 9, the transfer functions of the proposed converter for the three operating modes are obtained and expressed as.

$$G_{F,L}(s) = \frac{n_1 V_{def}}{V_{ac,f}} = \frac{n_1 V_o}{V_{in}}$$

$$= \frac{sL_{m,1-n1} // \left(sn_1^2 L_{r2} + \frac{1}{sC_{r2}/n_1^2} + R_{o,e} \right)}{2sL_{r1,1} + \frac{1}{sC_{r1,1}} + \left[sL_{m,1-n1} // \left(sn_1^2 L_{r2} + \frac{1}{sC_{r2}/n_1^2} + R_{o,e} \right) \right]} \times \frac{R_{o,e}}{sn_1^2 L_{r2} + \frac{1}{sC_{r2}/n_1^2} + R_{o,e}} \quad (1)$$

$$G_{F,H}(s) = \frac{n_2 V_{def}}{V_{bc,f}} = \frac{n_2 V_o}{V_{in}}$$

$$= \frac{sL_{m,1-n2} // \left(sn_2^2 L_{r2} + \frac{1}{sC_{r2}/n_2^2} + R_{o,e} \right)}{sL_{r1,1} + \frac{1}{sC_{r1,1}} + \left[sL_{m,1-n2} // \left(sn_2^2 L_{r2} + \frac{1}{sC_{r2}/n_2^2} + R_{o,e} \right) \right]} \times \frac{R_{o,e}}{sn_2^2 L_{r2} + \frac{1}{sC_{r2}/n_2^2} + R_{o,e}} \quad (2)$$

$$G_R(s) = \frac{V_{bc,f}/n_2}{V_{def}} = \frac{V_{in}}{n_2 V_o}$$

$$= \frac{sL_{m,2} // \left(sL_{r1,1}/n_2^2 + \frac{1}{sn_2^2 C_{r1,1}} + R_{in,e} \right)}{sL_{r2} + \frac{1}{sC_{r2}} + \left[sL_{m,2} // \left(sL_{r1,1}/n_2^2 + \frac{1}{sn_2^2 C_{r1,1}} + R_{in,e} \right) \right]} \times \frac{R_{in,e}}{sL_{r1,1}/n_2^2 + \frac{1}{sn_2^2 C_{r1,1}} + R_{in,e}} \quad (3)$$

where $G_{F,L}(s)$, $G_{F,H}(s)$, and $G_R(s)$ are the transfer functions of Figure 4b–d, respectively. To further simplify the transfer function Equations in (1)–(3), the circuit parameters are defined as $C_{r1,1} = C_{r1,2} = C_{r2}/n^2$, $L_{r1,1} = L_{r1,2} = n^2 L_{r2}$, $Q_{F,L} = \sqrt{4L_{r1,1}/C_{r1,1}/R_{o,e}}$, $Q_{F,H} = \sqrt{L_{r1,1}/C_{r1,1}/R_{o,e}}$, $Q_R = \sqrt{L_{r2}/C_{r2}/R_{in,e}}$, $K_{F,L} = L_{m,1-n1}/2L_{r1,1}$, $K_{F,H} = L_{m,1-n2}/L_{r1,1}$, $K_R = L_{m,2}/L_{r2}$ and $F_F = F_R = f_{sw}/f_r$. Then the voltage gains for the three operating modes are expressed as:

$$|G_{F,L}| = \frac{n_1 V_o}{V_{in}} = \frac{1}{\sqrt{\left[1 + \frac{1}{K_{F,L}^2} - \frac{1}{K_{F,L}^2 F_F^2} \right]^2 + Q_{F,L}^2 \left[F_F \left(2 + \frac{1}{K_{F,L}} \right) - \frac{1}{F_F} \left(2 + \frac{2}{K_{F,L}} - \frac{1}{K_{F,L} F_F^2} \right) \right]^2}} \quad (4)$$

$$|G_{F,H}| = \frac{n_2 V_o}{V_{in}} = \frac{1}{\sqrt{\left[1 + \frac{1}{K_{F,H}} - \frac{1}{K_{F,H} F_F^2}\right]^2 + Q_{F,H}^2 \left[F_F \left(2 + \frac{1}{K_{F,H}}\right) - \frac{1}{F_F} \left(2 + \frac{2}{K_{F,H}} - \frac{1}{K_{F,H} F_F^2}\right)\right]^2}} \quad (5)$$

