An Adjustable Turns Ratio Transformer Based LLC Converter for Deeply-depleted PEV Charging Applications

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Abstract—The charging of deeply-depleted plug-in electric vehicle (PEV) requires the dc/dc converter to have an ultra-wide output voltage range. However, the conventional LLC converter is unsuitable for such applications. To address this issue, a novel LLC converter with an adjustable turns ratio transformer is proposed. The transformer has three different effective turns ratios, which correspond to different output voltage ranges. Thus, three different operation modes are derived and the switching frequency range in each mode can be squeezed close to the resonant frequency. Therefore, the loss of the converter is reduced and the efficiency is improved. The operational principles, design considerations, and mode transition are analyzed. A 1 kW prototype that converts 390V input into a 126-420 V wide output is designed and tested to validate the concept. A smooth mode transition is achieved without transient issues.

Keywords—high efficiency, LLC converter, PEV charger, reconfigurable transformer, wide voltage range;

I. INTRODUCTION

LLC topology is a prevalent isolated dc/dc converter due to its soft-switching characteristic and simple frequency modulation. The operation of the LLC converter is deemed as optimal when its switching frequency is matched with the resonant frequency. However, in applications such as the plug in electric vehicle (PEV) charging, its output voltage needs to be adapted to a wide voltage range (typically 250 - 420 V, and even lower for deeply discharged scenarios) [1], [2]. Using the LLC topology to achieve such a wide voltage range, a small magnetizing inductance, and wide frequency modulation range are required. Both would jeopardize the system performance [3].

To cope with this issue, a series of solutions are investigated. In [4], adaptive dc-link voltage control is proposed to narrow down the switch frequency range of LLC converter. In [5]–[9], pulse width modulation (PWM) is adopted to regulate the resonant tank voltage, and a wide output voltage range is obtained. However, these solutions can’t provide enough voltage gain when the battery pack is in a deep depletion state. In [10], two resonant tanks and two transformers are stacked together to obtain higher voltage gain. However, the additional resonant tank and transformer degrade the system power density. Moreover, the mode transition comes with transient issues in the output voltage. In [11], high reactive power caused by the circulating current in an additional transformer jeopardizes the converter efficiency. In [12], [13], two modified LLC converters with two split resonant branches are proposed. However, there are still two transformers and the mode transition is discontinuous with severe dynamics. Those above-mentioned solutions either add complexity to the topology or fail to achieve a sufficient voltage range. In order to simplify the topology, an LLC converter with reconfigurable mode modulation is proposed in [14]. However, the mode transition is a gradual process which increases the complexity of transformation control.

In order to realize a wide voltage gain range with reduced control complexity, a series of efforts have been made in literature [15]–[17]. In [15], a parallel resonant converter is introduced to change the gain curve. However, the additional conduction loss in the parallel load is high. In [16], a CPT converter with dual transformer architecture is proposed to change the gain curve. However, additional transformers and resonant components are needed. In [17], a reconfigurable transformer LLC converter is proposed to realize multiple gain curves. However, the copper utilization in the transformer is low.

In this paper, an adjustable turns ratio transformer-based LLC resonant converter is proposed. An ultra-wide voltage gain range can be achieved through the transition of three effective transformer turns ratios. Soft switching can be implemented in an extended range over three different modes. In every mode, there are only four high frequency switches. Similar to the conventional LLC topology, singular frequency modulation is adopted. Each mode is assigned with a specific transformer turns ratio, which is customized for an specific voltage range. The mode transition can be done instantaneously without transient issues. Therefore, compared with the state of the art, the proposed converter can obtain higher voltage gain in a narrower frequency range.

II. TOPOLOGY DESCRIPTION AND OPERATION PRINCIPLES

A. Topology Description

Fig. 1 shows the schematic of the proposed LLC converter with an adjustable turns ratio transformer. The primary side is a full-bridge inverter with four MOSFETs. The resonant tank
can be configured as four-quadrant switches, the transformer secondary side facilitates mode transition. Each switch consists of two back-to-back diodes. Two four-quadrant switches are used to effectively extend the voltage gain of the converter.

