Frequency-Controlled Current-Fed Resonant Converter with No Input Ripple Current

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Abstract: This paper studies a frequency-controlled current-fed resonant circuit. The adopted direct current (DC)-to-DC converter contains two boost circuits and a resonant circuit on the primary side. First, two boost circuits are connected in parallel to achieve voltage step-up and reduce input ripple current by using interleaved pulse-width modulation. Therefore, the size and current rating of boost inductors are decreased in the proposed converter. Second, the boost voltage is connected to the resonant circuit to realize the mechanism of the zero-voltage switching of all active switches and zero-current switching of all diodes. Two boost circuits and a resonant circuit use the same power devices in order to lessen the switch counts. The voltage doubler topology is adopted on the secondary side (high-voltage side). Therefore, the voltage rating of diodes on the high-voltage side is clamped at output voltage. The feasibility of the studied circuit is confirmed by the experimental tests with a 1 kW prototype circuit.

Keywords: current-fed resonant converter; frequency control; ripple current; boost converter

1. Introduction

Due to increased global warming and climbing temperature issues, renewable energy sources have been developed to produce clean energy. A fuel cell is a kind of renewable energy source that converts chemical energy to electric direct current (DC) or alternating current (AC) power. A solar cell is the other source of renewable energy to convert photovoltaic (PV) energy to electric DC or AC power. The outputs of fuel cell stacks and PV panels are low voltage. For connecting fuel cell stacks and PV panels to AC grids or DC grids [1,2], a voltage boost circuit and an isolation transformer are necessary. The high-voltage boost converters can be the voltage source type [3,4] or the current source type [5–7] circuit topologies. Normally, the input current ripple of the current source converters is much smaller than the voltage source converters. However, the main problems of high-voltage boost converters operating under high switching frequency are serious core losses and switching losses. Power converters with soft-switching turn-on or turn-off characteristics have been proposed and studied to overcome these problems. Duty cycle control [8–13] and frequency control schemes [14–16] are normally used to regulate load voltage and also reduce switching losses under zero-voltage or zero-current. The advantages of duty cycle control with fixed switching frequency are easy implementation with commercial integrated circuits and many available circuit topologies. However, the drawback of these topologies is poor circuit efficiency at low load, due to hard switching at low output power. Resonant converters [14–16] with frequency modulation have developed to achieve advantages of low switching losses at whole load range and high efficiency under low load conditions. Two boost converters and two full-bridge resonant converters with interleaved pulse-width modulation (PWM) are adopted in [5] to obtain high voltage gain and less input ripple current for electric vehicle (EV) and hybrid electric vehicle (HEV) applications. However, more power devices (10 power switches) are used in this circuit topology such that the cost is increased and the reliability is reduced.
This paper proposes and studies a simple voltage-boost current-fed resonant DC/DC circuit. The studied converter has two voltage boost circuits, a frequency-controlled full-bridge circuit and a voltage doubler rectifier. The voltage boost circuits and full-bridge circuit use the same active devices so that the total active devices are decreased. Interleaved PWM is adapted to control two voltage boost circuits. Due to the interleaved PWM operation with one-half cycle phase shift, the input current ripple of the proposed converter is reduced to zero. Since the full-bridge circuit is controlled by frequency modulation, the mechanism of the zero-voltage switching of all active switches and zero-current switching of all diodes can be realized at whole load range. Therefore, the turn-on switching losses of active switches and turn-off switching losses or the reverse recovery current losses of the diodes are decreased. A voltage doubler circuit topology is used on the secondary side (high-voltage side) in order to limit the voltage stress of diode at load voltage. In Section 2, the circuit diagram and structure of the studied circuit topology are presented and discussed. The circuit operation is presented in Section 3. The circuit performance and design examples are provided in Section 4. Finally, the feasibility of the developed circuit is verified by a 1 kW prototype circuit in Section 5, followed by the conclusions.