$$|G_R| = \frac{V_{in}}{n_2 V_o} = \frac{1}{\sqrt{\left[1 + \frac{1}{K_R} - \frac{1}{K_R F_R^2}\right]^2 + Q_R^2 \left[F_R \left(2 + \frac{1}{K_R}\right) - \frac{1}{F_R} \left(2 + \frac{2}{K_R} - \frac{1}{K_R F_R^2}\right)\right]^2}} \quad (6)$$

These three voltage gains have similar equations with different circuit parameters and transformer turns ratios. According to Equations (4)–(6), Figure 10 illustrates the curves of voltage gain under $K = 5$.

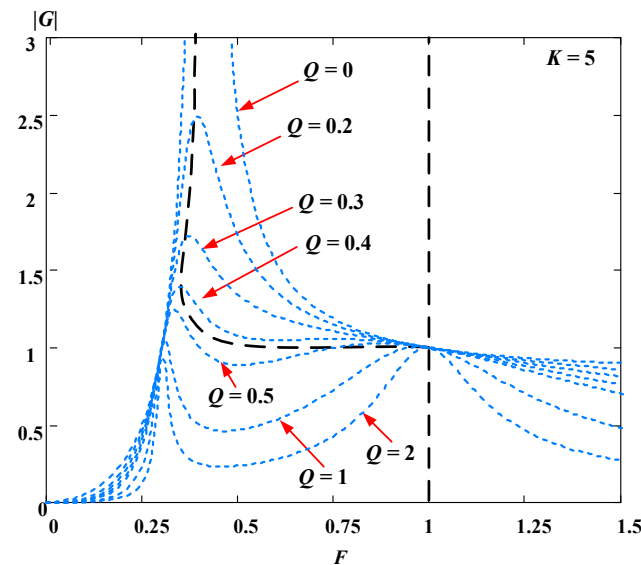


Figure 10. Voltage gain of the presented circuit.

The electric parameters of the circuit prototype are $V_{in} = 400$ V, $V_o = 150$ V–450 V, $P_{o,max} = 1$ kW at $V_o = 450$ V condition and $f_r = 80$ kHz. When $V_o = 150$ –300 V, a half bridge resonant circuit with transformer turns ratio $2n$ is used in Figure 4b, to control the load voltage. If $V_o = 300$ –450 V, then the half bridge resonant circuit with transformer turns ratio n is adopted in Figure 4c to control output voltage. When reverse power flow is required, the half bridge resonant circuit on the right-hand side is adopted (Figure 4d), to transfer power flow from the low side to input side. Since the three operating modes in Figure 4 have similar voltage transfer functions, the circuit parameters of the prototype circuit are obtained from Figure 4c with $V_o = 300$ –450 V in this design example. First, the unity gain of $M_{F,H}|$ is designed at $V_o = 300$ V. From Equation (5), the turn ratio $n = n_{p1}/n_s$ is obtained in Equation (7).

$$n_2 = n = n_{p1}/n_s = |G_{F,H}| \times \frac{V_{in}}{V_{o,min}} = 1 \times \frac{400}{300} \approx 1.333 \quad (7)$$

The magnetic core PC40/EE55 with primary turns $n_{p1} = n_{p2} = 26$ and secondary turns $n_s = 20$ is adopted to implement transformer T . Therefore, the turns ratio n_2 becomes $26/20 = 1.3$. Then, the voltage gains at $V_o = 300$ V and $V_o = 450$ V conditions are expressed as

$$|G_{F,H}|_{min} = \frac{n_2 V_{o,min}}{V_{in}} = \frac{1.3 \times 300}{400} \approx 0.975 \quad (8)$$

$$|G_{F,H}|_{max} = \frac{n_2 V_{o,max}}{V_{in}} = \frac{1.3 \times 450}{400} \approx 1.46 \quad (9)$$

Similarly, the voltage gains at $V_o = 150$ V and $V_o = 300$ V conditions under low voltage output range operation in Figure 4b are expressed as.