**B. Operation Principles**

Fig. 2 shows the three modes of the proposed converter. When $S_1$ is turned ON and $S_2$ is turned off, Mode I is enabled. The effective transformer turn ratio is $n_2 : n_1$. When $S_2$ is turned on and $S_1$ is turned off, Mode II is enabled. The effective transformer turn ratio is $n_3 : n_1$. When both $S_1$ and $S_2$ are turned on, Mode III is enabled. The effective transformer turn ratio is $(n_2+n_3) : n_1$. With the above three modes, the output voltage range can be divided into three segments: $[V_{a1}, V_{a0})$, $[V_{a1}, V_{a2})$ and $[V_{a2}, V_{b0})$. $V_{a1}$ and $V_{a2}$ are the threshold voltage which triggers the mode transition.

The steady-state waveforms with $V_a = V_{a0}$ in modes I and II are plotted in Fig. 3. Fig. 3 (a) demonstrates the circuit operation in Mode I. $f_c$ is lower than $f_s$. The maximum output voltage in Mode I is achieved. With further increase of the output voltage, the mode transition from Mode I to Mode II is triggered. Fig. 3 (b) demonstrates the circuit operation in Mode II. $f_c$ is higher than $f_s$. The minimum output voltage in Mode II is achieved. The circuit operation in the vicinity of $V_{a2}$ is similar.

In every operation mode, $f_c$ is the only control variable. The voltage gain decreases as $f_c$ increases. When $f_c$ equals $f_s$, the voltage gain equals the transformer turns ratio. When $f_c$ is lower than $f_s$, the resonant current intersects with the magnetizing current before the primary gate signal turns off as shown in Fig. 3 (a) at $t_0$. As a result, the secondary-side diodes operate in discontinuous mode. This leads to a naturally zero-current-switching (ZCS) turning off of diodes on the secondary side. On the primary side, the circulating current as shown in the shaded area of Fig. 3 (a) increases with the decrease of $f_c$. Then, the root-mean-square (RMS) value of the resonant inductor current $i_{Lr}$ increases with the decrease of $f_c$. Therefore, a tradeoff exists between the RMS current on the primary side and the turning off loss on the secondary side.

It should be noted that if $f_c$ is substantially lower than $f_s$, the primary-side MOSFETs may lose ZVS. Therefore, the lower bound of $f_c$, which determines the peak voltage gain in this mode, should be designed to be close to $f_s$. When $f_c$ is higher than $f_s$, the resonant current intersects with the magnetizing current after the primary gate signal turns off as shown in Fig. 3 (a) at $t_2$. As a result, the secondary side diodes lose ZVS turning-off feature. Moreover, the circulating current decreases with the increase of $f_c$. Therefore, a tradeoff exists between the turn-off loss on the secondary side and the RMS current on the primary side. With a substantial increase of $f_c$, the circulating current is sufficiently small. It barely affects the primary side current. However, the MOSFETs turning-off current and diodes turning-off $di/dt$ both increase, which contributed to an increased switching loss. Thus, the higher bound of $f_c$, which determines the valley voltage gain in the mode, should be designed close to $f_s$ as well.

**C. Mode Transition**

Mode transition is important during the charging process as a harsh transition incurs large transient, which may degrade the battery lifetime. When the battery voltage reaches $V_{a1}$, the mode transition is triggered to switch the circuit operation
from Mode I to Mode II. When the battery voltage reaches $V_{m}$, the mode transition is triggered to switch the circuit operation from Mode II to Mode III.

During the transition from Mode I to Mode II, $f_1$ steps from the lowest switching frequency in Mode I ($f_{L1}$) to the highest switching frequency in Mode II ($f_{L2}$). On the secondary side, $S_5$ is turned off and $S_6$ is turned on. The steady-state output voltage before and after the mode transition is unchanged. In addition, there is no net change in the dc bias of the resonant capacitor voltage. Therefore, a smooth transition from Mode I to Mode II can be achieved.

