A Wide Gain Range LLC Resonant Converter Based on Reconfigurable Bridge and Asymmetric Resonant Tanks

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Abstract—In wide voltage gain range applications, it is demanding to optimize the design of the conventional frequency modulated LLC resonant converter. In this paper, a novel reconfigurable LLC resonant topology is customized to mitigate this design challenge. On the primary side, six MOSFETs are utilized to formulate a reconfigurable inverter bridge. This bridge exhibits six different operation modes. Those operation modes enable a wide voltage gain range with a narrow switching frequency span. Furthermore, two asymmetric LLC resonant tanks with different turns ratios are deployed to further squeeze the switching frequency range. Zero-voltage switching is realized on all MOSFETs during all six modes. The conduction loss is reduced due to the reduced circulating current in each resonant tank. High efficiency and narrow frequency range over a wide voltage gain range are well achieved. The operating principles, design considerations, voltage gain analysis, and key experimental results are presented. A 1 kW laboratory prototype is built to verify the effectiveness of the proposed concept. The prototype converts a 390V input to an 80-420V output. A 97.42% peak efficiency is reported.

Keywords—asymmetric resonant tanks, LLC, reconfigurable bridge, wide gain range

I. INTRODUCTION

Wide input voltage or output voltage range converter is featured in many applications including renewable energy systems, electrical vehicle chargers, telecommunication power supplies and LED drivers. LLC resonant converter is very suitable for those scenarios due to its considerable advantages including zero voltages switching (ZVS), low EMI, and small components count [1]. However, in order to obtain a wide voltage gain range, LLC resonant converters need to operate over a wide switching frequency \( f_s \) range. This means \( f_s \) needs to deviate from the resonant frequency \( f_r \). It might incur issues such as the failure of soft switching and a jeopardized conversion efficiency.

To achieve a wide voltage regulation range and narrow \( f_s \) range simultaneously, different control strategies have been studied in the literature [2]–[5]. In [2], an asymmetric pulse width modulation (PWM) scheme is proposed to boost the voltage gain during hold-up time. Under normal operation, the LLC resonant converter is modulated by frequency. During the hold-up time, \( f_s \) goes down to the minimum \( f_s \) and the modulation method switches to asymmetric PWM. In [3], a control scheme which hybridizes frequency and phase shift modulations is proposed. The normalized voltage gain range below unity is enabled by phase shift modulation. Thus, the voltage gain is extended with a narrower frequency range.

Besides, modifications in circuit structure also help to extend the voltage gain range with narrowed \( f_s \) range [6]–[9]. In [6], a parallel loaded resonant converter with two resonant inductors \( L_r \) and one resonant capacitor \( C_r \) is proposed. This structure enables both soft switching and narrow \( f_s \) span. In [8], an LLC-LC resonant converter is introduced. The modified resonant tank helps to achieve wide voltage gain and load ranges with small inductance ratio \( m \) and narrow \( f_s \) window.

Adding control variables on the secondary side is another feasible method to narrow down the span of \( f_s \) [10]–[15]. In [10], a half-bridge LLC resonant converter with hybrid rectifier is proposed for PEV battery charging. By maintaining secondary-side MOSFETs in constant ON/OFF state, secondary rectifier operates as full bridge rectifier in low gain mode and voltage doubler rectifier in high gain mode, respectively. Similarly, a semiactive variable structure rectifier is proposed in [11]. With the help of active switches on the secondary side, the LLC resonant converter could operate in voltage-doubler-rectifier (VDR) mode and voltage-quadrupler-rectifier (VQR) mode. This mode transition effectively extends the voltage gain range. In [13], a semi-active structure and hybrid control method is introduced. The MOSFETs on the secondary side
operate as synchronous rectifier in normal state. During the hold-up time, the phase shift between the primary and secondary sides could boost the voltage gain.

Voltage modulation modes transition in the primary side could also squeeze $f_s$, span [16]-[21]. A topology morphing method is proposed in [18]. With the morphing method, the converter exhibits two gain curves when primary side bridge operates as full bridge or half bridge. In [17], the similar morphing method is applied to the converter with two resonant tanks. One resonant tank always operates as a half bridge while the other resonant tank could operate in full bridge mode or half bridge mode. Hence, the voltage gain in high gain mode is 1.5 time of that in low gain mode.