$$|G_{F,L}|_{\min} = \frac{n_1 V_{o,\min}}{V_{in}} = \frac{2 \times 1.3 \times 150}{400} \approx 0.975 \quad (10)$$

$$|G_{F,L}|_{\max} = \frac{n_1 V_{o,\max}}{V_{in}} = \frac{2 \times 1.3 \times 300}{400} \approx 1.95 \quad (11)$$

To obtain a voltage gain of more than 1.95 at $V_o = 300$ V under low voltage output range ($|G_{F,L}|_{\max} \approx 1.95$), the quality factors $Q = 0.2$ and $K = 5$ are selected. For forward power flow and high voltage output range (300–450 V), the quality factor $Q_{F,H} = \sqrt{L_{r1,1}/C_{r1,1}/R_{o,e}}$ and the inductor ratio $K_{F,H} = L_{m,1}n_2/L_{r1,1}$. The resistance $R_{o,e}$ at full load is calculated as:

$$R_{o,e} = \frac{2n_2^2}{\pi^2} R_o = \frac{2 \times 1.3^2}{3.14159^2} \times \frac{450^2}{1000} \approx 69 \, \Omega \quad (12)$$

Then, the resonant components can be calculated as:

$$L_{r1,1} = L_{r1,2} = Q_{F,H} R_{o,e} / (2\pi f_r) = 0.2 \times 69 / (2 \times 3.14159 \times 80000) \approx 27.45 \, \mu\text{H} \quad (13)$$

$$L_{r2} = L_{r1,1} / n_2^2 = 27.45 / 1.3^2 \approx 16.24 \, \mu\text{H} \quad (14)$$

$$L_{m,1} = K_{F,H} L_{r1,1} = 5 \times 27.45 \approx 137.25 \, \mu\text{H} \quad (15)$$

$$C_{r1,1} = C_{r1,2} = 1 / (2\pi Q_{F,H} f_r R_{o,e}) = 1 / (2 \times 3.14159 \times 0.2 \times 80000 \times 69) \approx 144 \, \text{nF} \quad (16)$$

$$C_{r2} = n_2^2 C_{r1,1} = 1.3^2 \times 144 \approx 243 \, \text{nF} \quad (17)$$

Table 2 shows the circuit components in the laboratory prototype.

Table 2. Circuit components in the laboratory prototype.

Items	Parameter
C_1, C_2	660 μF
C_3, C_4	660 μF
$C_{r1,1}, C_{r1,2}$	144 nF
C_{r2}	243 nF
$L_{r1,1}, L_{r1,2}$	27.45 μH
L_{r2}	16.24 μH
$L_{m,1}$	137.25 μH
S_1 – S_6	GP50B60PD1 (600 V/33 A)
Q	6R125P6 (650 V/19 A)
Transformer $n_{p1}:n_{p2}:n_s$	26:26:20

4. Experimental Verifications

The output voltage V_o between 150 V and 450 V and the switching signals of switches Q , S_1 , and S_3 are provided in Figure 11a. If V_o is between 150 V and 300 V, S_1 is off and Q is on. The half bridge circuit with switches S_3 and S_4 , as shown in Figure 4b, is controlled to regulate load voltage. If V_o is between 300 V and 450 V, Q and S_3 are off, and switches S_1 and S_2 are operated to control load voltage. The constant voltage (CV) and constant current (CC) modes are used to charge the battery. If the battery voltage is less than 450 V, then the battery charge current is controlled at CC mode with $I_o = 2.3$ A. If the battery voltage is close to 450 V, then CV mode with $V_o = 450$ V is selected to charge the battery. Therefore, the load current I_o will be decreased from 2.3 A. In Figures 12–15, the experimental waveforms operated at constant current mode with $V_o = 150$ V, 310 V and 450 V are provided, to show the circuit characteristics under a forward power flow condition. Figure 12a–c shows the measured waveforms of input voltage, load current, load voltage, and switch signal under $V_{in} = 400$ V, $I_o = 2.3$ A, and $V_o = 150$, 310, and 450 V, respectively, with CC

