Analysis and Design of Three-Level Full-Bridge LLC Resonant Converter Based on Pulse Frequency Modulation

J Z Liu¹,², Y Zhao¹, P Hu¹ and T T Chen¹

¹Hubei Collaborative Innovation Center for High-efficiency Utilization of Solar Energy, Hubei University of Technology, Wuhan 430068, China

E-mail: liujunzhe88@qq.com

Abstract. Aiming at the problem of serious switching loss in high-power and high-voltage occasions, the traditional LLC resonant network is difficult to adapt, the DC gain is difficult to obtain intuitively, and the voltage stress on the switch is too high, resulting in large switching losses and difficult selection of the switch. This paper uses a three-level full-bridge LLC resonant converter topology, the First Harmonic Approximation (FHA) is used to analyze the LLC resonant converter gain characteristics, and the parameter design ideas are given. A simulation experiment of a three-level full-bridge LLC resonant converter based on frequency modulation control is presented. The simulation results show that each primary side switch transistor can achieve zero voltage switching (ZVS), the secondary side diode realizes zero current switching (ZCS), and can obtain higher working efficiency, which verifies the feasibility of the design.

1. Introduction

At a high switching frequency, LLC resonant converter can still achieve zero voltage turn-on (ZVS) on the primary side and zero current turn-off (ZVS) on the secondary diode, which greatly reduces switching losses, and has a simple circuit structure and power density. The advantages of high efficiency, high efficiency, and ease of integration have received widespread attention. However, the poor voltage resistance of the MOSFET switch is limited to high-power applications. In the literature, the converter combines the advantages of LLC and three-level technology. By reasonably allocating the dead time of the inner and outer transistors, the voltage stress of each switch can be achieved. Both are half of the input voltage, thus adapting to high-power high-voltage occasions [1]. Literature [2-5] introduced some design schemes of LLC resonant converter.

This paper overcomes the difficulty that the DC gain of the three-level full-bridge LLC resonant converter cannot be obtained intuitively [6], and solves the difficulty of high design difficulty due to the high correlation of the resonant tank parameters [7]. It is necessary to design separately to avoid the dead zone of the inner and outer transistor the unreasonable time design causes the switch transistor to be unable to complete the zero voltage switch (ZVS) [8], and ensures the working efficiency of the circuit.

In this paper, the working process of three-level full-bridge LLC resonant converter in frequency modulation mode is analyzed in detail, and the equivalent working model in each mode is established. The fundamental wave analysis method is used to model the resonant converter [6-11], find the transfer function, and establish the gain curve in Mathcad to compare and analyze the effect of different power factor $Q$ and inductance ratio $k$ on the DC gain [10], through the analysis results to
complete the three-level full-bridge LLC resonant converter parameter design. Finally, through simulink simulation, it is verified that the design method in this paper can realize the soft switching characteristics of each switch, and the voltage stress of each switch is half of the input voltage.

2. Operational principle

Figure 1 shows the main circuit of a three-level full-bridge LLC resonant converter. Assuming the input capacitors \( C_{in1} \) and \( C_{in2} \), the output filter capacitor \( C_0 \) is large enough to be regarded as a constant output voltage source; the flying capacitors \( C_{ss1}, C_{ss2} \) are sufficiently large relative to the parasitic capacitance of the switch; Neutral grounding of the clamping diodes \( D_1, D_2, D_3, D_4 \) and the input capacitor; the outer transistors \( Q_1, Q_4, Q_5, Q_8 \) should be turned off earlier than the inner transistors \( Q_2, Q_3, Q_6, Q_7 \), to ensure that the voltage stress of each diode is half of the input voltage; \( C_1\sim C_8 \) are respectively Parasitic capacitance, and \( C_1=C_2=C_3=C_4=C_5=C_6=C_7=C_8; \) the resonant tank is composed of resonant capacitor \( Cr \), resonant inductance \( Lr \) and excitation inductance \( Lm \); secondary side rectifier diodes \( D_5 \sim D_8 \).

