An LLC and LCL-T Resonant Tanks Based Topology for Battery Charger Application

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Abstract—To achieve the constant current (CC) and constant voltage (CV) charge of the lithium battery, the traditional LLC resonant converter requires the switching frequency varies in a wide range, which brings difficulty to the magnetic components design, and the system efficiency would also be degraded. In this article, a novel topology based on LLC and LCL-T resonant tanks is proposed to reduce the range of operating switching frequency. During the CC charge state, the proposed converter is operating with the LCL-T resonant tank, and it can be regarded as a current source, which provides constant charging current to the battery. During the CV charge state, the LCL-T resonant tank is bypassed and the structure of the proposed converter is modified to a traditional LLC resonant converter, and it is functioning as a CV source. Owing to the high accuracy of the CC and voltage sources, the required operating switching frequency range can be significantly reduced when compared with traditional LLC approaches. Operational principles and design guidelines for the proposed converter are described. Experiment and simulation results from a 180 W prototype are provided to validate the theoretical analysis.

Index Terms—Battery charger, LCL-T resonant tank, LLC resonant converter.

I. INTRODUCTION

ELECTRIC vehicles (EVs) have drawn much attention due to their environmentally friendly characteristic. For EVs, the on-board battery charger is one of the most important technologies [1]. Typically, a two-stage system, including an AC/DC power factor correction (PFC) converter and an isolated DC/DC converter, is required. Compared with pulse width modulation (PWM) converters, LLC resonant converter has the following advantages: 1) Soft switching capability from zero to full load range, low primary switch turn off current and switching loss. 2) High efficiency is achieved at high input

voltage. 3) Voltage stress for secondary rectifier is low since there is no filter inductor, and the stress can be clamped by the output voltage. 4) The resonant inductor and transformer can be integrated into one magnetic component, and the leakage inductance can be utilized [2]. Therefore, the DC/DC power conversion is frequently realized by adopting LLC resonant converter [2]–[22].

Typically, for the lithium battery, the charging profile can be described as follows: First, the constant current (CC) charge state is used to achieve the normal operation voltage. Then, the constant voltage (CV) charge state is used to fully charge the battery. For traditional LLC resonant converters, to cover the wide voltage gain range, the switching frequency must vary in a wide range. For example, in [2], the required switching frequency range is from 85 kHz to 200 kHz, which leads to the following problems: 1) poor electronic-magnetic-interference (EMI) performance; 2) challenge for optimization of resonant inductor and transformer; 3) high circulating current and corresponding conduction losses [3]. Furthermore, the above switching frequency operation region is utilized to extend the voltage gain range, and the soft switching for rectifier is lost.

Based on the existing literature, there are several ways to achieve reduction of the switching frequency range: 1) hybrid control strategies [12]–[15]; 2) magnetic control [16]–[19] or switch-controlled capacitor (SCC) [20], [21]; 3) modified structure [22], [23]; 4) inherent characteristic of LLC converter [24]. Phase shift is another control freedom for LLC resonant converter. Therefore, by combining phase shift angle and the switching frequency, some hybrid control strategies have been proposed and analyzed [12]–[15]. To fully utilize the advantage of the converter, the optimization is necessary for hybrid control strategies due to the infinite solutions. The digital control method is necessary due to the complexity of the system control strategy, which increases the system cost.

LLC resonant converter with resonant frequency modulation (RFM) can achieve the fixed frequency operation. By adjusting the resonant inductor value, the output of the LLC resonant converter can be regulated and the switching frequency can be kept constant [16]–[19]. However, for magnetic control, the problems exist as follows: 1) Operating a variable inductor requires additional control current and control windings, which will decrease the system efficiency. 2) For the variable inductor based LLC resonant converter, the resonant inductor must be separated from the magnetizing inductor, which decreases power density. 3) The range of the variable inductor cannot be

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Fig. 1. Proposed converter for the battery charger application.

too wide due to material limitations, so it is difficult to adopt magnetic control over wide range operation. Another way to achieve the fixed frequency operation is the SCC LLC, where the resonant capacitor value can be modified to regulate the output voltage [20], [21]. However, the resonant current detection and synchronization circuit is necessary for SCC LLC, which increases the complexity.

To address the problems caused by magnetic control and SCC, some novel topologies based on the modifications of traditional topology are proposed [22], [23]. Although a wider operating range can be covered, this method suffers from the following drawbacks: 1) complicated structure; 2) large component count, which decreases its power density; 3) complicated control strategy.

Utilization of the inherent characteristics of the converter can also reduce the frequency range. In [24], the topology based on a dual full-bridge LLC resonant converter is proposed. By utilizing the CC and CV operation of the converter, the different charger mode for battery is achieved. However, the accuracy of fixed output current mode still needs to be examined. The details regarding the accuracy of different resonant tank characteristics are presented in this paper.

LCL-T resonant tank has been previously investigated and analyzed. In these papers [25]–[27], the operational analysis for LCL-T resonant tank is described and its CC operation is analyzed and validated. This characteristic has been utilized as a current source to drive LEDs. Inspired by the fixed output voltage characteristic of the LLC resonant tank and the fixed output current characteristic of the LCL-T resonant tank, a novel topology based on these two resonant tanks, as shown in Fig. 1, is proposed in this paper for battery charger applications. Due to the high accuracy of these two characteristics, a narrow switching frequency operating range is achieved. The magnetic core size reduction of resonant inductor and transformer makes the size increase of additional LCL-T resonant tank insignificant.

This paper begins with brief descriptions of the fixed output voltage and current characteristics as well as the operational analysis of the converter in Section II. The design guidelines are investigated in Section III based on a particular specification. In addition, the accuracies of different resonant tank characteristics are investigated and compared. Then, a 180 W experimental prototype was built, and the results are presented to validate the theoretical analysis. The paper closes with a summary and conclusion.

Fig. 2. Fixed output voltage mode. (a) Equivalent circuit and (b) FHA model.

II. ANALYSIS OF THE PROPOSED CONVERTER

Two operation modes exist for the proposed topology: namely, fixed output voltage or LLC operation mode, and fixed output current or LCL-T operation mode, the operational principles of these two modes are described in this section.

A. Fixed Output Voltage or LLC Operation Mode

For the proposed converter, when the mode selection switch S_{mode1} is ON, and S_{mode2} is OFF. The converter is operating in fixed output voltage mode. The equivalent circuit of the converter is shown in Fig. 2(a), which is a simple half-bridge LLC resonant converter. The model based on fundamental harmonic analysis (FHA) of the LLC resonant converter is obtained as shown in Fig. 2(b). In Fig. 2(b), V_{ac} is the fundamental harmonic of the half-bridge inverter output voltage, and V_{p1} is the fundamental harmonic of the transformer voltage.