The mode transition from Mode II to Mode III is similar. $f_1$ steps from the lowest switching frequency in Mode II ($f_{L2}$) to the highest switching frequency in Mode II ($f_{L3}$). On the secondary side, $S_5$ is turned on. The steady-state output voltage before and after the mode transition is unchanged. In addition, there is no net change in the dc bias of the resonant capacitor voltage. Therefore, a smooth transition from Mode II to Mode III can be achieved.

III. DESIGN CONSIDERATIONS

A. Voltage Gain

Fig. 4 is the equivalent circuit of the proposed converter, where

$$
\begin{align*}
\text{In mode I : } n &= n_2 / n_1 \\
\text{In mode II : } n &= n_1 / n_2 \\
\text{In mode III : } n &= (n_2 + n_3) / n_1
\end{align*}
$$

(1)

The total required voltage gain range is divided into three segments. Thus, the frequency range of each segment can be squeezed and designed to be close to $f_o$. The first harmonics approximation (FHA) method can be employed to establish the circuit model. Thus, the voltage gain of the proposed converter is,

$$
M_g = \frac{1}{\sqrt{[1 + (1/L_m - 1/L_r) f_o^2]^2 + Q_o^2 (f_o - 1/f_o)^2}}
$$

(2)

where,

$$
L_r = \frac{L_0 Q_o}{f_o R_m}, f_o = \frac{f_o}{f_1}, n_0 = \frac{Q_o}{\sqrt{2} \pi f_1 R_m}, n_1 = \frac{1}{2 \pi f_1 C}
$$

(3)

B. Parameter Design

The voltage gain is a function of $L_m$ and transformer turns ratio. Fig. 5 shows the voltage gain with different $L_m$ when the transformer turns ratio $n = 1$ and $Q_o = 0.7$. In order to obtain a better gain curve, $L_m = 3.5$ is selected in this paper. Then the voltage gain versus normalized $f_o$ with different transformer turns ratios is shown in Fig. 6. In order to obtain a continuous voltage gain in a narrower frequency range, two turns ratios of the transformer can be selected as $n_2/n_1 = 0.35$, $n_3/n_1 = 0.65$, then the turns ratio in mode III is $(n_2 + n_3) / n_1 = 1.0$.

In order to reduce the volume of magnetic elements without causing particularly large switching losses, the resonant frequency is selected to be 140 kHz. To ensure ZVS operation, from [18], the upper limit of magnetizing inductor $L_m$ can be derived as,

$$
L_m < \frac{V_{m} C_{eq}}{16 C_{m}}
$$

(4)

where $C_{eq}$ is the equivalent output capacitance of the MOSFET, $t_{dead}$ is the deadband, and $T_{min}$ is the minimum switching period. To reduce the MOSFET turning off current and circulating current, $L_m$ should be sufficiently large. On the other hand, $L_m$ affects the voltage gain range. In the proposed method, the output voltage range is divided into three segments. It means a higher $L_m$ can be chosen to reduce the switching loss and conduction loss. In this paper, $L_m$ is selected to be 350 $\mu$H. Then, $L_m$ and $C$ are calculated as 100 $\mu$H and 12.8 $nF$, respectively.

Fig. 7 shows the voltage gain versus normalized $f_o$ of the proposed converter. The red line represents the operation region. Threshold voltages and frequency are indicated in the figure.
C. Transformer Design

Fig. 8 shows the configuration of the transformer windings and the equivalent circuit. As shown the secondary side is intermediately tapped with three-port. The resonant inductor is integrated into the transformer to reduce the volume of the converter.

D. Voltage stress

When the topology operates in Mode I, \( S_5 \) is on and \( S_6 \) is off. The equivalent circuit on the secondary side is shown in Fig. 9.