2. Proposed Converter

Figure 1a gives the circuit schematic of the studied converter. \( V_{in} \) is input voltage, \( V_o \) is output voltage, \( Q_1 \sim Q_4 \) are active switches, \( L_{b1} \) and \( L_{b2} \) are boost inductors, \( C_b \) is boost capacitor, \( C_r \) and \( L_r \) are resonant capacitor and inductor, \( T \) is an isolation transformer, \( D_1 \) and \( D_2 \) are fast recovery diodes, and \( C_{o1} \) and \( C_{o2} \) are output capacitors. On the secondary side (high-voltage side), a voltage doubler rectifier is employed to limit the voltage rating of fast recovery diodes for high voltage output applications. The proposed circuit includes two interleaved boost converters (\( V_{in}, Q_1 \sim Q_4, L_{b1}, L_{b2} \) and \( C_b \)) and a full-bridge resonant converter (\( C_b, Q_1 \sim Q_4, L_r, C_r, T, D_1, D_2, C_{o1} \) and \( C_{o2} \)) to achieve high voltage gain, no input ripple current and soft switching of active switches. In Figure 1b, two voltage boost converters are operated by using interleaved PWM to achieve voltage step-up. Since the duty cycle of each of the active switches \( Q_1 \sim Q_4 \) is 0.5, it can obtain \( V_b = 2 V_{in} \). Since the gated signals of two voltage boost converters are phase-shifted by one-half of the switching period, the boost inductor ripple currents \( \Delta i_{Lb1} \) and \( \Delta i_{Lb2} \) are cancelled each other so that the input ripple current \( \Delta i_{in} = \Delta i_{Lb1} + \Delta i_{Lb2} = 0 \). Therefore, no ripple current (\( \Delta i_{in} = 0 \)) is realized at input side. In Figure 1c, the full-bridge resonant circuit is operated by frequency control to produce a nearly sinusoidal current and voltage on the resonant tank by \( L_r, C_r \) and \( L_m \). Due to the resonant tank of the full-bridge circuit being worked at inductive load, the zero-voltage switching of \( Q_1 \sim Q_4 \) is realized at the whole load. Frequency modulation is used to control load voltage. The studied converter can be used in a battery charger/discharger with switches on the high-voltage terminal in the proposed converter and low-voltage input renewable energy source such as solar cell panel.
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Two boost circuits with interleaved PWM schemes are connected in parallel to decrease the current stress of active devices and to obtain no input ripple current. Since \(Q_1 \sim Q_4\) all have 0.5 duty cycles, it can obtain the boost voltage \(V_b = 2V_{in}\). The full-bridge resonant circuit is operated at inductive load with frequency control. Therefore, \(Q_1 \sim Q_4\) have zero-voltage switching with low switching loss, and \(D_1\) and \(D_2\) have zero-current switching with no reverse recovery current loss. The studied converter is assumed \(L_{b1} = L_{b2} = L_b\), \(C_{Q1} = C_{Q2} = C_{Q3} = C_{Q4} = C_{Qo}\), \(C_{o1} = C_{o2}\) and \(V_{o1} = V_{o2} = V_o/2\). Figure 2 gives the PWM waveforms of the studied circuit at a switching cycle. Based on the switching states of \(D_1\), \(D_2\) and \(Q_1 \sim Q_4\), the studied converter can be divided into six operating steps in each switching period when the series resonant frequency is more than the switching frequency. Figure 3 gives these six equivalent circuits. Before step 1, \(Q_1 \sim Q_4\) are off, \(D_2\) conducts, \(i_{Lb1} > 0\), \(i_{Lb2} > 0\) and \(i_{Ir} > 0\).

Figure 1. Proposed converter (a) circuit diagram (b) boost converter operation (c) full-bridge resonant converter operation.

3. Circuit Operation

Two boost circuits with interleaved PWM schemes are connected in parallel to decrease the current stress of active devices and to obtain no input ripple current. Since \(Q_1 \sim Q_4\) all have 0.5 duty cycles, it can obtain the boost voltage \(V_b = 2V_{in}\). The full-bridge resonant circuit is operated at inductive load with frequency control. Therefore, \(Q_1 \sim Q_4\) have zero-voltage switching with low switching loss, and \(D_1\) and \(D_2\) have zero-current switching with no reverse recovery current loss. The studied converter is assumed \(L_{b1} = L_{b2} = L_b\), \(C_{Q1} = C_{Q2} = C_{Q3} = C_{Q4} = C_{Qo}\), \(C_{o1} = C_{o2}\) and \(V_{o1} = V_{o2} = V_o/2\). Figure 2 gives the PWM waveforms of the studied circuit at a switching cycle. Based on the switching states of \(D_1\), \(D_2\) and \(Q_1 \sim Q_4\), the studied converter can be divided into six operating steps in each switching period when the series resonant frequency is more than the switching frequency. Figure 3 gives these six equivalent circuits. Before step 1, \(Q_1 \sim Q_4\) are off, \(D_2\) conducts, \(i_{Lb1} > 0\), \(i_{Lb2} > 0\) and \(i_{Ir} > 0\).