mode control. When $V_o = 150$ V, the converter is controlled in low voltage output mode (Figure 4b). Therefore, switch S_3 is controlled with variable switching frequency, to regulate load voltage, as shown in Figure 12a. When $V_o = 310$ V and 450 V, the proposed circuit is operated in high voltage output mode (Figure 4c). Thus, switch S_1 is controlled to adjust the load voltage, as shown in Figure 12b,c. Figure 13 provides the test results of the leg voltage and the primary current and voltage under $V_o = 150$ V, 310 V, and 450 V outputs. The square voltage with ± 200 V is generated on the leg voltage V_{ac} at $V_o = 150$ V condition or V_{bc} at $V_o = 310$ V and 450 V conditions. The resonant voltage $v_{Cr1,1}$ is almost a sinusoidal voltage. In low voltage output mode, the voltage gain at $V_o = 150$ V output is close to unity. Therefore, f_{sw} (switching frequency) is close to 80 kHz (series resonant frequency) and i_{Lr1} resembles a sinusoidal waveform. In high voltage output mode, the proposed converter has a high (or low) voltage gain under $V_o = 450$ V (or 310 V) output. Thus, $f_{sw} < f_r$ at $V_o = 450$ V condition. Therefore, the primary current i_{Lr1} resembles a quasi-sinusoidal waveform at $V_o = 450$ V output. The measured secondary side waveforms are shown in Figure 14. Since $f_{sw} > f_r$ at $V_o = 150$ V and 310 V conditions are more than series resonant frequency, the diode currents i_{DS5} and i_{DS6} are turned off at hard switching, as shown in Figure 14a,b. However, D_{S5} and D_{S6} turn off at zero current under $V_o = 450$ V output (Figure 14c), due to $f_{sw} < f_r$. Figure 15 provides the experimental waveforms of the active switch for different output voltage conditions. The switch waveforms at full rated current and 150 V output case are shown in Figure 15a. The switch S_3 is tuned on with zero voltage switching, due to the collect-to-emitter voltage being reduced to zero voltage before S_3 turns on. In the same way, S_1 also turns on at zero voltage switching under 310 V output (Figure 15b) and 450 V output (Figure 15c,d). Figure 16 shows the experimental results under reverse power conversion at $V_{in} = 400$ V, $V_o = 450$ V, and full power condition. Only switches S_5 and S_6 are controlled to realize reverse power flow operation. Figure 16a gives the measured waveforms V_o , V_{in} , I_{in} , and $v_{S5,g}$. The measured right hand side waveforms are given in Figure 16b, and the measured left hand side waveforms are given in Figure 16c. Since the voltage gain under the reverse power flow operation is close to unity, f_{sw} is almost equal to f_r . Thus, the resonant voltage and current waveforms are almost sinusoidal waveforms, as shown in Figure 16b. Figure 16d,e show the test results of switch S_5 at 100% and 50% rated power conditions. S_5 turns on at zero voltage switching. The measured circuit efficiencies are 90.7%, 92.5%, and 91.7% at $V_o = 150$ V, 310 V, and 450 V under CC mode control.