To achieve a wide voltage gain range and a narrow $f_s$ range, this paper proposes a novel asymmetric dual LLC resonant converter. Due to the mode transition on the primary side, the converter could operate in a narrow $f_s$ span near $f_s$ and achieve an ultra wide voltage gain range. The proposed converter’s advantages include: a) wide voltage modulation range with narrow $f_s$ span; b) all MOSFETs’ soft switching over the whole load range; c) smaller resonant current ($i_r$) and conduction loss; d) high efficiency.

This paper is organized as follows. In Section II, the topology introduced and its operation principles are explained. Design considerations and voltage gain analysis are presented in Section III. The experimental results are shown in Section IV. Section V concludes this work.

II. TOPOLOGY DESCRIPTION AND OPERATION PRINCIPLES

The schematic of the converter is plotted in Fig. 1. A reconfigurable bridge with six MOSFETs is placed on the primary side. Two asymmetric LLC resonant tanks, namely RT1 and RT2, are connected with this inverter bridge. Since the turns ratios are different, their voltage gains are also different. Two full-wave rectifiers are connected in series on the secondary side. By maintaining the MOSFETs in constant ON/OFF states, RT1 and RT2 could operate as full bridge or half bridge mode. Hence, six different operating modes with different voltage gain curves are obtained: RT2 half-bridge (2H), RT1 half-bridge (1H), dual half-bridge (DH), RT1 half-bridge RT2 full-bridge (1H2F), RT1 full-bridge RT2 half-bridge (1F2H), and dual full-bridge (DF). Fig. 2 depicts the primary-side switch combinations at six different modes respectively. Fig. 3 demonstrates the corresponding steady-state waveforms.

Mode 1: In this mode (2H), Q1,3 are constantly OFF, Q2,5 are constantly ON. Q2 and Q6 are controlled with complementary driving signals with certain deadband. Mode 1 is illustrated in Fig. 2 (a) and Fig. 3 (a). The input of RT2, $v_{bc}$, is a two-level square wave (0 to $V_{in}$). $v_{bc}$’s RMS value is $V_{in}$/2. Due to point a is floating, $v_{bc}$’s RMS value is 0. This means only RT2 operates in half-bridge mode while RT1 is idle.

Mode 2: The operational principles of this mode (1H) is plotted in Fig. 2 (b) and Fig. 3 (b). In this mode, Q1,4 are always ON and Q2,5 are always OFF. Q2 and Q6 are controlled with complementary driving signals with certain deadband. Therefore, the input of RT1, $v_{ab}$, is a two-level square wave (0 to $V_{in}$). $v_{ab}$’s RMS value is $V_{in}$/2. Since point c is floating, $v_{ab}$’s RMS value is 0. This indicates that only RT1 operates as half bridge and RT2 is idle. It should be noted that voltage gains in Mode 1 and Mode 2 are different. This is because RT1 and RT2 are asymmetric.

Mode 3: In this mode (DH), Q1,5,6 are constantly ON while Q3 is constantly OFF as shown in Fig. 2 (c) and Fig. 3 (c). Q2 and Q4 are controlled with complementary driving signals with certain deadband. Hence, $v_{ab}$ and $v_{bc}$ are two identical square waves, (0 to $V_{in}$). RT1 and RT2 both operate in half-bridge mode. The RMS values of the two resonant tanks’ inputs are $V_{in}$/2.

Mode 4: In this mode (1H2F mode), Q1,6 are constantly ON. Q3,5 and Q2,4 are controlled with complementary driving signals with certain deadband. Mode 4 is illustrated in Fig. 2 (d) and Fig. 3 (d). $v_{ab}$ and $v_{bc}$ are two different two-level square waves. The voltage levels in those square waves are 0 to $V_{in}$ and -$V_{in}$ to $V_{in}$, respectively. This indicates RT1 operates in half-bridge mode and RT2 operates in full-bridge mode. The RMS values of RT1 and RT2’s input voltages are $V_{in}$/2 and $V_{in}$/2, respectively.