Figure 2. PWM waveforms of the developed circuit.
Step 1 [$t_0$–$t_1$]: At time $t_0$, capacitors $C_{Q1}$ and $C_{Q4}$ are discharged to zero voltage. Since $i_{L31} - i_{Lr} < 0$ and $i_{Lr} + i_{L32} > 0$, the body diodes $D_{Q1}$ and $D_{Q4}$ are forward biased. At this time instant, switches $Q_1$ and $Q_4$ are turned on under zero-voltage switching. Since $Q_1$ and $Q_4$ are turned on in this step, $v_a = \frac{V_o}{2}$.
0 and \( v_b = V_b \). The boost inductor voltages \( v_{l1b1} = V_{in} \) and \( v_{l1b2} = V_{in} - V_b \). Since \( V_b > V_{in} \), \( i_{l1b1} \) increases linearly and \( i_{l1b2} \) decreases linearly.

\[
\frac{di_{l1b1}}{dt} = \frac{V_{in}}{L_b}, \quad \frac{di_{l1b2}}{dt} = \frac{V_{in} - V_b}{L_b}
\]

(1)

For the full-bridge resonant converter, energy is transferred from \( C_b \) to the secondary side (high-voltage side) through a resonant tank with \( L_r \) and \( C_r \). Since \( D_2 \) is forward biased, the primary winding voltage \( v_{lm} = -V_o/2 \), the current of \( L_{mr} \), \( i_{lm} \), decreases linearly with \( -nV_o/(2L_{mr}) \) and \( C_{o2} \) is charged from the secondary current of transformer \( T \). In this step, \( L_r \) and \( C_r \) are resonant with \( f_r = 1/2\pi\sqrt{L_rC_r} \). The solutions of primary side current and voltage are given as

\[
i_{lr}(t) = \frac{nV_o}{2} - 2V_{in} - v_{Cr}(t_0) \sqrt{L_r/C_r} \sin \frac{t - t_0}{\sqrt{L_rC_r}} + i_{lr}(t_0) \cos \frac{t - t_0}{\sqrt{L_rC_r}}
\]

(2)

\[
v_{Cr}(t) = i_{lr}(t_0) \sqrt{L_r/C_r} \sin \frac{t - t_0}{\sqrt{L_rC_r}} + \frac{nV_o}{2} - 2V_{in} - \left[ \frac{nV_o}{2} - 2V_{in} - v_{Cr}(t_0) \right] \cos \frac{t - t_0}{\sqrt{L_rC_r}}
\]

(3)

where \( n = n_p/n_s \). If the series resonant frequency \( f_r \) is more than the switching frequency \( f_{sw} \), then the primary side current of transformer \( T \) decreases to zero current at time \( t_1 \) and the circuit operation goes to step 2. Otherwise, the circuit operation will go to step 3 under \( f_{sw} > f_r \) condition.

**Step 2** \([t_1 \rightarrow t_2]\): \( D_2 \) decreases to zero current at time \( t_1 \) and \( D_2 \) becomes reverse biased. In this step, \( i_{l1b1} \) increases linearly, \( i_{l1b2} \) decreases linearly and \( i_{lr} \) freewheels through \( Q_1 \) and \( Q_4 \). In this step, \( L_r \), \( L_{mr} \) and \( C_r \) are resonant with \( \omega_p = 1/\sqrt{(L_r + L_{mr})C_r} \). The solutions of primary current \( i_{lr} \) and resonant capacitor voltage \( v_{Cr} \) in this freewheeling state are given in (4) and (5).

\[
i_{lr}(t) = -\left( \frac{2V_{in} + v_{Cr}(t_1)}{\sqrt{(L_r + L_{mr})C_r}} \right) \sin \frac{t - t_1}{\sqrt{(L_r + L_{mr})C_r}} + i_{lr}(t_1) \cos \frac{t - t_1}{\sqrt{(L_r + L_{mr})C_r}}
\]

(4)

\[
v_{Cr}(t) = -2V_{in} + \left[ 2V_{in} + v_{Cr}(t_1) \right] \cos \frac{t - t_1}{\sqrt{(L_r + L_{mr})C_r}} + i_{lr}(t_1) \sqrt{(L_r + L_{mr})C_r} \sin \frac{t - t_1}{\sqrt{(L_r + L_{mr})C_r}}
\]

(5)

**Step 3** \([t_2 \rightarrow t_3]\): \( Q_1 \) and \( Q_4 \) are turned off at time \( t_2 \) under zero voltage. Since \( i_{l1b1} - i_{lr} > 0 \) and \( i_{lr} + i_{l1b2} < 0 \), \( C_{Q2} \) and \( C_{Q3} \) are discharged linearly.

\[
\frac{dv_{CQ2}}{dt} = -\frac{2i_{l1b1} - i_{lr}}{2C_Q}, \quad \frac{dv_{CQ3}}{dt} = \frac{i_{lr} + i_{l1b2}}{2C_Q}
\]

(6)

In this step, \( D_1 \) is forward biased and the primary winding voltages \( v_{lm} \) is clamped at \( nV_o/2 \). The discharge time of \( C_{Q2} \) and \( C_{Q3} \) is soon enough so that \( i_{l1b1}, i_{l1b2} \) and \( i_{lr} \) are almost constant during this step.