![Three-level full-bridge LLC resonant circuit main circuit diagram](image)

Figure 1. Three-level full-bridge LLC resonant circuit main circuit diagram

Figure 2 is the key waveform diagram of the three-level full-bridge LLC resonant converter operating in the frequency conversion mode, where \( V_{gs} \) is the gate voltage of the switch, \( i_{Lr} \) is the resonant current, \( i_{Lm} \) is the excitation current, and \( V_{Cr} \) is the voltage of the resonant capacitor \( Cr \); Input current for the secondary side, \( V_{rc} \) is the input voltage of the resonant tank, and \( i_d \) is the current flowing through the secondary side diode.
Figure 2. Key waveforms in the three-level full-bridge LLC frequency conversion mode

In order to simplify the analysis, the negative half cycle is similar to the positive half cycle, only the positive half cycle working mode of the switch is analyzed, as shown in Figure 3:
Mode 2(t1,t2)

Mode 3(t2,t3)

Mode 4(t3, t4)
Mode 1 (t0–t1): The switch transistor \( Q_1Q_2Q_7Q_8 \) is turned on, and the input voltage acts on the resonant tank through the switch transistor. At this time, the two elements of the resonance cavity \( C_r \) and \( L_r \) resonate, and the resonance frequency is \( f_0 = \frac{1}{2\pi}\sqrt{\frac{L_r}{C_r}} \). The current \( i_{L_r} \) increases in a sine form, the secondary diode \( D_5D_8 \) is turned on, and the current size is \( n(i_{L_r} - i_{L_m}) \), which charges the output capacitor \( C_0 \) and supplies power to the load. The primary voltage is clamped to \( nV_o \) by the secondary, and the excitation current \( i_{L_m} \) increases linearly.

Mode 2 (t1–t2) resonance current \( i_{L_r} \) is equal to the excitation current \( i_{L_m} \), the excitation inductance is out of voltage clamping. Resonant tank \( C_r, L_r, L_m \) three elements resonate, the resonance frequency at this time \( f_1 = \frac{1}{2\pi}\sqrt{\frac{L_r + L_m}{C_r}} \). The secondary diode \( D_5D_8 \) has no current through natural shutdown, to achieve zero current switching (ZCS), to prevent reverse recovery of the diode, the load is powered by the output capacitor \( C_0 \).

Mode 3 (t2–t3) switch \( Q_1Q_8 \) is turned off earlier than \( Q_2Q_7 \). At this time, the direction of the resonant circuit is not changed. The junction capacitor \( C_1C_8 \) of switch \( Q_1Q_8 \) is charged, and the junction capacitor \( C_4C_5 \) of switch \( Q_2Q_7 \) is discharged. Because the flying capacitors \( C_{ss1}, C_{ss2} \) can...
not be abruptly changed at both ends, the voltage across the switch \( Q_2Q_6 \) is always half of the input voltage \( U_{in} \).

Mode 4 (t3–t4) at t3, the resonant tank voltage is 0. When the junction capacitor \( C_1C_8 \) voltage reaches half of the input voltage \( U_{in} \), the clamping diode \( D_2D_3 \) conducts to prevent voltage overcharge. At this time, the voltage across the \( Q_3Q_6 \) complementary switch transistor \( Q_4Q_5 \) is also clamped to 0, the body diode is turned on, in preparation for the switch transistor \( Q_4Q_5 \) to achieve zero voltage switching (ZVS).

Mode 5 (t4–t5) at time t4, the switching transistor \( Q_2Q_7 \) is turned off, and the resonant circuit is freewheeling through \( D_2D_3 \), charging the junction capacitor \( C_2C_7 \) of the switching transistor \( Q_2Q_7 \), and discharging the junction capacitor \( C_3C_6 \) of the switching transistor \( Q_3Q_6 \) at the same time. The resonance current is less than the excitation current, the primary current is reversed, the secondary diode \( D_6D_7 \) is turned on, and the output capacitor \( C_0 \) starts to charge, and the load is powered. The primary voltage clamped by the secondary diode is \( nV_0 \), the excitation inductance of the resonant tank becomes out of resonance and becomes two-element resonance, the excitation current changes linearly, and the resonant current sinusoidal change.

Mode 6 (t5) The voltage across the junction capacitor \( C_2C_7 \) rises to half of the input voltage and is clamped. At the same time, the voltage across the junction capacitor \( C_3C_6 \) is clamped to 0. (ZVS) Prepare switch \( Q_3Q_6 \) for the next zero-voltage turn-on.

3. Performance analysis

Since the three-level full-bridge switch transistor converts the input constant current voltage into a square wave input resonant tank, the secondary side can be converted to the primary side, which can simplify the analysis of the three-level full-bridge LLC converter, as shown in Figure 4.