In an LLC resonant converter, there exist two resonant frequencies, and they can be expressed as (1) and (2), respectively.

$$f_r = \frac{1}{2\pi\sqrt{L_r C_r}} \quad (1)$$

$$f_m = \frac{1}{2\pi\sqrt{(L_r + L_m)C_r}} \quad (2)$$

Based on Fig. 2(b), the output voltage is derived as

$$V_{p1}(s) = V_{ac} \frac{\frac{L_m s R_{ac}}{L_m s + R_{ac}}}{L_r s + \frac{1}{C_r s} + \frac{L_m s R_{ac}}{L_m s + R_{ac}}} \quad (3)$$

$$R_{ac} = \frac{8n^2}{\pi} R_L \quad (4)$$

$$V_{p1} = \frac{2\sqrt{2}n}{\pi} V_o \quad (5)$$

$$V_{ac} = \frac{\sqrt{2}}{\pi} V_{in} \quad (6)$$

Fig. 3. Fixed output voltage operation mode. (a) equivalent circuit; (b) FHA model; (c) simplified model.

The term $(1/C_r s + L_r s)$ equals 0 when $f_s = f_r$. Then, (3) can be simplified to

$$V_o(s) = \frac{V_{in}}{2n} \quad (7)$$

From (7), the fixed output voltage operation is achieved. In addition, the input impedance of the converter can be expressed as

$$Z_{inl}(s) = \frac{L_m s R_{ac}}{L_m s + R_{ac}} \quad (8)$$

From (8), the input impedance during this operation mode is inductive, which is beneficial for the zero-voltage switching (ZVS) operation of the switches.

The output current expression can be simply derived from (3) as

$$I_o(s) = n V_{ac} \frac{1}{\left(L_r s + \frac{1}{C_r s}\right) + \frac{R_{ac}}{L_m s} \left(L_m s + L_r s + \frac{1}{C_r s}\right)} \quad (9)$$

The term $(L_m s + L_r s + 1/C_r s)$ becomes 0 when the switching frequency is the same as f_m . Then, (9) can be simplified to

$$I_o(s) = \frac{n V_{ac}}{L_r s + \frac{1}{C_r s}} \quad (10)$$

From (10), the fixed output current operation is achieved at this operating point. The accuracy of these two characteristics are examined in Section III. In this paper, only the fixed output

voltage characteristic is adopted.

B. Fixed Output Current or LCL-T Operation Mode

When the mode selection switch S_{mode1} is OFF, S_{mode2} is ON, the converter is operating in fixed output current mode. The equivalent circuit is shown in Fig. 3(a), and the corresponding FHA model is shown in Fig. 3(b). Since f_s equals f_r , the resonant inductor L_r and resonant capacitor C_r in Fig. 3(b) can be removed, the voltage source is transferred from the primary side to the secondary side, and Fig. 3(b) can be simplified as shown in Fig. 3(c). Please note that v_{ac1}/n and v_{D1} are the fundamental harmonic of the transformer secondary side voltage and the rectifier output voltage, and R_{eq} is the equivalent resistance of the rectifier network.

$$v_{ac1} = V_{acm} \sin(\omega_s t) \quad (11)$$

$$R_{eq} = \frac{8}{\pi} R_L \quad (12)$$

where V_{acm} is the peak value of the fundamental harmonic voltage, which can be expressed as

$$V_{acm} = \frac{4}{\pi} \frac{V_{in}}{2n} \quad (13)$$

For the sake of analysis, the angular resonant frequency of the LCL-T resonant network ω_o , angular normalized switching frequency ω_n are defined as

$$\omega_o = \frac{1}{\sqrt{L_1 C_1}} \quad (14)$$

$$\omega_n = \frac{\omega_o}{\omega_s} \quad (15)$$

The ratio between the inductor L_1 and L_2 is defined as $\lambda = L_1/L_2$. The characteristic impedance Z_n and quality factor of the LCL-T resonant network Q are

$$Z_n = \sqrt{\frac{L_1}{C_1}} \quad (16)$$

$$Q = \frac{Z_n}{R_{eq}} \quad (17)$$

Based on Fig. 3(c), by only considering the fundamental harmonics of the circuit voltages and currents, the current I_{L2} can be expressed as

$$I_{L2} = \frac{V_{acm}}{Z_n} \frac{1}{\frac{1}{Q} (1 - \omega_n^2) + j[(1 + \lambda)\omega_n - \lambda\omega_n^3]} \quad (18)$$

From (18), it can be seen that when the angular normalized switching frequency ω_n equals 1, (18) can be simplified as

$$I_{L2}|_{\omega_n=1} = -j \frac{V_{acm}}{Z_n} \quad (19)$$

According to (19), the current I_{L_2} is independent of the load when $\omega_n = 1$, so the converter is operating in the fixed output current mode.

In time domain, the current $i_{L_2}(t)$ can be expressed as

$$i_{L_2}(t)|_{\omega_n=1} = \frac{V_{acm}}{Z_n} \sin\left(\omega_s t - \frac{\pi}{2}\right) \quad (20)$$

Therefore, the average value after the diode bridge rectifier can be obtained as

$$I_{o|\omega_n=1} = \frac{1}{\pi} \int_0^\pi i_{L_2}(t) d\omega_s t = \frac{2}{\pi} \frac{V_{acm}}{Z_n} \quad (21)$$

When $\omega_n = 1$, according to Fig. 4(c), the current I_{ac1} can be expressed as

$$I_{ac1} = \frac{V_{acm}}{Z_n} \left[\frac{1}{Q} - j(1 - \lambda) \right] \quad (22)$$

From (22), it can be seen that when the inductor ratio λ equals 1, the current I_{ac1} is in phase with the voltage v_{ac1} , which can be expressed as

$$I_{ac1}|_{\lambda=1} = \frac{V_{acm} R_{eq}}{Z_n^2} \quad (23)$$

Based on Fig. 3(b), the input impedance of the proposed converter in fixed output current mode can be derived as

$$\begin{aligned} Z_{in2}(s) &= Z_{LLC}(s) // Z_{LCL-T}(s) \\ &= L_m s // n^2 \left(\frac{(L_2 s + R_{eq}) \frac{1}{C_1 s}}{L_2 s + R_{eq} + \frac{1}{C_1 s}} + L_1 s \right) \end{aligned} \quad (24)$$

Since the input impedance is purely inductive, only the impedance of the LCL-T resonant network is investigated, which can be expressed as

$$Z_{LCL-T}(s) = n^2 \left(\frac{(L_2 s + R_{eq}) \frac{1}{C_1 s}}{L_2 s + R_{eq} + \frac{1}{C_1 s}} + L_1 s \right) \quad (25)$$

By substituting $s = \omega_s j$ into (25), the following equation can be obtained.

$$Z_{LCL-T} = n^2 \left[\frac{L_2}{R_{eq} C_1} + \left(L_1 \omega_s - \frac{1}{C_1 \omega_s} \right) j \right] \quad (26)$$

If the angular resonant frequency ω_o equals the angular switching frequency ω_s , the imaginary part of Z_{LCL-T} equals zero, so the impedance of the LCL-T resonant network is pure resistive. From (24), the total input impedance is inductive due to the term of $L_m s$. Therefore, the total input impedance when the converter operating in fixed output current mode is

Fig. 4. Control diagram of the proposed converter.

inductive. Similarly, it is beneficial for ZVS operation of the switches.