**Step 4** \([t_3 \rightarrow t_4]\): The voltages on \( C_{Q2} \) and \( C_{Q3} \) are decreased to zero voltage at time \( t_3 \). Since \( i_{l1b1} - i_{lr} > 0 \) and \( i_{lr} + i_{l1b2} < 0 \), \( D_{Q1} \) and \( D_{Q4} \) are forward biased. Therefore, \( Q_2 \) and \( Q_3 \) can be turned on after \( t_3 \) and the zero-voltage switching of \( Q_2 \) and \( Q_3 \) is achieved. In this step, \( v_a = V_b \) and \( v_b = 0 \) so that \( v_{l1b1} = V_{in} - V_b \) and \( v_{l1b2} = V_{in} \). Therefore, \( i_{l1b1} \) decreases linearly and \( i_{l1b2} \) increases linearly.

\[
\frac{di_{l1b1}}{dt} = \frac{V_{in} - V_b}{L_b}, \quad \frac{di_{l1b2}}{dt} = \frac{V_{in}}{L_b}
\]

(7)

The energy is transferred from \( C_b \) to \( C_{o1} \) through \( L_r \) and \( C_r \). Since \( D_1 \) is forward biased, the primary winding voltage \( v_{l1m} = V_o/2 \), \( i_{lm} \) increases linearly and \( C_{o1} \) is charged. If the series resonant frequency \( f_r \) is more than the switching frequency \( f_{sw} \), then the diode current \( i_{q1} \) will decrease to zero at time \( t_4 \) and circuit goes to step 5. On the other hand, the circuit will go to step 6 under the \( f_{sw} > f_r \) condition.
Step 5 [t4~t5]: \(i_D1\) decreases to zero current at \(t_4\) and \(D_1\) becomes reverse biased. During step 5, \(i_{Lb1}\) decreases, \(i_{Lb2}\) increases and \(i_{Lr}\) freewheels through \(Q_2\) and \(Q_3\). \(C_r\), \(L_r\) and \(L_m\) are resonant in step 5.

Step 6 [t5~t0]: At time \(t_5\), \(Q_2\) and \(Q_3\) turn off under zero voltage. Since \(i_{Lb1} - i_{Lr} < 0\) and \(i_{Lr} + i_{Lb2} > 0\), \(C_{Q1}\) and \(C_{Q4}\) are discharged linearly.

\[
\frac{dv_{CQ1}}{dt} = \frac{i_{Lb1} - i_{Lr}}{2C_Q}, \quad \frac{dv_{CQ4}}{dt} = -\frac{i_{Lr} + i_{Lb2}}{2C_Q}
\]

Rectifier diode \(D_2\) is forward biased and \(v_{Lm} = -nV_o/2\). Since the discharge time of \(C_{Q1}\) and \(C_{Q4}\) in step 6 is very soon, \(i_{Lb1}, i_{Lb2}\) and \(i_{Lr}\) are almost constant in this time interval.

4. Circuit Characteristics

4.1. Boost Converter

Two interleaved voltage boost circuits are worked in continuous conduction mode under the following assumptions: (1) power devices are ideal; and (2) inductors and capacitors are linear and time-invariant. Since the average value of the voltages across \(L_{b1}\) and \(L_{b2}\) is zero, it is possible to calculate the following equation based on the flux balance on the boost inductors.

\[V_b = V_{in} / (1 - D)\]  

where \(D\) is a duty ratio of \(Q_1\) and \(Q_4\). Since \(D\) is fixed at 0.5 in the studied circuit, it can be calculated as \(V_b = 2V_{in}\) and the DC voltage transfer function of boost converter is given as \(M_{V} = V_o / V_{in} = 2\). The PWM waveforms of \(Q_1\) and \(Q_3\) are phase shifted by \(T_{sw}/2\). The ripple currents, \(\Delta i_{Lb1}\) and \(\Delta i_{Lb1}\), on two boost inductors \(L_{b1}\) and \(L_{b2}\) are cancelled each other which gives \(\Delta i_{in} = \Delta i_{Lb1} + \Delta i_{Lb2} = 0\). Thus, there is no ripple current (\(\Delta i_{in} = 0\)) from input voltage. The voltage ratings of \(Q_1\)–\(Q_4\) are equal to \(V_b (=2 V_{in})\). The average inductor currents \(i_{Lb1}\) and \(i_{Lb2}\) equal to \(P_o/(2 V_{in})\).