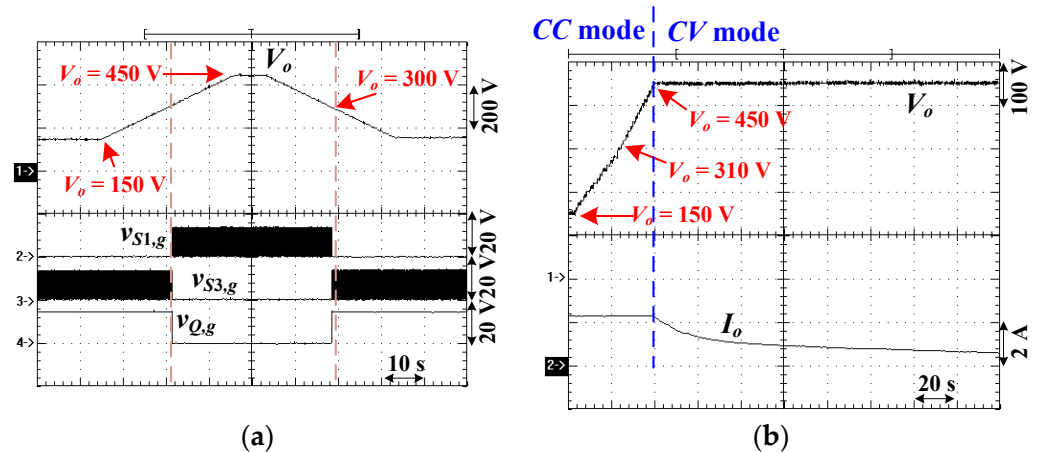


Figure 11. Measured waveforms: (a) V_o , $v_{S1,g}$, $v_{S3,g}$, and $v_{Q,g}$, (b) V_o and I_o .

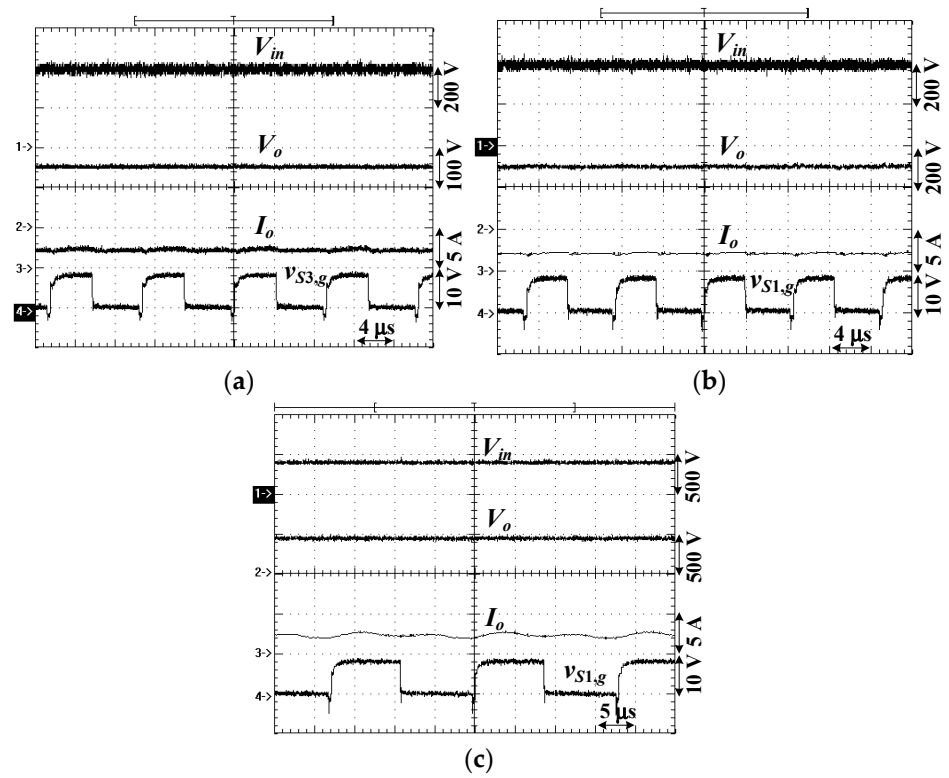


Figure 12. Experimental waveform of V_{in} , V_o , I_o and switch signal $v_{S3,g}$ (or $v_{S1,g}$) under constant current mode at (a) $V_o = 150$ V, (b) $V_o = 310$ V, (c) $V_o = 450$ V.