![Figure 4. Simplified equivalent diagram](image)

In order to improve efficiency, the LLC resonant converter is best to work near the resonant frequency. At this time, the cavity current is approximately sinusoidal, and the input voltage is a square wave with positive and negative amplitude. The First Harmonic Approximation (FHA) can be used to analyze the frequency conversion mode the relationship between gain and frequency of LLC resonant converter.

Assuming that the input voltage is a 50% duty cycle, the amplitude is \( V_{in} \) square wave, the transformer is an ideal transformer with a primary side and secondary side turns ratio of \( n \): 1, all the switching rectifier diodes are ideal diodes, the output capacitance is large enough, and the output The voltage is constant and the load is purely resistive.

Then the input voltage can be expressed as:

\[
    u(t) = \begin{cases} 
    V_{in}, & (0 < t < \frac{T}{2}) \\
    -V_{in}, & (\frac{T}{2} < t < T) 
    \end{cases}
\]  

(1)

Fourier decomposition and expansion of the square wave of formula (1) can be obtained as follow:
\[ u_i(t) = \frac{4}{\pi} V_n \sum_{n=1,3,5,...}^{\infty} \frac{1}{n} \sin(n \omega_0 t) \]  

(2)

Bring \( n=1 \) into equation (2) to get the fundamental component:

\[ u_{i,FHA}(t) = \frac{4}{\pi} V_1 \sin(\omega_1 t) \]  

(3)

Due to the clamping effect of the diode on the secondary side, the input voltage on the secondary side is a square wave with an amplitude of \( V_o \). The input current on the secondary side is approximately a sine wave as shown in fig2. Expand its Fourier

\[ u_o(t) = \frac{4}{\pi} V_o \sum_{n=1,3,5,...}^{\infty} \frac{1}{n} \sin(n \omega_o t - \theta) \]  

(4)

Bring \( n=1 \) into equation (4) to get the fundamental component:

\[ u_{o,FHA}(t) = \frac{4}{\pi} V_o \sin(\omega_o t - \theta) \]  

(5)

The secondary load is purely resistive, so the secondary current and voltage are in phase as follow:

\[ i_{o,FHA}(t) = \sqrt{2} I_o \sin(\omega_o t - \theta) \]  

(6)

The equivalent resistance of the secondary side can be obtained as follow:

\[ R_o = \frac{u_{o,FHA}(t)}{i_{o,FHA}(t)} = \frac{4V_o}{\sqrt{2} \pi I_o} \]  

(7)

It can be seen from the figure that the current flowing through the load \( R \) as follow:

\[ I = \frac{2}{\pi} \left[ T \sqrt{2} I_o \sin(\omega t - \theta) \right] dt = \frac{2\sqrt{2}}{\pi} I_o \]  

(8)

Available in (7) as follow:

\[ R_o = \frac{8V_o}{\pi^2 I} = \frac{8}{\pi^2} R \]  

(9)

It can be obtained by converting the impedance of the secondary side to the primary side as follow:

\[ R_{ac} = n^2 R_o \]  

(10)

Finally available

\[ R_{ac} = \frac{8n^2}{\pi^2} R \]  

(11)

Transfer Function:

\[ G_S = \frac{V_o(s)}{V_{in}(s)} = \frac{sR_{ac}L_m}{sL_r + \frac{1}{sC_r} + \frac{sR_{ac}L_m}{R_{ac} + sL_m}} = \frac{s^2 R_{ac} C_r L_m}{s^3 C_r L_m + s^2 C_r R_{ac} (L_r + L_m) + sL_m + R_{ac}} \]  

(12)

Define characteristic impedance \( Z_r = \sqrt{L_r / C_r} \); Inductance ratio \( k = L_m / L_r \); Quality factor \( Q = Z_r / R_{ac} \); Frequency normalization \( f_0 = f_s / f_0 \), \( f_s \) is the operating frequency; Resonant
frequency \( f_o = 1 / 2\pi \sqrt{L/C} \); The amplitude-frequency characteristic of the transfer function can be obtained by taking the modulus value of (12):

\[
|G(s)| = \frac{kf_n^2}{\sqrt{\left(1 - f_n^2(1+k)\right)^2 + k^2Q^2f_n^2(1-f_n^2)^2}}
\]  

(13)