4.2. Full-Bridge Resonant Converter

The full-bridge resonant circuit includes two switching legs to draw two current pulses per switching cycle from the input voltage terminal and to deliver more output power than the half-bridge resonant circuit. To obtain advantages of the good light load efficiency of the series resonant circuit and the ability to control load voltage at light or open load of the parallel resonant circuit, the full-bridge LLC resonant circuit is employed in the studied circuit topology. Fundamental harmonic analysis from Steigerwald’s article [17] is used to obtain the AC gain characteristics. Due to the PWM waveforms of \(Q_1\)–\(Q_4\), the voltage \(v_{ab}\) is a square waveform with two voltage levels, \(V_b (=2 V_{in})\) and \(-V_b (=2 V_{in})\). Thus, the fundamental frequency voltage \(v_{ab,f}\) is expressed as

\[v_{ab,f} = (8V_{in} / \pi) \sin(2\pi f_{sw}t)\]

Due the on-off states of \(D_1\) and \(D_2\), the transformer primary voltage \(v_{Lm}\) is a square waveform and the fundamental magnetizing voltage is given as

\[v_{Lm,f} = \frac{2nV_o}{\pi} \sin(2\pi f_{sw}t - \theta)\]

The fundamental secondary current of transformer is derived as

\[i_{s,T} = \pi I_o \sin(2\pi f_{sw}t - \theta)\]  

Based on (11) and (12), \(R_{ac}\) is obtained from the secondary load \(R_o\) reflected into the primary side.
The amplitude of the AC gain from (14) can be further expressed as

\[ R_{dc} = \frac{v_{Lm,f}}{i_s/T} = \frac{2n^2}{\pi^2} R_o \]  

(13)

Figure 4 gives the AC equivalent circuit of the adopted resonant circuit. \( v_{ab,f} \) is an effectively sinusoidal input voltage and \( R_{dc} \) is an effective load. Therefore, the gain characteristics of the studied resonant circuit under different switching frequencies is calculated as

\[ G_{ac}(f_{sw}) = \frac{R_{ac} \times j\omega_{sw} L_{ac}}{\frac{1}{\omega_{sw} C_{o}} + \frac{R_{ac} \times j\omega_{sw} L_{ac}}{R_{sw} + j\omega_{sw} L_{sw}}} = \frac{1}{1 + K(1 - \frac{1}{f}) + jQ(F - \frac{1}{f})} \]  

(14)

where \( f_{sw} \) is the switching frequency, \( f_r = 1/2\pi \sqrt{L_r C_r}, F = f_{sw}/f_r, K = L_r/L_m \) and \( Q = \sqrt{L_r/C_r}/R_{ac} \). The amplitude of the AC gain from (14) can be further expressed as

\[ |G_{ac}(f_{sw})| = \frac{1}{\sqrt{[1 + K(1 - \frac{1}{f})]^2 + Q^2(F - \frac{1}{f})^2}} \]  

(15)

Figure 4. AC equivalent circuit of the full-bridge resonant circuit.

4.3. Design Procedure of the Developed Converter

A 1000 W prototype is set up to investigate the main circuit parameter design considerations. The electrical specifications of the developed converter are input voltage \( V_i = 44\sim52 \) V, output voltage \( V_o = 400 \) V, output power \( P_o = 1000 \) W, and the series resonant frequency \( f_r = 100 \) kHz. First, the boost converter with interleaved PWM is used to boost input voltage and reduce input ripple current. The duty cycle of \( Q_1\sim Q_4 \) is fixed at 0.5 with frequency modulation. The ripple currents \( \Delta i_{L1} \) and \( \Delta i_{L2} \) are calculated as

\[ \Delta i_{L1} = \Delta i_{L2} = \frac{V_i \times T_{sw}}{2L_b} \]  

(16)