When the current of the transformer \( i_{i1} \) is positive, the terminal voltage of the diode bridge intermediate point and the transformer is,

\[
V_A = 0, V_E = V_A, V_C = 0, V_{s3} = n_2 V_A / n_1
\]

Therefore, the maximum voltage on \( S_6 \) is,

\[
V_{\text{max}} = V_{s3} - V_C = -n_2 V_A / n_1
\]

When the current of the transformer \( i_{i1} \) is negative, similar result can be obtained. Therefore, the voltage stress on \( S_6 \) is,

\[
V_{\text{stress}} = n_2 V_A / n_1
\]

Similarly, the voltage stress on \( S_5 \) during Mode II can be obtained as,

\[
V_{\text{stress}} = n_3 V_A / n_2
\]

When the converter operates in Mode III, both \( S_5 \) and \( S_6 \) are on. Their voltage drop is close to zero. Moreover, since the power stored in the leakage inductance can be released during the mode transition, no extra voltage stress is generated during the mode transition.

IV. EXPERIMENTAL RESULT

To verify the proposed concept, a 1 kW converter prototype is designed and tested experimentally. The input voltage is 390 V and the output voltage is 126 – 420 V. The transformer turns ratio is \( n_2/n_1 = 0.35 \), \( n_3/n_1 = 0.65 \). The leakage inductance is 100 \( \mu \)H, the magnetic inductance is 354 \( \mu \)H, and the resonant capacitance is 12.8 nF (resonant frequency \( f_r = 140 \) kHz).

![Fig. 8. Transformer winding schematic and its equivalent circuit.](image)

![Fig. 9. Equivalent circuit on the secondary side during Mode I.](image)

![Fig. 10. Experiment steady state waveforms of Mode I when \( f_r = 220 \) kHz, \( V_o = 126 \) V.](image)

![Fig. 11. Experimental steady state waveforms of Mode II when \( f_r = 140 \) kHz, \( V_o = 251 \) V.](image)

![Fig. 12. Experimental steady state waveforms of Mode III when \( f_r = 130 \) kHz, \( V_o = 423 \) V.](image)

![Fig. 13. Experimental efficiency versus the output voltage in CC charging stage.](image)
converter. Fig. 12 shows the steady-state waveforms of the converter when the output voltage is 423 V in Mode III. From the experiment result, ZVS can be achieved in almost all of the operation points.

Fig. 13 shows the efficiency versus the output voltage in constant current (CC) charging stage of the designed prototype. In every mode, when $f_i$ is close to $f_o$, a high efficiency is obtained. When $f_i$ deviates from $f_o$, the efficiency slowly drops. Therefore, the efficiency intersection occurs between two adjacent modes. This can be marked as the mode transition point as shown in Fig. 13. In Fig. 14, the mode transition from Mode II to Mode III is captured. The mode transition occurs instantaneously without transient issues.

Fig. 15 shows the curve of efficiency versus output power in the constant voltage (CV) charging stage. As shown, the converter maintains high overall efficiency and 97.18% peak efficiency. The large magnetizing inductance leads to low circulating current. This helps to improve light load efficiency.

V. CONCLUSION

In this paper, an adjustable turns ratio transformer-based LLC Converter for PEV charging applications is proposed. The converter can be operated with three transformer turns ratios. Therefore, the proposed topology has the following advantages: 1) the frequency range is squeezed in each mode, which allows the LLC converters to work with $f_i$ close to $f_o$; and 2) design of resonance network parameters is less constrained by the voltage gain requirement.

Soft switching can be implemented over a wide range. A very wide output voltage range (126 – 420 V) can be achieved. The transformer is featured with high copper utilization. Singular frequency modulation is adopted, which simplify the circuit control. A 1 kW converter prototype is established to verify the proposed concept. The designed prototype demonstrates good efficiency performance over this wide output voltage range. 97.18% peak efficiency is reported. A smooth mode transition is realized instantaneously without transient issues.

REFERENCES