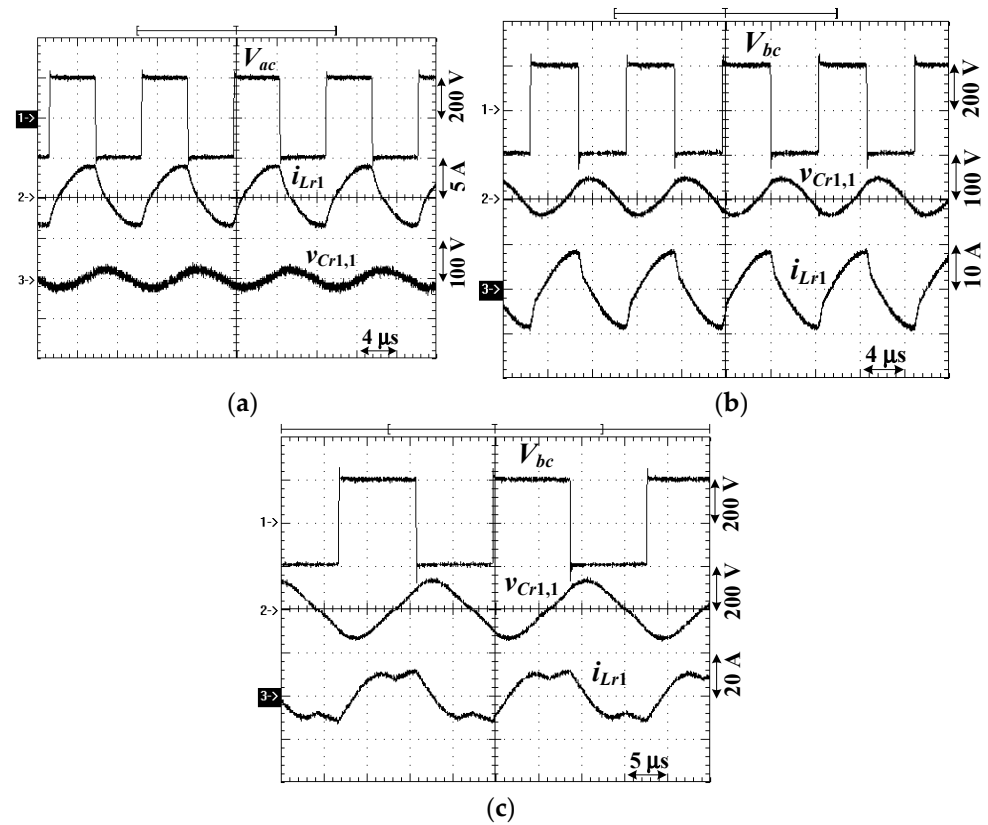


Figure 13. Experimental waveform of leg voltage V_{ac} (or V_{bc}), $v_{Cr1,1}$ and i_{Lr1} under constant current mode at (a) $V_o = 150$ V, (b) $V_o = 310$ V, (c) $V_o = 450$ V.

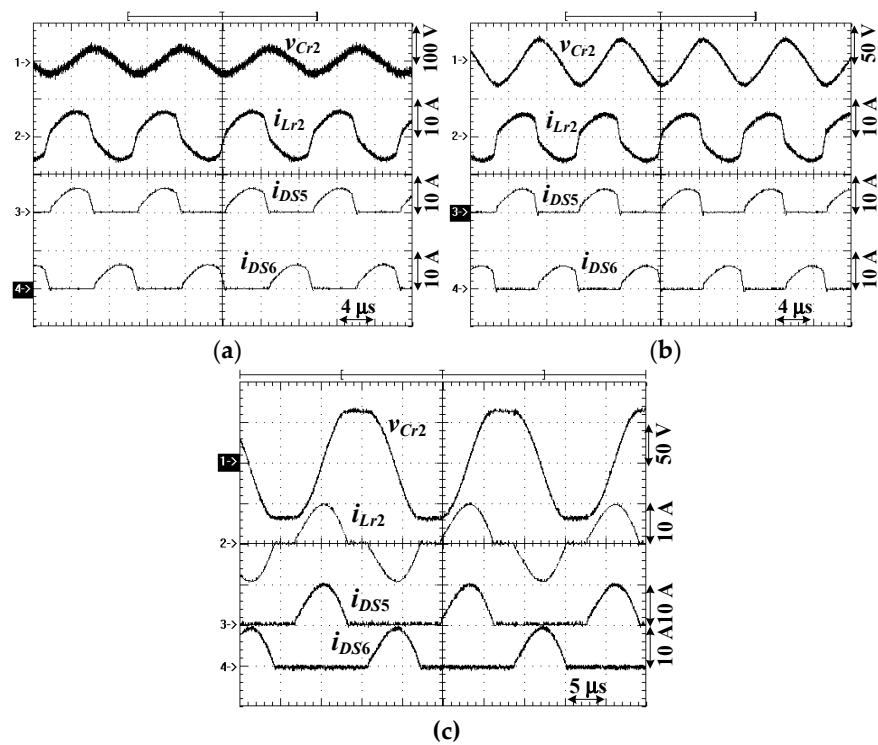


Figure 14. Experimental waveform on the secondary side v_{Cr2} , i_{Lr2} , i_{DS5} and i_{DS6} under constant current mode at (a) $V_o = 150$ V, (b) $V_o = 310$ V (c) $V_o = 450$ V.