C. Operational Analysis of the Proposed Converter for the Battery Charger Application

Li-ion batteries mainly have two charge states, namely, CC charge state and CV charge state. Fig. 4(a) shows the system control diagram. It can be seen that during the CC charge state, the mode selection switch S_{mode1} is OFF, S_{mode2} is ON, and the converter is operating in the fixed output current or LCL-T operation mode. During the CV charge state, the mode selection switch S_{mode1} is ON, S_{mode2} is OFF, and the LCL-T resonant network is bypassed, and the converter is operating in the fixed output voltage or LLC operation mode. Please note that the error signal between the measured output voltage and the reference voltage is used as the control signal for the mode selection switch. When the error is greater than 0, the mode selection switch is at upper position, so the battery is in the CC charge state to increase the battery voltage. If the battery voltage reaches the reference voltage or the maximum charge voltage, the converter operation mode will be changed from CC to CV, and the mode selection switch is at lower position.

III. DESIGN GUIDELINE

A. Design Guidelines for the Proposed Converter

The battery pack charging profile is shown in Fig. 5. During the CC charging state, the converter's output voltage increases from 50 V to 90 V, while its output current is fixed at 2 A; during the CV charging state, the converter's output voltage is fixed at 90 V, and the output current drops from 2 A to 0.2 A. The system specifications are summarized in Table I. Please note that the nominal input voltage is 340 V.

1) Design Considerations for LLC Resonant Tank

The LLC resonant tank is adopted to achieve the fixed output voltage function. Therefore, f_r should be designed equal to f_s . Normal operation is selected when the input voltage equals 340 V. The following design and experimental verification are

Fig. 5. Battery charging profile.

TABLE I
SYSTEM SPECIFICATIONS

Parameter	Designator	Value
Input voltage	$V_{i, \min} - V_{i, \max}$	320 – 360 V
Output voltage	V_o	50 – 90 V
Output current	I_o	2 – 0.2 A
Output power	$P_{o, \min} - P_{o, \max}$	18 – 180 W
Switching frequency	f_s	100 kHz

based on the normal input voltage.

First, the unity gain operation occurs at the CV charge state, so n can be calculated as

$$n = \frac{V_{in}}{2V_o} = 1.9 \quad (27)$$

Then, the voltage stress limitation and the commercially available capacitance values are considered when designing the resonant capacitor C_r . (28) is the expression of the voltage stress on the resonant capacitor. A minimum resonant capacitor value of 31.5 nF is calculated by setting the voltage stress as 250 V. Ultimately, $C_r = 66$ nF was selected (a convenient value available).

$$V_{C_r, \max} = nV_o + \frac{I_{o, \max}}{4nf_r C_{r, \min}} \quad (28)$$

The resonant inductor L_r can be calculated based on (29).

$$f_s = f_r = \frac{1}{2\pi\sqrt{L_r C_r}} \quad (29)$$

According to (29), $L_r = 38.34$ μH is calculated.

Next, the magnetizing inductance L_m is determined based on: 1) the resonant network root-mean-square (RMS) current; 2) ZVS operation condition; 3) the operating switching frequency range. The resonant network RMS current $I_{L_r, \text{rms}}$ can

Fig. 6. Relationship between $I_{L_r, \text{rms}}$ and L_m .

Fig. 7. Relationship between the load resistance and the input impedance angle with different magnetizing inductance value.

be expressed as

$$I_{L_r, \text{rms}} = \sqrt{\frac{\pi^2 I_o^2}{8n^2} + \frac{n^2 V_o^2}{32L_m^2 f_r^2}} \quad (30)$$

Based on (30), the relationship between $I_{L_r, \text{rms}}$ and L_m can be drawn as shown in Fig. 6. Since the conduction loss is proportional to the resonant network RMS current, a smaller magnetizing inductance value will lead to a higher conduction loss. In addition, when the L_m reaches about 200 μH , the RMS current has begun to level out and changes more slowly with increasing inductance.

Another consideration for the magnetizing inductance design is the ZVS condition for the switches. From (8), the ZVS angle will affect the reactive power and the power efficiency of the converter. Specifically, a low reactive power or a low input impedance angle is desired to minimize the reactive power and improve the system efficiency. In this research, the switching frequency is selected based on the primary switch performance and passive elements size. The load resistance is determined based on the application requirement. Fig. 7 shows the relationship between the input impedance angle and different load resistance with different magnetizing inductance values. Clearly, a higher magnetizing inductance value can reduce the input impedance angle and the corresponding reactive power and conduction loss. However, a large magnetizing inductance value will also decrease the converter voltage gain and the soft

switching operation may not guaranteed due to the decrease of resonant inductor current during the dead time. Therefore, the design of magnetizing inductance should take these factors into consideration.

The resonant inductor current during dead time $i_{L_r_dead}$ equals the magnetizing inductor current, by ignoring the current variation during the dead time, it can be simply expressed as

$$i_{L_r_dead} = i_{L_m_pk} = \frac{nV_o D}{2L_m f_s} \quad (31)$$

where D is the duty cycle.

To completely extracted the charge stored on the junction capacitor of the switch, the dead time and $i_{L_r_dead}$ should satisfy (32).

$$Q = i_{L_r_dead} \cdot t_{dead} \geq 2C_{ds1} V_{in, max} \quad (32)$$

From (32), the boundary value for L_m can found.

$$L_m \leq \frac{nV_o D t_{dead}}{4C_{ds} V_{in, max} f_s} \quad (33)$$

where t_{dead} is the dead time of primary switches, and C_{ds} is the junction capacitance of the primary switch. In this experiment, $t_{dead} = 100$ ns and $C_{ds} = 50$ pF are selected. Then, the maximum value for the magnetizing inductance L_m equals to 330.8 μ H can be calculated.

Meanwhile, to avoid wide switching frequency operation, a small magnetizing inductance value is preferred to cover the entire input voltage range. Therefore, by taking both the resonant network RMS current and the switching frequency operating range into consideration, finally, $L_m = 287$ μ H is selected.