Since the gate signals of \( Q_1 \) and \( Q_3 \) are phase shifted by \( T_{sw}/2 \) and each duty cycle of switches is equal to 0.5, it is obvious that the current ripple on the input side is decreased significantly and equals zero. The adopted boost inductors \( L_{b1} \) and \( L_{b2} \) are 37 \( \mu \)H. That means that the ripple currents on \( L_{b1} \) and \( L_{b2} \) at maximum input voltage and series resonant frequency are

\[ \Delta i_{L1} = \Delta i_{L2} = \frac{V_i \times T_{sw}}{2L_b} = \frac{52 \times 10^{-5}}{2 \times 37 \times 10^{-6}} \approx 7 \text{ A} \]  

(17)

In order to make sure the output voltage can be regulated at all input voltage ranges, the minimum DC voltage gain \( G_{dc,min} \) under \( V_{Ch,max} \) input is designed as unity. Thus, the turn-ratio is obtained in (18).

\[ n = \frac{G_{dc,min} V_{Ch,max}}{V_i/2} = \frac{1 \times 52 \times 2}{400/2} = 0.52 \]  

(18)

Transformer \( T \) is implemented by ferrite core TDK PC40 EER-42 with \( n_p = 13 \) and \( n_s = 25 \). Then, the theoretical maximum DC voltage gain under minimum input voltage is given as

\[ G_{dc,max} = \frac{nV_o}{2V_{Ch,min}} = \frac{0.52 \times 400}{2 \times 44 \times 2} = 1.18 \]  

(19)
Based on (13), $R_{ac}$ under full load conditions can be calculated as

$$R_{ac} = \frac{2n^2}{\pi^2} R_o = \frac{2 \times 0.52^2}{3.1416} \times \frac{400^2}{1000} \approx 8.77 \Omega$$  \hspace{1cm} (20)

The inductor ratio $K = L_r/L_m$ will affect the circulating current loss on the primary side of the resonant converter. The lower inductor ratio $K$ can reduce the circulating current loss due to larger $L_m$. However, the AC gain of resonant converter is reduced. The higher inductor ratio $K$ can obtain larger AC voltage gain. However, the circulating current loss is increased to reduce circuit efficiency. Therefore, the selection $K$ is a compromise between the AC voltage gain and the circulating current loss. Normally, the $K$ is selected between 0.08–0.5. Considering these factors, the inductor ratio and quality factor are designed as $K = L_r/L_m = 1/6$ and $Q = 0.3$. Therefore, $L_r$ and $C_r$ are calculated as

$$L_r = \frac{QR_{ac}}{2\pi f_r} = \frac{0.3 \times 8.77}{2\pi \times 100,000} \approx 4.18 \mu H$$ \hspace{1cm} (21)

$$C_r = \frac{4\pi^2 L_r f_r}{1} = \frac{4\pi^2 \times 4.18 \times 10^{-6} \times (100,000)^2}{1} \approx 606 \text{ nF}$$ \hspace{1cm} (22)

Considering the practical value of $C_r$, a 600 nF film capacitor is used for $C_r$. In a similar way, the resonant inductor $L_r$ is actually selected as 4.22 $\mu H$. Since the inductor ratio $K = L_r/L_m = 1/6$ is adopted, the magnetizing inductance $L_m$ is calculated as

$$L_m = L_r/k = \frac{4.22 \mu H}{1/6} \approx 25.32 \mu H$$ \hspace{1cm} (23)

The voltage stress and average current of fast recovery diodes $D_1$ and $D_2$ are calculated in (24) and (25).

$$v_{D1,\text{stress}} = v_{D2,\text{stress}} \approx V_o = 400 \text{ V}$$ \hspace{1cm} (24)

$$i_{D1,av} = i_{D2,av} = I_o = 2.5 \text{ A}$$ \hspace{1cm} (25)

Diodes OM5262SW with 1000 V voltage rating and 12 A average current rating are adopted for $D_1$ and $D_2$. The voltage rating of $Q_1$–$Q_4$ are 2 $\times$ 52 V = 104 V. The MOSFETs IRF52N15D with 150 V voltage rating and 60 A current rating are adopted for $Q_1$–$Q_4$. The 1100 $\mu F$ capacitor is used for $C_b$ and the 360 $\mu F$ capacitors are used for $C_{o1}$ and $C_{o2}$.