Phase frequency characteristics as follow:

\[
\angle G(s) = 180^\circ - \arctg \left[\frac{Qkf_n(1-f_n)}{1-f_n^2(1+k)}\right]
\]  

(14)

The net gain of the resonant converter can be sorted out:

\[
G(f_n,Q,k) = \frac{1}{\sqrt{1 + \frac{1}{k}\left(\frac{1}{f_n^2} - 1\right)^2 + Q^2\left(\frac{1}{f_n} - f_n\right)^2}}
\]  

(15)

According to equation (15), it can be seen that the gain of the resonant network is related to the values of the operating frequency \( f_n \), the circuit quality factor \( Q \) and the inductance coefficient \( k \), and the corresponding gain curve can be drawn through Mathcad[12].

By fixing the \( Q \) value of the circuit quality factor and changing the inductance coefficient \( k \) value, it can be seen from Figure 5 that as the inductance coefficient \( k \) value increases, the gain range decreases, especially when \( k=50 \), the gain never exceeds 1. And as the value of \( k \) increases, the operating frequency required to change the gain becomes wider, which will make the dynamic adjustment capability of the resonant converter worse, but the stability of the system can be improved. Although a too small inductance can obtain higher gain sensitivity, it will result in poor system stability, and a too small excitation inductance \( Lm \) will cause a larger current to flow in the resonant tank, thereby reducing the operation of the resonant converter effectiveness. Therefore, the choice of inductance \( k \) should be considered in compromise.

![Figure 5. DC gain at different inductance ratios](image)

By fixing the inductance \( k \) value and changing the \( Q \) value of the circuit quality factor, it can be
seen from Figure 6 that increasing the $Q$ value reduces the gain range and reduces the operating frequency range. If the secondary side equivalent impedance $R_{ac}$ is fixed, the circuit quality factor $Q$ is inversely proportional to the resonance capacitance $C_{r}$ and proportional to the resonance inductance $L_{r}$. Increasing the $Q$ value will increase the resonant inductance $L_{r}$, thereby increasing the excitation inductance $L_{m}$, which can reduce the current flowing into the resonant tank, reduce the switching transistor losses, and improve the efficiency of the resonant converter. However, too large $Q$ value results in a too narrow gain range to meet the demand. Therefore, the selection of $Q$ value should be as high as possible within the range of required gain.

![Figure 6. DC gain at different quality factors](image)

4. Parameter design
The LLC resonant converter mainly controls the output voltage through the resonant characteristics of the resonant tank[13]. Therefore, the design of the parameters of each component of the resonant tank is extremely important. The resonant capacitor mainly blocks DC in the circuit and participates in resonance. It can pass the quality factor $Q$, resonant frequency $f_{0}$, equivalent The impedance $R_{ac}$ is obtained as[14]:

$$C_{r} = \frac{1}{2\pi Q f_{0} R_{ac}}$$

(16)

Calculate the resonant capacitance $C_{r}$ to derive the resonant inductance:

$$L_{r} = \frac{1}{(2\pi f_{0})^2 C_{r}}$$

(17)

The excitation inductance can be obtained by the inductance:

$$L_{m} = kL_{r}$$

(18)

The dead time of the switch transistor affects the voltage stress of the switch transistor and the realization of the zero voltage switch (ZVS) of the switch transistor.

After the switch $Q_{2}Q_{7}$ is turned off, the junction capacitor $C_{2}C_{7}$ starts to charge and discharge $C_{3}C_{6}$. When the voltage of $C_{3}C_{6}$ drops to 0, $Q_{3}Q_{6}$ is turned on again to realize the zero voltage turn on. Therefore, the dead time between $Q_{2}Q_{7}$ and $Q_{3}Q_{6}$ must be longer than the charging and discharging time of its upper junction capacitor.