2) Design Considerations for LCL-T Resonant Tank

The LCL-T resonant tank is adopted to achieve the fixed output current characteristic. As previously stated, when the LCL-T resonant frequency f_o equals the switching frequency f_s , the average output current I_o is fixed. The LCL-T resonant network parameters are determined based on the battery charging current during the CC charge state I_{B_CC} .

$$f_s = f_o = \frac{1}{2\pi\sqrt{L_1 C_1}} \quad (34)$$

$$I_{B_CC} = I_o = \frac{2}{\pi} \frac{U_{acm}}{Z_n} \quad (35)$$

Based on (34) and (35), $L_1 = 57.75$ μ H, $C_1 = 43.86$ nF can be calculated, $L_2 = L_1 = 57.75$ μ H is selected. In practice, $L_1 = L_2 = 57$ μ H, $C_1 = 44$ nF are selected.

For the traditional PFM LLC, according to [8], the transformer turns ratio is determined based on the minimum input voltage and output voltage: $n_1 = 320/2/50 = 3.2$. In comparison to the proposed topology, the other resonant tank parameters remain unchanged.

TABLE II
CONVERTER PARAMETERS

Parameters	Values
Resonant capacitor C_r	66 nF
Resonant inductor L_r	38 μ H
Magnetizing inductor L_m	287 μ H
Turns ratio n	1.9 : 1
Resonant inductor L_1	57 μ H
Resonant capacitor C_1	44 nF
Resonant inductor L_2	57 μ H
Series resonant frequency f_r	100 kHz
Parallel resonant frequency f_m	34 kHz
Resonant frequency f_o for LCL-T	100 kHz
Switching frequency f_s	100 kHz
Input voltage	100 V

The proposed converter with LLC and LCL-T resonant tanks.

Fig. 8. Experimental waveform when the LLC converter is operating at the fixed output voltage mode. (a) Ch1: 1 A/div, Ch2: 1 A/div, Ch3: 50 V/div, Ch4: 10 V/div, time: 4 μ s/div; (b) (c) (d) Ch1: 200 mA/div, Ch2: 500 mA/div, Ch3: 50 V/div, Ch4: 10 V/div, time: 4 μ s/div.

B. Examination and Comparison of the Accuracy of Different Resonant Tanks Characteristics

Although the output voltage or output current are independent of the load for these three characteristics theoretically, variations still exist in practice. Therefore, before we adopt these characteristics, the error between the experimental results and theoretical results are examined and compared in this subsection. The experimental setup parameters are given in Table II.

According to (7), when $f_s = f_r$, the output voltage can be calculated as 27.7 V. The experimental waveform of the output voltage V_o with different loads is shown in Fig. 8. With the load changes from 10 Ω to 110 Ω , the output voltage variation is very small, from 26.2 V to 27.8 V. Please note that the load variation range is selected mainly based on the equivalent load shown in Fig. 5. The error is defined as the difference between theoretical value and practical value over practical value, and Fig. 9 shows

Fig. 9. Error of the fixed output voltage mode for LLC resonant tank.

Fig. 11. Error of the fixed output current mode for LLC resonant tank.

Fig. 10. Experimental waveform when the LLC converter is operating at the fixed output current mode. (a) (b) Ch1: 1 A/div, Ch2: 2 A/div, Ch3: 50 V/div, Ch4: 10 V/div, time: 10 μs/div; (c) (d) Ch1: 500 mA/div, Ch2: 2 A/div, Ch3: 50 V/div, Ch4: 25 V/div, time: 10 μs/div.

Fig. 12. Experimental waveform when the LLC-T converter is operating at the fixed output current mode. (a) Ch1: 200 mA/div, Ch2: 500 mA/div, Ch3: 50 V/div, Ch4: 5 V/div, time: 4 μs/div; (b) 200 mA/div, Ch2: 500 mA/div, Ch3: 50 V/div, Ch4: 10 V/div, time: 4 μs/div; (c) (d) Ch1: 500 mA/div, Ch2: 200 mA/div, Ch3: 50 V/div, Ch4: 25 V/div, time: 4 μs/div.

the error of the fixed output voltage operation mode. Clearly, the error is lower than 6% over the entire operating range.

According to (10), when $f_s = f_m$, the output current is calculated as 1.29 A. The waveform of output current I_o with different loads is shown in Fig. 10. With the load changes from 10 Ω to 110 Ω, the output current variation is large, from 1.49 A to 0.775 A. Fig. 11 shows the error of the fixed output current operation mode for LLC resonant tank. Clearly, the accuracy of this characteristic is unsatisfactory. This can be explained by observing the harmonics of the resonant current. Considerable errors would occur by only considering fundamental harmonics in this case.

For the LCL-T resonant tank, according to (21), when $f_s = f_o$, the output current is calculated as 0.585 A. Fig. 12 shows the experimental waveform of the output current I_o with different loads. It can be seen that with the load changes from 10 Ω to 110 Ω, the output current variation is not large, from 0.583 A to 0.474 A. Fig. 13 shows the error of the fixed output current operation mode. Clearly, the error is lower than 20% over the entire operating range.

Fig. 14 shows the error comparison results of different

Fig. 13. Error of the fixed output current mode for LCL-T resonant tank.

resonant tank characteristics, the fixed output voltage characteristic of the LLC resonant tank has the lowest error; while the error of fixed output current characteristic of the LLC resonant tank is large. For the battery charger application in this paper, the equivalent load resistance varies from 25 Ω to 45 Ω, the error of the LCL-T resonant tank is still very low in this

Fig. 14. Error comparison results.

Fig. 15. Experimental system setup.

range (lower than 5%). While for the LLC fixed output current operation mode, the error is large, from 13% to 28%. Therefore, in this paper, the fixed output voltage characteristic of the LLC resonant tank and the fixed output current characteristic of the LCL-T resonant tank are adopted to achieve the CV and CC charge state of the battery, respectively. Due to the high accuracy feature of these two characteristics, the required operating switching frequency range is significantly reduced when compared with the traditional LLC resonant converter.

IV. EXPERIMENT RESULTS

In this section, a scaled-down prototype was built and tested. The main purpose of the experiment is to validate the concept of using LLC and LCL-T resonant tanks to achieve CV operation and CC operation with a very narrow switching frequency operation range.