Mode 5: The operational principles of this mode (1F2H) is plotted in Fig. 2 (e) and Fig. 3 (e). In this mode, Q2,6 are constantly ON. Q1,4 and Q2,3 are controlled with complementary driving signals with certain deadband. Therefore, $v_{ab}$ and $v_{bc}$ are two different two-level square waves. The voltage levels of the square waves are -$V_{in}$ to $V_{in}$ and 0 to $V_{in}$, respectively. This indicates RT1 operates in full-bridge mode and RT2 operates in half-bridge mode. The RMS values of RT1 and RT2’s input voltages are $V_{in}$ and $V_{in}$/2, respectively. Similarly, due to two resonant tanks’ parameters are different, voltage gains in Mode 4 and Mode 5 are different.
Resonant tank’s parameters, including turns ratios in RT1 and RT2 (n1, n2) should meet two basic requirements: a) provide continuous output/input voltage range, largest voltage gain in the low voltage gain curve should be larger than the smallest voltage gain in the adjacent higher voltage gain curve; b) obtain narrower f1 range, the ratio between the largest voltage gain and smallest voltage gain in the same voltage gain curve should be minimized.

From Table I and those two basic requirements introduced above, the optimum w could be obtained.

\[
w_0 = \frac{1 + \sqrt{5}}{2} = 1.6
\]

Hence, n2 is designed to be 1.6n1. Then, using the conventional first harmonics analysis (FHA) method, Q, m, Lm, Lr and Cr in two resonant tanks could be obtained.

### B. Voltage gains

The equivalent circuit model of the proposed topology based on FHA is shown in Fig. 4. \(R_e\) is expressed and resonant tank’s voltage gain, \(G_{LLC}\), is derived.

\[
G_{LLC} = \frac{mf^2}{\sqrt{[(m+1)f_1^2 - 1]^2 + m^2Q^2 f_1^2(f_1^2 + 1)^2}}
\]
The six voltage modulation modes facilitate a narrow and a wide voltage gain range. Furthermore, the efficiency is

\[
G_0 = \frac{v_\text{ab, rms}}{v_\text{bc, rms}}
\]

w = n_2/n_1.

<table>
<thead>
<tr>
<th>Modes</th>
<th>Tanks</th>
<th>Gain</th>
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<tbody>
<tr>
<td>Mode 1</td>
<td>1</td>
<td>V_\text{ab}/2</td>
</tr>
<tr>
<td>Mode 2</td>
<td>1</td>
<td>V_\text{bc}/2</td>
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<tr>
<td>Mode 3</td>
<td>2</td>
<td>V_\text{ab}/2</td>
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<tr>
<td>Mode 4</td>
<td>2</td>
<td>V_\text{bc}/2</td>
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<tr>
<td>Mode 5</td>
<td>2</td>
<td>V_\text{ab}/2</td>
</tr>
<tr>
<td>Mode 6</td>
<td>2</td>
<td>V_\text{bc}/2</td>
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Table II: Normalized Voltage Gains in Six Modes

To normalize the voltage gains in different modes, the proposed converter’s voltage gain in Mode 1 is defined as \(G_0\). Thus, the voltage gains in modes 1, 2, 3, 4, and 5 are derived as 1.6\(G_0\), 2.6\(G_0\), 3.6\(G_0\), 4.2\(G_0\), and 5.2\(G_0\), as shown in Table II.

Therefore, six different voltage gain curves versus normalized \(f_s\) are demonstrated in Fig. 5. It should be noted that the six voltage modulation modes facilitate a narrow \(f_s\) span and a wide voltage gain range. Furthermore, the efficiency is

\[
Q = \frac{\sqrt{L_0/C_0}}{L_0}
\]

\[
m = L_0/L_r
\]

\[
f_s = f_1/f_2
\]

\[
R_s = \frac{8\mu_0n^2}{\pi}
\]

Considering \(Q_1 = Q_2\), \(m_1 = m_2\), \(G_{\text{LLC1}} = G_{\text{LLC2}}\), the total voltage gain of the converter depends on the number of resonant tanks, \(V_\text{ab, rms}, V_\text{bc, rms}\), and \(n\).

\[
V_o = v_{\text{ab, rms}} G_{\text{LLC1}} n_1 + v_{\text{bc, rms}} G_{\text{LLC2}} n_2
\]

Fig. 4. Equivalent FHA circuit model.

Fig. 5. Voltage gain curves versus normalized \(f_s\).