Junction capacitance discharge time:

$$t_{s} = 8C_{i} L_{m} f_{s}$$

(19)

Where $C_{i}$ is the junction capacitance of the switch; $L_{m}$ is the excitation inductance; $f_{s}$ is the switching frequency;

The dead time of $Q_{2}Q_{7}$ and $Q_{3}Q_{6}$ of the switch transistor also needs to be less than the zero-crossing time of the primary current. The reverse of the primary current will recharge the junction
capacitance after the discharge is completed, resulting in the failure to reach the zero-voltage switching (ZVS) condition. Calculate the time when the primary current reverses zero crossing:

\[
t_f = \frac{\tan^{-1}\left(\frac{n^2 V_o}{2\pi f_0 P_o V_o L_m f_0}\right)}{2\pi f_0}
\]  

(20)

Where \( t_f \) is the zero-crossing time of the primary current; \( n \) is the transformer ratio; \( V_o \) is the output voltage; \( P_o \) is the output power; \( f_s \) is the operating frequency; \( f_0 \) is the resonance frequency;

When the resonance current is not equal to the excitation current, the switch transistor is turned off, then the switch transistor cannot achieve zero current shutdown (ZCS), so the dead time should also be less than the difference between the resonance half cycle and the switch half cycle:

\[
t_h = \frac{1 - f_0}{2 f_s^2}
\]  

(21)

Simultaneous (19), (20), (21) can get the dead time conditions:

\[
t_s \leq t_d \leq t_s + t_f \text{ and } t_d \leq t_h
\]  

(22)

Switching transistors \( Q_1Q_8 \) and \( Q_4Q_5 \) need to be considered for early turn-off time. Switching transistor \( Q_2Q_7 \) is turned off in advance to charge junction capacitor \( C_1C_8 \) and discharging junction capacitor \( C_4C_5 \). \( C_4C_5 \) has been discharged to 0 before the switching transistor \( Q_2Q_7 \) is turned off. The early turn-off time is simplified. It is the dead time \( t_d \) of \( Q_2Q_7 \) and \( Q_4Q_5 \).

5. Simulations

Design a high-power three-level full-bridge LLC resonant converter according to the above method, the specific parameters are as table 1:

The method of fixed duty cycle and conduction duty cycle is adopted, and the working frequency is changed by changing the working cycle length of the switch as a whole. This method will reduce the dead time when the frequency is higher. Let the converter lose the soft switching condition, but only need to account for the maximum operating frequency, the dead time can be greater than the charge and discharge time of the parallel capacitance of the converter switch transistor to ensure soft switching. This article uses simulink’s voltage-frequency module and comparator to achieve.

| Algorithm                  | Value          |
|---------------------------|----------------|
| Rated input voltage \( V_{\text{nom}} \) | 1000V          |
| Rated output power \( P_{\text{nom}} \) | 10kW           |
| Transformer ratio \( n \)   | 2.1            |
| Maximum voltage gain \( G_{\text{max}} \) | 1.11           |
| Minimum voltage gain \( G_{\text{min}} \) | 0.91           |
| Inductance \( k \)          | 5              |
| Quality factor \( Q \)      | 0.36           |
| Equivalent resistance \( R_{\text{dc}} \) | 80.59Ω         |
| Resonant inductance \( L_r \) | 46.17uH        |
| Resonant capacitor \( C_r \) | 54.86nF        |
| Excitation inductance \( L_m \) | 230.86uH       |
| Operating frequency range   | 70-140kHz      |
6. Results and discussion

Figure 7 is the drain voltage $V_{ds}$ voltage and the gate voltage $V_{gs}$ of the switch transistor when the input voltage is rated at 1000V, the load is 23Ω, and the output voltage is 480V. From top to bottom are the switch transistors $Q_2Q_7$, $Q_1Q_8$, $Q_3Q_6$, $Q_4Q_5$, respectively. The withstand voltage stress never exceeds half of the input voltage $V_{i,nom}$ and the magnitude is about 500V. When the voltage of the switching transistor ($V_{ds}$) drops to 0, the driving voltage acts on the switching transistor ($V_{gs}$) again. The transistor $Q_1Q_8$ and $Q_4Q_5$ needs to be turned off in advance, so the dead time is longer than $Q_2Q_7Q_3Q_6$, but it can be seen from Figure 7 that each switch transistor can achieve zero voltage switching.

![Figure 7. Waveform diagram of drain voltage $V_{ds}$ and gate voltage $V_{gs}$](image)

Figure 8 is a waveform diagram of the excitation current $i_{Lr}$ and the excitation current $i_{Lm}$ of the resonant tank when the operating frequency is about 130 kHz. It can be seen that the amplitude of the resonant current $i_{Lr}$ is about ±20A, and the excitation current $i_{Lm}$ is about ±8A. When $i_{Lr} = i_{Lm}$, the three-element resonance occurs in the resonant tank, and the secondary diode realizes zero current shutdown (ZCS).