**5. Experimental Results**

The developed converter is implemented by a 1 kW prototype to demonstrate the circuit performance and verify the feasibility and effectiveness of the studied circuit. Figure 5 gives the experimental circuit diagram of the developed circuit. The photocoupler PC817 and voltage regulator TL431 are adopted to control load voltage. The resonant mode control integrated circuit UCC25600 is used to realize frequency modulation and achieve zero-voltage switching. The circuit parameters of active and passive components of the developed circuit are discussed and obtained from the previous section. From the experimental results in Figures 6–12, the test and experimental results agree with the PWM waveforms in Figure 2. The experimental results of the gate voltages of $Q_1$–$Q_4$ at rated power are illustrated in Figure 6. It can be observed that $Q_1$ ($Q_2$) and $Q_3$ ($Q_4$) have complementary PWM waveforms. The switching frequency of $Q_1$, at 44 V input is less than the switching frequency at 52 V input so that the resonant circuit can obtain larger voltage gain to regulate load voltage. The experimental waveforms of input current and boost inductor currents at the rated power are given in Figure 7. It can be observed that $i_{Lb1}$ and $i_{Lb2}$ are interleaved with each other and balanced well. The PWM waveforms of two boost circuits are interleaved by a phase shift of $T_{sw}/2$ so that the input ripple current $\Delta i_{in}$ is reduced almost to zero.
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\[ v_{Q1,g} \]

\[ v_{Q2,g} \]

\[ v_{Q3,g} \]

\[ v_{Q4,g} \]

\( V_{in} = 44 \text{ V} \) 
\( V_{in} = 52 \text{ V} \)

\[ \text{Figure 6. Experimental results of the gating signals of } Q_1 - Q_4 \text{ under the rated power and (a) } V_{in} = 44 \text{ V} \] 
\[ \text{ (b) } V_{in} = 52 \text{ V} \] 

\[ [v_{Q1,g} - v_{Q4,g}: 10 \text{ V/div}; \text{time: } 4 \mu\text{s/div}.] \]

\[ i_{in} \]

\[ i_{L1A} \]

\[ i_{L1B} \]

\[ i_{L2A} \]

\[ i_{L2B} \]

\[ i_{L1} \]

\[ i_{L2} \]

\( \text{Figure 7. Experimental results of input current and boost currents under the rated power and (a) } V_{in} = 44 \text{ V} \) 
\[ \text{ (b) } V_{in} = 52 \text{ V} \]

\[ [i_{in}, i_{L1A}, i_{L1B}, i_{L2A}, i_{L2B}: 5 \text{ A/div}; \text{time: } 4 \mu\text{s/div}.] \]

The test results of boost voltage \( V_b \), boost current \( i_{Cb} \) and switch currents \( i_{Q2} \) and \( i_{Q4} \) under the rated power are shown in Figure 8. It is clear that the frequency of \( i_{Cb} \) is twice the frequency of \( i_{Q2} \) and \( i_{Q4} \). Figure 9 gives the test results of primary voltage and current of full-bridge resonant circuit under the rated power. It can be seen that the higher switching frequency at 52 V input will result in low circulating current compared to at 44 V input.
Figure 8. Measured waveforms of $V_b$, $i_{Q2}$, $i_{Q4}$ and $i_{Q4}$ under the rated power and (a) $V_{in} = 44$ V (b) $V_{in} = 52$ V [\(v_b\): 50 V/div; \(i_{Q2}, i_{Q4}, i_{CB}\): 20 A/div; time: 4 $\mu$s/div].

Figure 9. Measured waveforms of \(v_{ab}\), \(i_{Lr}\) and \(v_{Cr}\) under the rated power and (a) 44 V input voltage (b) 52 V input voltage [\(v_{ab}, v_{Cr}\): 100 V/div; \(i_{Lr}\): 20 A/div; time: 2 $\mu$s/div].