IV. Experimental Verification

A 1 kW experimental prototype is built to validate the effectiveness of this concept. Its circuit design specifications are listed in Table III.

The steady-state waveforms in modes 3, 4 and 6 are shown in Figs. 6, 7 and 8, respectively. Fig. 6 demonstrates the converter’s steady-state waveforms when it operates in Mode 3. It should be noted that \(V_\text{ab}\) and \(V_\text{bc}\) are two identical two-level (0V to 390V) square waves. Therefore, RT1 and RT2 both operate in half-bridge mode. Also, since \(n_1\) is smaller than \(n_2\), \(i_{11}\) is larger than \(i_{21}\). Therefore, RT1 delivers more power, and RT2 deliver less power to the load. This experimental result coincides with the theoretical analysis.

Fig. 7 demonstrates converter’s steady-state waveforms when it operates in Mode 4. It can be seen that \(V_\text{ab}\) and \(V_\text{bc}\) are two different two-level square waves. Their voltage range is 0V to 390V, and -390V to 390V, respectively. These different square waves validate that two resonant tanks operate in half-bridge mode (RT1) and full-bridge mode (RT2), respectively. Hence, \(i_{11}\) is smaller than \(i_{21}\), and RT2 delivers higher power than RT1. As shown in Fig. 8, \(V_\text{ab}\) and \(V_\text{bc}\) are two identical two-level square waves. The same voltage range, from -390V to 390V indicates that two resonant tanks both operate in full-bridge mode. \(i_{11}\) is unequal to \(i_{21}\). This indicates two resonant tanks deliver different power to the load due to \(n_1\) and \(n_2\) are different.
The steady-state waveforms at 10% load, 50% load, and 100% load are captured in Figs. 9, 10 and 11. Fig. 9 shows the circuit operation in Mode 2 at 10% load. ZVS turning ON of $Q_6$ is achieved. $i_{r1}$ is much larger than $i_{r2}$. This is because RT1 operates at half-bridge mode and RT2 delivers no power to the load.

Fig. 10 demonstrates the circuit operation in Mode 5 at 50% load. It can be seen that $v_{ds4}$ is discharged to zero before $Q_4$ is turned on. Hence, switches are turned on with ZVS at 50% load. Steady-state waveforms in Mode 6 at 100% load is captured in Fig. 11. As shown, $v_{ds6}$ drops to zero before the conduction of $Q_6$. Since $n_1$ is smaller than $n_2$, $i_{r1}$ is larger than $i_{r2}$. This indicates two resonant tanks deliver different power to the load. ZVS is achieved for all MOSFETs over very wide load range, from 10% load to 100% load.

Conversion efficiency is measured in the laboratory with 390V input and 80-420V output. The operation mode changes from 1 to 6 as $V_o$ increases, accordingly. Since the squeezed $f_s$ span is close to $f_s$, the overall efficiency performance is enhanced with a 97.42% peak efficiency. It should be noted that efficiency in Mode 5 is lower than that in Mode 4. The main reason is the different resonant parameters in RT1 and RT2. $L_{m1}$ is smaller than $L_{m2}$. Hence, circulating current in RT1 is larger than that in RT2. The power loss will be larger when RT1 delivers more power to the load. Furthermore, the efficiency data in modes 1 and 2 are much smaller than those in the other four modes. The main reasons are a) power levels
in modes 1 and 2 are small, and b) the idle resonant tanks incur extra power loss due to the series-connected secondary side.

V. CONCLUSION

In this paper, a novel asymmetric dual LLC resonant converter is proposed for wide voltage gain applications. Due to the six voltage modulation modes, this converter could provide wide voltage gain range with squeezed \( f_r \) range near \( f_s \). Furthermore, this converter achieves a) all MOSFETs' ZVS during over wide load range, b) reduced conduction loss, and c) high efficiency. The proposed topology, operation principles, design considerations, circuit model and voltage gains analysis are presented. To validate the effectiveness of this topology and theoretical analysis, a 1 kW laboratory prototype with 390V input, 80–420V output, 100 kHz \( f_s \), and 70–130 kHz \( f_r \) is designed. The overall experimental performance is good and peak efficiency is 97.42%. This proves that this topology is suitable in wide input/output voltage range application including renewable energy systems, and electrical vehicle onboard chargers.

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