![Figure 8. Resonance current and excitation current waveform](image)

Figure 9 shows the output voltage and output current waveforms. You can see that the output current is about 21A, the output voltage is about 480V, and the output power is 10kW.

![Figure 9. Output voltage and current waveforms](image)
7. Conclusions

This paper designs a three-level full-bridge LLC resonant converter based on frequency modulation control, which is suitable for high power and high voltage environment. Through the analysis of typical topology, the equivalent circuit in each mode is established, the design scheme is given in combination with the gain characteristics, and the detailed design process of each parameter is given.

The three-level full-bridge LLC resonant converter designed in this paper has the following advantages: 1. The voltage stress of each switch on the primary side is half of the input voltage, and both can achieve ZVS. 2. The secondary diode can achieve ZCS, and the output current is stable, which meets the design requirements. 3. Through the analysis of the DC gain of the LLC circuit, the values of the quality factor $Q$ and the inductance coefficient $k$ are discussed. 4. Simple topology and high output efficiency.

The simulation in simulink verified the feasibility of the parameter design scheme. In the input voltage DC1000V and output voltage DC480V systems, the switch transistors have achieved soft switching characteristics, and the output efficiency is as high as 96.5%.

References

[1] Haga H and Kurokawa F 2015 A novel modulation method of the full bridge three-level LLC resonant converter for battery charger of electrical vehicles IEEE Energy Conversion Congress and Exposition 5498-504
[2] Li H, Zhang Z, Wang S, Tang J, Ren X and Chen Q 2020 A 300-kHz 6.6-kW SiC Bidirectional LLC Onboard Charger IEEE Trans. Ind. Electron. 67(2) 1435-45
[3] Lee I and Moon G 2012 Analysis and Design of a Three-Level LLC Series Resonant Converter for High- and Wide-Input-Voltage Applications IEEE Trans. Power Electron. 27(6) 2966-79
[4] Liu J, Zhang J, Zheng T Q and Yang J 2017 A Modified Gain Model and the Corresponding Design Method for an LLC Resonant Converter IEEE Trans. Power Electron. 32(9) 6716-27
[5] Mohammadi M and Ordonez M 2019 Synchronous Rectification of LLC Resonant Converters Using Homopolarity Cycle Modulation IEEE Trans. Ind. Electron. 66(3) 1781-90
[6] Zhou D, Zhang X, Liu X, Wang Y, Xu D and Tian H 2018 Design of LLC Converter Parameters Based on ZVS Characteristic Analysis International Conference on Electrical Machines and Systems 2251-55
[7] Rahman A N, Chiu H and Hsieh Y 2018 Design of wide input voltage range high step-up DC-DC converter based on secondary-side resonant tank full bridge LLC International Conference on Intelligent Green Building and Smart Grid 1-6
[8] Gao T, Cheng Z, Wang Q, Li Y, Zeng K and Yang Y 2019 The Analysis of Dead time's Influence on the Operating Characteristics of LLC Resonant Converter IEEE Conference on
[9] Wang S, Zheng Z, Li C, Xu L, Wang K and Li Y 2019 Accurate frequency-domain analysis and hybrid control method for isolated dual active bridge series resonant DC/DC converters IET Power Electronics 12(11) 2932-41

[10] Wei Y, Luo Q and Mantooth A 2019 Comprehensive analysis and design of LLC resonant converter with magnetic control CPSS Transactions on Power Electronics and Applications 4(4) 265-75

[11] Li Z, Lin H, Mao W, Cai C, Guo X and Shu Z 2018 Optimized Parameter Design of An Improved LLC Resonant Converter IEEE International Power Electronics and Application Conference and Exposition 1-6

[12] Wang C, Zhang S, Wang Y, Chen B and Liu J 2019 A 5-kW Isolated High Voltage Conversion Ratio Bidirectional CLTC Resonant DC–DC Converter With Wide Gain Range and High Efficiency IEEE Trans. Power Electron. 34(1) 340-55.

[13] Lee Y, Hong S, Hyun S and Won C 2016 Design of Phase Control LLC resonant converter for fast load transient IEEE Transportation Electrification Conference and Expo Asia-Pacific 793-7

[14] Zhu T, Ji Y, Wang J, Liu Y and Gao C 2018 LLC Converter's Design Method Based on Simulation Analysis IEEE Advanced Information Management Communicates Electronic and Automation Control Conference 1996-2000