The picture of the laboratory prototype is shown in Fig. 15, and Table III summarizes the circuit parameters and part numbers. The load resistor is used to simulate the battery equivalent resistance. The efficiency is measured by using HIOKI 3193 power analyzer. To achieve the function of S_{mode1} and S_{mode2} , four MOSFETs are required to block the voltage and current, which increases the circuit complexity and cost. To reduce the circuit complexity and cost, two power relays are used

TABLE III
CIRCUIT PARAMETERS AND COMPONENT PART NUMBERS

Parameters	Part number	Value
Resonant capacitor C_r	TDK B32621A4333J000	66 nF
Resonant inductor L_r	TDK B66319G0000X187 EE 35×15×7 with N87 material	$N_{Lr} = 10$ 38 μ H
Magnetizing inductor L_m	TDK 445-175091-ND PQ32/30Z-12 with PC44 material	$N_{pri} = 19$ $N_{sec} = 10$ 287 μ H
Turns ratio n		1.9:1
Resonant inductor L_1	TDK B66319G0000X187 EE 35×15×7 with N87 material	$N_{L1} = 8$ 57 μ H
Resonant capacitor C_1	TDK B32520C6223K189	44 nF
Resonant inductor L_2	TDK B66319G0000X187 EE 35×15×7 with N87 material	$N_{L2} = 8$ 57 μ H
Diode rectifier	VT10200C SCHOTTKY RECTIFIER	200 V/5 A
Primary switch	CREE SiC C3M0120100K	1000 V/22 A
Mode selection switch	J120D1AS12VDC	300 VAC/20A
Relay driver	ULN2803A	8 channels
Gate driver IC	Silicon Lab Si 8271	2.5 kV RMS isolation

The proposed converter with LLC and LCL-T resonant tanks.

to achieve the function of mode selection switches due to the bi-directional voltage and current blocking capability of relays. The selection guidelines for the relays are based on the maximum operating voltage and maximum operating current, which can be easily obtained from simulation results. From simulation, it can be found that the maximum voltage and current for S_{mode1} is 180 V and 3.34 A, respectively. The maximum voltage and current for S_{mode2} is 97 V and 5.39 A. Finally, a power relay (J120D1AS12VDC) with maximum operation voltage and maximum operation current of 300 VAC and 20 A is selected and the unit price is around 2 \$ based on Digikey. The typical operate time and release time for the relay are 20 ms and 10 ms, respectively. Please note that fast operation time is not required since for the battery charger application, the charging process is continuous and slow. Therefore, low cost power relays can be implemented for the proposed converter. Moreover, the typical initial contact resistance of the relay is less than 50 m Ω , which has better performance than most of the power MOSFETs. Furthermore, the driving circuitry for power relays are more simple and cost-effective than power MOSFETs. In this article, the relay driver ULN2803A from Texas Instrument with unit price of around 1 \$ is used to drive two power relays. Due to the complementary operation of two power relays, a pair of PWM signal from digital signal processor (DSP) is used as the logic input to drive the power relays.

Fig. 16 shows the experimental waveform of the resonant

Fig. 16. Experimental waveform during the CC charge state. Ch1: 50 V/div, Ch2: 2 A/div, Ch3: 250 V/div, Ch4: 1 V/div (500 mV/A), time: 4 μ s/div.

Fig. 17. Experimental waveform of the soft switching operating for the proposed topology during the CC charge state. (a) and (b) Ch1: 2 A/div, Ch2: 2 A/div, Ch3: 100 V/div, Ch4: 5 V/div, time: 2 μ s/div.

tank input voltage V_{HB} , resonant tank current i_{Lr} , output current I_o , and output voltage V_o during the CC charge state with different load values. During the CC charge state, the output current is fixed at 2 A, the output voltage increases from 50 V to 90 V. It can be seen that the output current is remained constant at 2 A with a very narrow switching frequency operation range of [99 kHz, 102 kHz].

The details of the soft switching operation for the semiconductors of the proposed converter during the CC charge state is shown in Fig. 17, where V_{ds2} is the voltage across the primary switch S_2 , V_{gs2} is the gate drive signal for S_2 , i_{sec} is the transformer secondary side current, and i_{D1} is the current for rectifier diode D_1 . During the CC charge state, the ZVS operation is achieved since the voltage across the switch V_{ds2} decreases to zero before its driving signal arrives. In addition, the zero-current switching (ZCS) operation for rectifier diodes is also achieved.

Similarly, Fig. 18 shows the experimental waveforms during the CV charge state with different load values. During this state, the output voltage is fixed at 90 V, while its current decreases from 2 A to 0.26 A. A very narrow switching frequency operating range of [100 kHz, 107 kHz] is required to achieve output voltage regulation during CV charge state. Therefore, the CC charging and CV charging characteristics claimed by the proposed converter can be validated.

Fig. 19 shows the details of the soft switching operation for the semiconductors of the proposed converter during the

Fig. 18. Experimental waveform during the CV charge state. Ch1: 50 V/div, Ch2: 2 A/div, Ch3: 250 V/div, Ch4: 1 V/div (500 mV/A), time: 4 μ s/div.

Fig. 19. Experimental waveforms for the proposed topology during the CV charge state. (a) and (b) Ch1: 2 A/div, Ch2: 2 A/div, Ch3: 100 V/div, Ch4: 5 V/div, time: 2 μ s/div.

Fig. 20. Experiment waveform for traditional LLC converter in CC charging state with output voltage of 70 V. Ch1: 50 V/div, Ch2: 5 A/div, Ch3: 250 V/div, Ch4: 1 V/div (500 mV/A), time: 4 μ s/div.

CV charge state. Clearly, the ZVS and ZCS operation for the switches and rectifier diodes are guaranteed, respectively.

To make comparisons between the proposed converter and traditional LLC converter. The comparison is first made based on the same prototype with LLC operation mode. With the same parameters, the resonant frequency operating point is at the constant voltage charging value. Therefore, above resonant frequency operation is required during the CC charging state to reduce the converter voltage gain. Fig. 20 shows the experiment waveform when the LLC converter is operating in CC charging state with output voltage of 70 V. Clearly, to achieve output voltage range from 70 V to 90 V, the required switching frequency range is [100 kHz, 200 kHz]. For the application in this paper, the output voltage is in the range of

TABLE IV
COMPARISONS BETWEEN THE PROPOSED CONVERTER AND CONVENTIONAL LLC CONVERTER

Parameter	LLC converter	Proposed converter
Resonant capacitor C_r/nF	10.9	66
Resonant inductor $L_r/\mu\text{H}$	232	38
Magnetizing inductor $L_m/\mu\text{H}$	1500	287
Turns ratio n	3.4	1.9
Switching frequency range f_s/kHz	[44.8, 100]	100
Resonant capacitor maximum voltage $V_{Cr, \max}/\text{V}$	673	224
Switch turn-off current $I_{S, \text{off}}/\text{A}$	0.0866	1.5023
Primary switch maximum current $I_{S, \max}/\text{A}$	2.5187	2.2323
Secondary rectifier maximum current $I_{D, \max}/\text{A}$	7.08	3.34

[50 V, 90 V], a very wide switching frequency operating range is required.