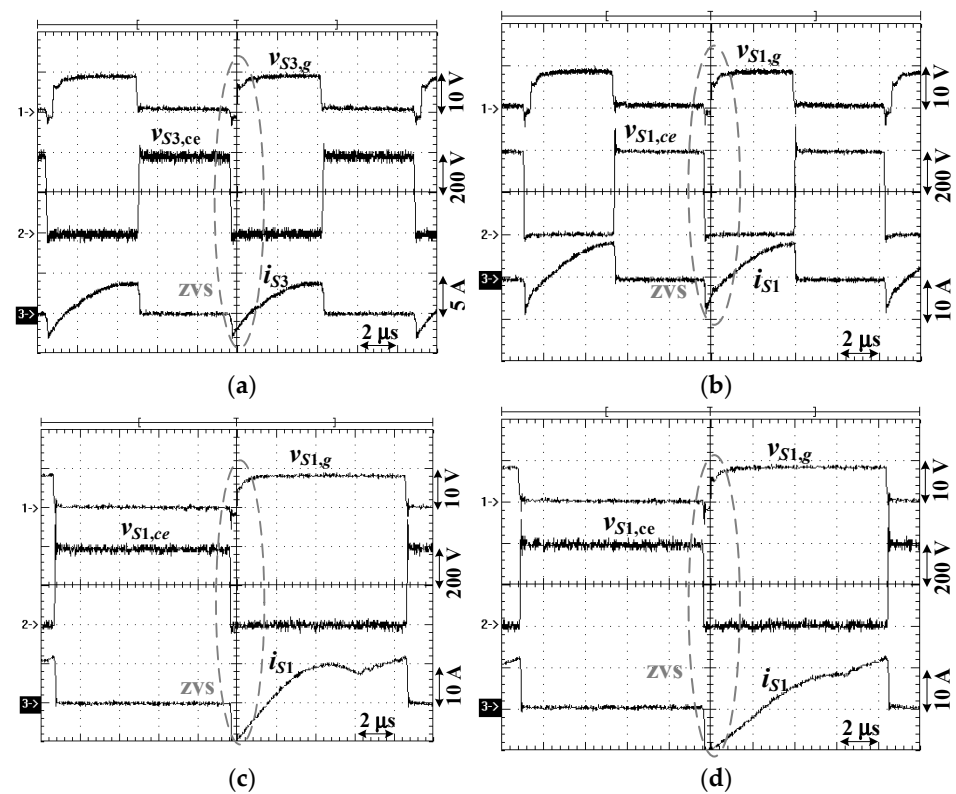


Figure 15. Experimental waveforms of active switch: (a) S_3 at $V_o = 150$ V and $I_o = 2.3$ A (b) S_1 at $V_o = 310$ V and $I_o = 2.3$ A (c) S_1 at $V_o = 450$ V and $I_o = 2.3$ A (rated power) (d) S_1 at $V_o = 450$ V and $I_o = 1.15$ A (50% rated power).