Figure 10. Test results of the switch current and inductor current on the primary side under the rated power (a) \(i_{Lb1}, i_{Q1}, i_{Q2}\) and \(i_{Lr}\) at 44 V input case (b) \(i_{Lb2}, i_{Q3}, i_{Q4}\) and \(i_{Lr}\) at 44 V input case (c) \(i_{Lb1}, i_{Q1}, i_{Q2}\) and \(i_{Lr}\) at 52 V input case (d) \(i_{Lb2}, i_{Q3}, i_{Q4}\) and \(i_{Lr}\) at 52 V input case [\(i_{Lb1}, i_{Lb2}: 10\ A/div; i_{Q1} - i_{Q4}, i_{Lr}: 20\ A/div\); time: 2 $\mu$s/div].
under 52 V input. Switching frequencies are 131 kHz (at 20% load), 118 kHz (at 50% load) and 99 kHz (at 100% load) are 90.5% (at 20% load), 92.6% (at 50% load) and 94.5% (at 100% load) under 52 V input. The measured efficiencies of the developed converter with the 1 kW laboratory prototype clearly provide the claimed high-performance resonant circuit with low switching losses and input current ripple-free is achieved with the developed circuit topology. The zero-voltage switching of active devices on the low-voltage side and zero-current switching of rectifier diodes on the high-voltage side are also realized at the whole load range. The test results with the 1 kW laboratory prototype clearly demonstrate that a theoretical examination and test verification of the proposed converter can be used in a battery charger/discharger with switches on the high-voltage terminal in the proposed converter. If the input voltage is from the PV solar cell panel, a slightly studied converter can be used in a battery charger/discharger with switches on the high-voltage terminal in the proposed converter. If the input voltage is from the PV solar cell panel, a slightly studied converter can be used in a battery charger/discharger with switches on the high-voltage terminal in the proposed converter. If the input voltage is from the PV solar cell panel, a slightly studied converter can be used in a battery charger/discharger with switches on the high-voltage terminal in the proposed converter. If the input voltage is from the PV solar cell panel, a slightly studied converter can be used in a battery charger/discharger with switches on the high-voltage terminal in the proposed converter. If the input voltage is from the PV solar cell panel, a slightly studied converter can be used in a battery charger/discharger with switches on the high-voltage terminal in the proposed converter. If the input voltage is from the PV solar cell panel, a slightly studied converter can be used in a battery charger/discharger with switches on the high-voltage terminal in the proposed converter. If the input voltage is from the PV solar cell panel, a slightly studied converter can be used in a battery charger/discharger with switches on the high-voltage terminal in the proposed converter. If the input voltage is from the PV solar cell panel, a slightly studied converter can be used in a battery charger/discharger with switches on the high-voltage terminal in the proposed converter.

Figure 11. Test results of $V_{o1}$, $V_{o2}$, $i_{D1}$ and $i_{D2}$ under the rated power and (a) 44 V input case (b) $V_{in}$ = 52 V input case [$V_{o1}$, $V_{o2}$: 100 V/div; $i_{D1}$, $i_{D2}$: 5 A/div; time: 4 µs/div].

Figure 12. Measured results of switch $Q_1$ under (a) $V_{in}$ = 44 V and 20% load (b) $V_{in}$ = 44 V and 50% load (c) $V_{in}$ = 44 V and full load (d) $V_{in}$ = 52 V and 20% load (e) $V_{in}$ = 52 V and 50% load (f) $V_{in}$ = 52 V and full load [$v_{Q1,q}$: 10 V/div; $v_{Q1,d}$: 50 V/div; $i_{Q1}$: 20 A/div for (a,b,d,e) and 50 A/div for (c,f); time: 2 µs/div].

Figure 10 illustrates the test results of $i_{L1}$, $i_{L2}$, $i_{Q1}$~$i_{Q4}$ and $i_{Lr}$ under the rated power. The peak current of $i_{Q1}$ is larger than the peak current of $i_{Q2}$ due to $i_{Q1} = i_{L1} - i_{Lr}$ and $i_{Lr} < 0$ when $Q_1$ is conducting. The test and experimental results of the output capacitor voltages and rectifier diode currents under the rated power are shown in Figure 11. $V_{o1}$ and $V_{o2}$ are balanced well and $D_1$ and $D_2$ are turned off under zero-current switching. Figure 12 illustrates the test results of switch $Q_1$ at 20%, 50% and 100% loads. It can be observed that zero-voltage switching of switch $Q_1$ is realized. Since $Q_2$~$Q_4$ have the same operation characteristics of $Q_1$, it can be expected that the zero-voltage switching of switches $Q_2$~$Q_4$ are also achieved. The measured efficiencies of the developed converter are 90.5% (at 20% load), 92.6% (at 50% load) and 94.5% (at 100% load) under 52 V input. The measured switching frequencies are 131 kHz (at 20% load), 118 kHz (at 50% load) and 99 kHz (at 100% load) under 52 V input.
6. Conclusions

A novel frequency-controlled current-fed resonant circuit with no input current ripple is proposed and verified in this paper. Theoretical examination and test verification demonstrate that a high-performance resonant circuit with low switching losses and input current ripple-free is achieved with the developed circuit topology. The zero-voltage switching of active devices on the low-voltage side and zero-current switching of rectifier diodes on the high-voltage side are also realized at the whole load range. The test results with the 1 kW laboratory prototype clearly provide the claimed characteristics. The constructed converter can be applied in the renewable energy conversion system with low-voltage input (48 V) and high-voltage output (400 V) with no input current ripple. The studied converter can be used in a battery charger/discharger with switches on the high-voltage terminal in the proposed converter. If the input voltage is from the PV solar cell panel, a slightly wider input voltage variation is expected. Then, the wider switching frequency range and maximum power point tracking must be implemented to regulate the load voltage.

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