Another comparison is made based on different LLC converter parameters. Normally, the LLC converter is designed in below resonant frequency operation region [8]. Thus, the transformer turns ratio is designed as

$$n = \frac{V_{\text{in}}}{2V_{\text{o, min}}} = 3.4 \quad (36)$$

By keeping the quality factor and inductor ratio same for the traditional LLC converter and the proposed converter, the following resonant circuit parameters can be calculated: $C_r = 10.9 \text{ nF}$, $L_r = 232 \mu\text{H}$, $L_m = 1.5 \text{ mH}$. Please note that the inductor ratio is reduced in order to satisfy the maximum voltage gain requirement. The comparison results are summarized in Table IV. The following benefits are achieved when compared with traditional LLC converter: 1) The switching frequency range is reduced significantly, almost fixed frequency operation can be achieved for the proposed converter, which is beneficial for the magnetic components and EMI filter design. 2) The voltage stress on the resonant capacitor is reduced significantly. The voltage stress for the resonant capacitor of traditional LLC converter is more than three times larger than the proposed converter. The reliability of the system can be improved with reduced voltage stress. 3) For traditional LLC converter, it is operating around the peak gain point, which is also the boundary of ZVS and ZCS operation. Therefore, the ZVS operation for primary switch is tricky to be satisfied for traditional LLC converter. From the switch turn-off current, the switch turn-off current for traditional LLC converter is only 0.0866 A, while the turn-off current for the proposed converter is 1.5023 A. Thus, either an extreme large deadtime is required to achieve ZVS operation or the ZVS operation may lost for the traditional LLC converter. 4) The current stress for the traditional LLC converter is larger than the proposed. Both the peak currents for primary switch and secondary rectifier are larger than the proposed converter.

The dynamic test for the proposed converter is presented in Fig. 21. Fig. 21(a) shows the dynamic test waveform when load resistor changes from 100 Ω to 50 Ω during CV charge state. It can be seen that during the CV charge state, the output

Fig. 21. Dynamic waveform of the proposed converter. (a) dynamic waveform during CV charge state; (b) dynamic waveform during CC charge state. Ch1:50 V/div, Ch2: 5 A/div, Ch3: 250 V/div, Ch4: 1 V/div (500 mV/A), time: 200 ms/div.

Fig. 22. Mode transitioning from CC charging state to CV charging state of the proposed converter. Ch1: 50 V/div, Ch2: 5 A/div, Ch3: 250 V/div, Ch4: 1 V/div (500 mV/A), time: 200 ms/div.

voltage remains constant at 90 V during the dynamic transition. In addition, there is no voltage overshoot or undershoot during the transition since during the CV charge state, the LLC converter is operating at resonant frequency, its output voltage is constant and independent of load. Therefore, when the load varies, the output voltage remains constant at 90 V. Fig. 21(b) shows the dynamic test waveform when load resistor changes from 40 Ω to 30 Ω in CC charge state. Similarly, the output current remains constant at 2 A and no current overshoot or undershoot occurs during the transition. In CC charge state, due to the existence of LCL-T resonant tank, the output current is constant and independent of load. Thus, when the load changes, the output current remains constant at 2 A. Another main advantage for the proposed LLC and LCL-T resonant tanks based battery charger topology is that it has good dynamic performance, no voltage/current overshoot/undershoot exist and the response time is very fast.

The equivalent load resistance at mode transitioning point is 45 Ω . Thus, in the experiment, a step change on the load resistor is generated to emulate the mode transitioning from CC charging state to CV charging state. Fig. 22 shows the dynamic experiment waveform when the load changes from 40 Ω to 45 Ω . When the load resistance equals 40 Ω , the converter is operating in LCL-T operation mode to provide CC of 2 A and the output voltage is 80 V. When the load resistance is changed to 45 Ω , the converter is transitioning from CC charging state to CV charging state so that the output voltage is regulated at 90 V. From the experiment waveform, the dynamic response of the proposed converter is good and there are no large transients. The main reason is that for the proposed converter, its operating switching frequency is always around the resonant

Fig. 23. Efficiency curve of the proposed converter.

frequency (100 kHz). Therefore, when the mode transitioning occurs, there is no large mismatch in the control variable. In real application, the dynamic performance is better since the charging process is continuous and there would not be a step change.

Fig. 23 shows the efficiency performance comparison between the proposed converter during the CC charge state, CV charge state and the traditional LLC converter. It can be seen that the maximum efficiency for CV charge state is around 96%, while the maximum efficiency for CC charge state is around 95%. In addition, at the same power level, the efficiency of CV charge state is higher than that of the CC charge state. This is because that additional LCL-T resonant tank is required during the CC charge state, which will result in some additional power loss. This is a scale down prototype, the converter efficiency can be improved with higher power rating.

A. Comparison of Power Loss Distribution

In this subsection, the power loss distribution of the proposed converter is presented, and the comparison between the CC charge state, CV charge state and the traditional LLC converter at the rated power is summarized in Table V, the power loss model introduced in [4] is adopted to calculate power loss for each component. During the CV charge state, the LCL-T resonant tank is bypassed and only the conduction loss of the additional relay is introduced. The conduction loss is about 0.1597 W, which has very small influence on the system efficiency. During the CC charge state, the power loss generated from the additional LCL-T resonant tank is about 2.095 W, which results in 1.2% efficiency degradation. However, compared with traditional LLC converters, the overall efficiency for the proposed topology is still higher. The reason is that at this operating point, the switching frequency of traditional LLC is much lower than the series resonant frequency, which increases the circulating current and conduction losses of the components.

B. Comparison to Existing Literature

Comprehensive comparisons between different topologies and the proposed one are given in Table VI. The reduced switching frequency operating range will bring many benefits for the resonant converter: 1) Converter electro-magnetic