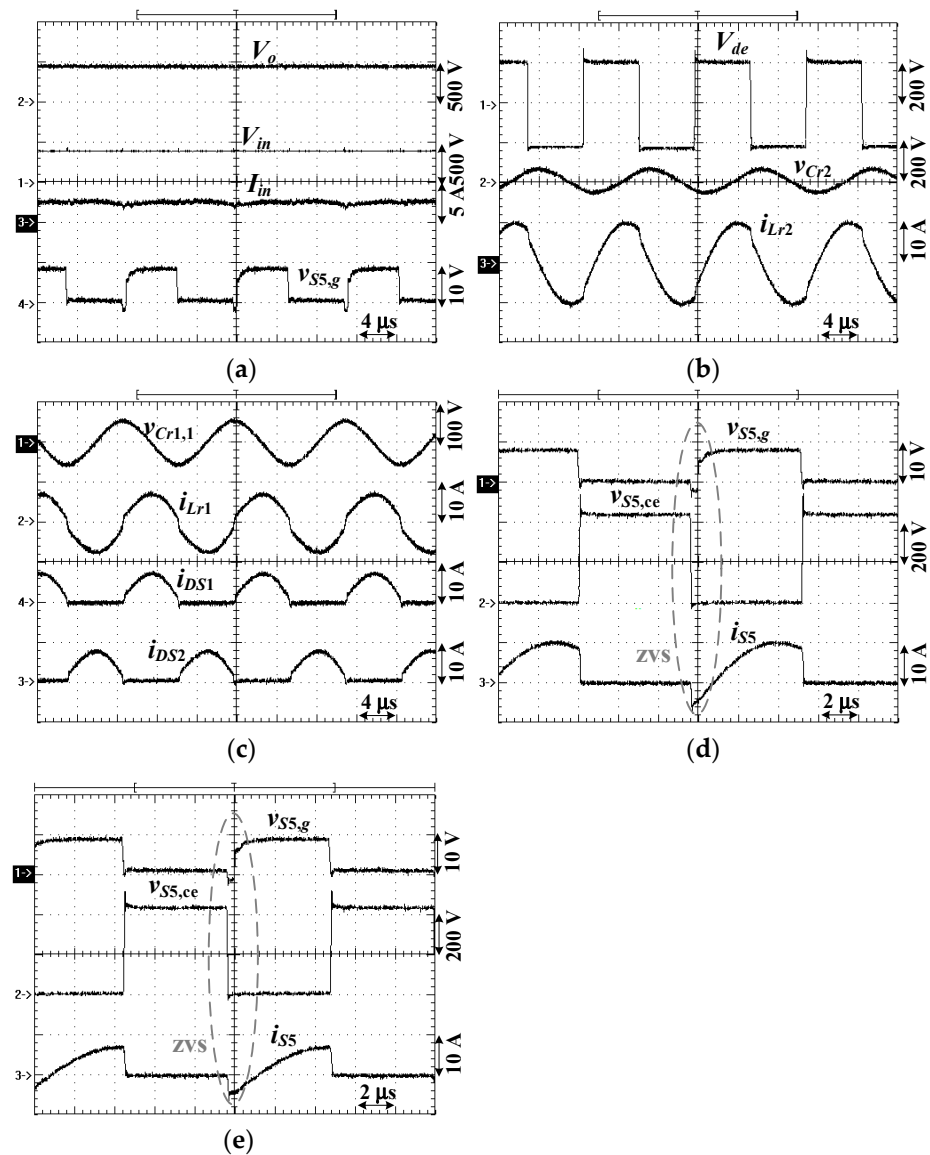


Figure 16. Experimental waveform under reverse power operation with $V_o = 450$ V and $V_{in} = 400$ V at rated power (a) V_o , V_{in} , I_{in} and $v_{S5,g}$, (b) v_{de} , v_{Cr2} and i_{Lr2} , (c) $v_{Cr1,1}$, i_{Lr1} , i_{DS1} and i_{DS2} , (d) $v_{S5,g}$, $v_{S5,ce}$ and i_{S5} at 100% load, (e) $v_{S5,g}$, $v_{S5,ce}$ and i_{S5} at 50% load.

5. Conclusions

A frequency control DC/DC converter with the advantages of bidirectional power conversion, soft switching operation, and wide voltage operation is presented and implemented for battery charging and discharging applications. Symmetric half bridge resonant circuits are used to achieve bidirectional power flow action and voltage step up/down operation. By using a variable transformer turns ratio, the proposed converter has a wide voltage output capability, to overcome the drawback of wide frequency variation in traditional resonant converters. The presented circuit is controlled at inductive load characteristics, so that power semiconductors are turned on (or off) at zero voltage (or zero current) switching. Thus, the electromagnetic interference and switching losses can be reduced. The circuit schematic, system operation, and design procedures of the proposed circuit are discussed and provided in detail. The circuit effectiveness was confirmed and verified by the experimental results with a 1-kW power scaled down prototype.

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