TABLE V
POWER LOSS DISTRIBUTION COMPARISON BETWEEN THE PROPOSED CONVERTER AND THE TRADITIONAL LLC CONVERTER

Power loss categorizations	Proposed		Traditional PFM LLC
	CC charge state	CV charge state	
Primary switch loss/W	0.800	0.809	1.328
Secondary rectifier loss/W	3.7320	3.7912	3.868
Magnetic core loss/W	0.9322	0.9491	2.3814
Winding loss/W	0.669	0.931	2.559
Additional resonant inductors loss/W	0.93	0	0
Additional resonant capacitor loss/W	0.807	0	0
Additional mode section switch loss/W	0.119	0.053	0
Other loss/W	0.806	0.817	0.925
Total loss/W	8.895	7.1703	11.07
Estimated efficiency/%	95.29	96.17	94.2
Measured efficiency/%	94.8	96.01	-

interference (EMI) performance is better than traditional resonant topologies. 2) Design and optimization for magnetic components, including EMI filter, resonant inductor, and transformer, becomes easy since the operating frequency for these magnetic components is almost fixed. 3) Size reduction for magnetic components can be achieved. The minimum operating switching frequency should be selected to design the magnetic components, while for the proposed topology, the magnetic components can be designed at normal operating switching frequency. By increasing the switching frequency, the magnetic components size can be reduced. 4) Simple analyze and design for resonant converter. It is well known that for resonant converters, the analysis accuracy of FHA is poor when the operating switching frequency is away from the resonant frequency. Therefore, considerable errors would be generated if the FHA analysis is adopted to analyze and design the resonant converters. However, for the proposed topology, the operating switching frequency is around the resonant frequency, traditional simple FHA method can provide enough accuracy for converter design and the system efficiency can be improved when compared with traditional LLC converter with FHA analysis method. 5) Control stability and system reliability are improved. Since the proposed topology is always operating around the resonant frequency point, the relationship between the control variable (switching frequency) and output is always monotonous, and there is no control stability issue.

V. CONCLUSION

In this article, a novel topology based on the LLC and LCL-T resonant tanks for battery charger applications is proposed. Compared with the traditional LLC resonant converter, the required operating switching frequency range is reduced significantly due to the high accuracy of the fixed output voltage characteristic of the LLC resonant tank and the fixed output current characteristic of the LCL-T resonant tank.

TABLE VI
COMPARISONS BETWEEN DIFFERENT TOPOLOGIES

Topology	Conventional LLC [10]	Hybrid control strategy [30]	Modified structure [23]	Dual-bridge LLC [29]	Dual full-bridge LLC [24]	Dual half-bridge with magnetic control [16]	PWM type LLC [31]	Hybrid LLC with multiple operation modes [32]	LLC with adjustable transformer turns ratio [33]	Proposed
Additional switch	None	None	2	2	1	None	1	None	4 MOSFETs and 2 diodes	2
Losses of additional switch	None	None	High (both conduction and switching loss)	High (both conduction and switching loss)	Low (only conduction loss)	None	Yes (both conduction loss and switching loss)	None	Yes (both conduction loss and switching loss)	Low (only conduction loss)
Number of rectifier network	1	1	2	1	2	2	1	1	1	1
Number of transformer	1	1	2	1	2	2	1	2	1	1
Number of resonant components	3	3	4	3	6	6	3	5	3	6
Modulation strategy	PFM	PFM+PSM	PFM	PWM	PFM	MC	PWM	PFM	PFM	PFM
Switching frequency range	Wide range	Medium range	Medium range	Fixed	Medium range	Fixed	Fixed	Medium	Medium	Narrow range
Magnetic components and gate driver design	Complicate	Medium	Medium	Simple	Medium	Simple	Simple	Medium	Medium	Simple
Analysis and design method	FHA (Simple)	TDA (Complex)	TDA (Complex)	TDA (Complex)	FHA (Simple)	FHA (Simple)	FHA (Simple)	FHA (Simple)	FHA (Simple)	FHA (Simple)
Accuracy of analysis	Poor	High	High	High	Poor	Poor	Poor	Poor	Poor	High
Control complexity	Simple	Complex	Complex	Complex	Medium	Simple	Simple	Complex	Complex	Simple
Control stability and reliability	Low	Medium	Low	High	Low	Low	High	Medium	Medium	High

In addition, a larger magnetizing inductance value compared with the traditional LLC can be selected to reduce circulating current, resonant tank components conduction losses, and turn-off loss of the primary switches. The operational principles and design considerations for the proposed converter are presented, and the accuracy of the resonant tanks' characteristics are investigated. Moreover, the experimental results are provided to validate the theoretical analysis and demonstrate the advantages of the proposed converter in terms of operating switching frequency range. The dynamic performance for the proposed converter is good with fast response time.

REFERENCE

- [1] M. Yilmaz and P. T. Krein, "Review of battery charger topologies, charging power levels, and infrastructure for plug-in electric and hybrid vehicles," in *IEEE Transactions on Power Electronics*, vol. 28, no. 5, pp. 2151–2169, May 2013.
- [2] J. Deng, C. C. Mi, R. Ma, and S. Li, "Design of LLC resonant converters based on operating-mode analysis for level two PHEV battery chargers," in *IEEE/ASME Transactions on Mechatronics*, vol. 20, no. 4, pp. 1595–1606, Aug. 2015.
- [3] Bo Yang, F. C. Lee, A. J. Zhang, and Guisong Huang, "LLC resonant converter for front end DC/DC conversion," in *Proceedings of APEC Seventeenth Annual IEEE Applied Power Electronics Conference and Exposition (Cat. No. 02CH37335)*, Dallas, TX, USA, 2002, pp. 1108–1112.
- [4] C. Yang, T. Liang, K. Chen, J. Li, and J. Lee, "Loss analysis of half-bridge LLC resonant converter," in *Proceedings of 2013 1st International Future Energy Electronics Conference (IFEEEC)*, Tainan, 2013, pp. 155–160.
- [5] Y. Wei, Q. Luo, S. Chen, P. Sun, and N. Altin, "Comparison among different analysis methodologies for LLC resonant converter," in *IET Power Electronics*, vol. 12, no. 9, pp. 2236–2244, Aug. 2019.
- [6] K. Kim, H. Youn, J. Baek, Y. Jeong, and G. Moon, "Analysis on synchronous rectifier control to improve regulation capability of high-frequency LLC resonant converter," in *IEEE Transactions on Power Electronics*, vol. 33, no. 8, pp. 7252–7259, Aug. 2018.
- [7] R. Beiranvand, B. Rashidian, M. R. Zolghadri, and S. M. H. Alavi, "Using LLC resonant converter for designing wide-range voltage source," in *IEEE Transactions on Industrial Electronics*, vol. 58, no. 5, pp. 1746–1756, May 2011.
- [8] J. Deng, S. Li, C. C. Mi, and R. Ma, "Design methodology of LLC resonant converter for electric vehicle battery chargers," in *IEEE Transactions on Vehicular Technology*, vol. 63, no. 4, pp. 1581–1592, May 2014.
- [9] F. Musavi, M. Craciun, D. S. Gautam, and W. Eberle, "Control strategies for wide output voltage range LLC resonant DC–DC converters in battery chargers," in *IEEE Transactions on Vehicular Technology*, vol. 63, no. 3, pp. 1117–1125, Mar. 2014.
- [10] F. Musavi, M. Craciun, D. S. Gautam, W. Eberle, and W. G. Dunford, "An LLC resonant DC–DC converter for wide output voltage range battery charging applications," in *IEEE Transactions on Power Electronics*, vol. 28, no. 12, pp. 5437–5445, Dec. 2013.
- [11] Y. Wei, Q. Luo, X. Lv, P. Sun, and X. Du, "Analysis and design of function decoupling high voltage gain DC/DC converter," in *Journal of Power Electronics*, vol. 19, no. 2, pp. 380–393, Mar. 2019.
- [12] Z. Li, T. Wu, G. Zhang, and R. Yang, "Hybrid modulation method combining variable frequency and double phase-shift for a 10 kW LLC resonant converter," in *IET Power Electronics*, vol. 11, no. 13, pp. 2161–2169, 2018.

- [13] X. Yudi, M. Xingkui, Z. Zhe, and Y. Shi, "New hybrid control for wide input full-bridge LLC resonant DC/DC converter," in *Proceedings of 2018 3rd International Conference on Intelligent Green Building and Smart Grid (IGBSG)*, Yi-Lan, 2018, pp. 1–4.
- [14] J. H. Kim, C. E. Kim, J. K. Kim, J. B. Lee, and G. W. Moon, "Analysis on load-adaptive phase-shift control for high efficiency full-bridge resonant converter under light-load conditions," in *IEEE Transactions on Power Electronics*, vol. 31, no. 7, pp. 4942–4955, Jul. 2016.
- [15] B. McDonald and F. Wang, "LLC performance enhancements with frequency and phase shift modulation control," in *Proceedings of 2014 IEEE Applied Power Electronics Conference and Exposition - APEC 2014*, pp. 2036–2040, Mar. 2014.
- [16] Y. Wei, Q. Luo, X. Du, N. Altin, A. Nasiri, and J. M. Alonso, "A dual half-bridge LLC resonant converter with magnetic control for battery charger application," in *IEEE Transactions on Power Electronics*, vol. 35, no. 2, pp. 2196–2207, Feb. 2020.
- [17] Y. Wei, Q. Luo, X. Du, N. Altin, J. M. Alonso, and H. A. Mantooth, "Analysis and design of the LLC resonant converter with variable inductor control based on time-domain analysis," in *IEEE Transactions on Industrial Electronics*, vol. 67, no. 7, pp. 5432–5443, Jul. 2020.
- [18] Y. Wei, N. Altin, Q. Luo, and A. Nasiri, "A high efficiency, decoupled on-board battery charger with magnetic control," in *Proceedings of 2018 7th International Conference on Renewable Energy Research and Applications (ICRERA)*, Paris, 2018, pp. 920–925.
- [19] V. S. Costa, M. S. Perdigão, A. S. Mendes, and J. M. Alonso, "Evaluation of a variable-inductor-controlled LLC resonant converter for battery charging applications," in *Proceedings of IECON 2016-42nd Annual Conference of the IEEE Industrial Electronics Society*, Florence, 2016, pp. 5633–5638.
- [20] Z. Hu, Y. Qiu, L. Wang, and Y. Liu, "An interleaved LLC resonant converter operating at constant switching frequency," in *Proceedings of 2012 IEEE Energy Conversion Congress and Exposition (ECCE)*, Raleigh, NC, 2012, pp. 3541–3548.
- [21] S. K. Chung, B. G. Kang, and M. S. Kim, "Constant frequency control of LLC resonant converter using switched capacitor," in *Electronics Letters*, vol. 49, no. 24, pp. 1556–1558, 2013.
- [22] Z. Liang, R. Guo, G. Wang, and A. Huang, "A new wide input range high efficiency photovoltaic inverter," in *Proceedings of 2010 IEEE Energy Conversion Congress and Exposition*, Atlanta, GA, 2010, pp. 2937–2943.
- [23] H. Hu, X. Fang, F. Chen, Z. J. Shen, and I. Batarseh, "A modified high-efficiency LLC converter with two transformers for wide input-voltage range applications," in *IEEE Transactions on Power Electronics*, vol. 28, no. 4, pp. 1946–1960, Apr. 2013.
- [24] H. Vu and W. Choi, "A novel dual full-bridge LLC resonant converter for CC and CV charges of batteries for electric vehicles," in *IEEE Transactions on Industrial Electronics*, vol. 65, no. 3, pp. 2212–2225, Mar. 2018.
- [25] Q. Luo, S. Zhi, C. Zou, W. Lu, and L. Zhou, "An LED driver with dynamic high-frequency sinusoidal bus voltage regulation for multistring applications," in *IEEE Transactions on Power Electronics*, vol. 29, no. 1, pp. 491–500, Jan. 2014.
- [26] M. Borage, S. Tiwari, and S. Kotaiah, "Analysis and design of an LCL-T resonant converter as a constant-current power supply," in *IEEE Transactions on Industrial Electronic*, vol. 52, no. 6, pp. 1547–1554, Dec. 2005.
- [27] M. Borage, S. Tiwari, and S. Kotaiah, "LCL-T resonant converter with clamp Diodes: A novel constant-current power supply with inherent constant-voltage limit," in *IEEE Transactions on Industrial Electronic*, vol. 54, no. 2, pp. 741–746, Apr. 2007.
- [28] H. Wang, S. Dusmez, and A. Khaligh, "Design and analysis of a full-bridge LLC-based PEV charger optimized for wide battery voltage range," in *IEEE Transactions on Vehicular Technology*, vol. 63, no. 4, pp. 1603–1613, May 2014.
- [29] X. Sun, X. Li, Y. Shen, B. Wang, and X. Guo, "Dual-bridge LLC resonant converter with fixed-frequency PWM control for wide input applications," in *IEEE Transactions on Power Electronics*, vol. 32, no. 1, pp. 69–80, Jan. 2017.
- [30] S. Liu and F. Zhang, "State variable derivation with numerical approach and efficiency optimisation method for phase-shift LLC converters under wide voltage-gain range," in *IET Power Electronics*, vol. 12, no. 7, pp. 1752–1762, Jul. 2019.
- [31] H. Wang and Z. Li, "A PWM LLC type resonant converter adapted to wide output range in PEV charging applications," in *IEEE Transactions on Power Electronics*, vol. 33, no. 5, pp. 3791–3801, May 2018.
- [32] L. A. D. Ta, N. D. Dao, and D. Lee, "High-efficiency hybrid LLC resonant converter for on-board chargers of plug-in electric vehicles," in *IEEE Transactions on Power Electronics*, vol. 35, no. 8, pp. 8324–8334, Aug. 2020.
- [33] D. Shu and H. Wang, "An ultrawide output range LLC resonant converter based on adjustable turns ratio transformer and reconfigurable bridge," in *IEEE Transactions on Industrial Electronics*, vol. 68, no. 8, pp. 7115–7124, Aug. 2021.

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