Research Article

The Phase Shifting Soft Startup of L-LLC Resonant Bidirectional DC-DC Converter Based on Current-Limiting Curve

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The new L-LLC resonant bidirectional DC-DC converter (L-LLC-BDC) will produce a large resonance current and voltage inrush during the startup, posing a threat to the safe operation of the power device. Although a very high starting frequency can effectively suppress the inrush, it will also increase the output current demand of the driving ICs. This paper proposes a phase-shifting soft-start control strategy based on the current-limiting curve. Using operating mode analysis, the peak value of the resonant current is limited according to the output voltage and the phase shift angle of the switch, with the limit current curve at the startup stage drawn. By this current curve, a one-to-one correspondence between the output voltage and the phase shift angle of the switch is obtained. The phase-shifted soft-start control strategy can quickly establish the output voltage on the basis of a resonant frequency and can effectively suppress the resonance current inrush. An experimental prototype with a power of 6 kW and an input of 760 V and an output of 380 V is built. The experimental results prove the correctness and effectiveness of the soft start control strategy proposed in this paper.

1. Introduction

Due to LLC resonant converter's high efficiency, high power density, and wide gain and power, there are many researchers focusing on it [1–3]. Twelve switches are used in the LLC converter in literature [4], the cost of the converter is increased, and it also reduces the power density because of the two transformers. Literature [5] proposed an LLC converter with two interleaved pulse-width rectifiers, and it only operated with forward power transmission. The CLLC resonant converter is proposed in [6] for bidirectional power flow. However, the voltage gain of the converter is not monotonous in inductive region under heavy load. In literature [7], an additional inductance is added to the traditional LLC resonant converter to form an L-LLC resonant bidirectional DC-DC converter. Eight switches are used in this converter, and the converter can be operated at both boost and buck voltage gain with forward and reverse transmission. The efficiency of the topology used in this paper is up to 97% by using synchronous variable width variable frequency control. This paper employs the topology

from literature [7] as shown in Figure 1. It can be used as the key equipment for interaction between ESSs and the DC microgrid for stabilization of the DC bus voltage. Figure 2 shows the DC characteristics of the converter with different loads. In order to obtain the highest working efficiency, the converter generally works around the switching frequency f equal to the resonance frequency f_r , i.e., the normalized frequency $f_n = f/f_r = 1$. The voltage gain M is the ratio of the output voltage to the input voltage of the converter, and M is 1. When the input voltage decreases, the converter will work in the $f_n < 1$, with M being greater than 1 to keep the output voltage constant. Conversely, when the input voltage increases, the converter works in the $f_n > 1$, so as to prevent the converter from entering the zero current (ZCS) region during startup process, and the switching frequency at the initial stage of startup must be greater than the resonance frequency [8, 9]. When the switching frequency in the startup is not large enough, that is, when the maximum switching startup frequency $f_{\max} = 1.5f_r$, a large voltage and current inrush will occur as shown in Figure 3. When the switching frequency in the startup phase is large enough,

FIGURE 1: L-LLC-BDC circuit topology.

FIGURE 2: DC characteristics of L-LLC-BDC.

$f_{\max} = 5f_r$. Although there is no large voltage and current stress at the initial stage of the start-up process, too large voltage and current surge will still occur if the switching frequency is reduced too quickly, and if the switching frequency decreases too slowly, the output voltage will be set up too slowly.

The pulse frequency modulation (PFM) control strategy is adopted, and the relationship between f_n and maximum primary current $i_{p\max}$ with different filter parameters is plotted in the soft startup as shown in Figure 4. The maximum resonant current will not decrease as the switching frequency increases when $f_n > 3$. The minimum switching start frequency can be determined if the maximum resonant current is set. Therefore, the switching frequency is relatively large as long as the PFM control strategy is used during soft startup.

At a lower switching frequency, no large voltage and current stress occur during the rapid establishment of the output voltage, which is a key issue for the soft startup of the converter. The phase shift control is widely used in resonant converters [10–14]. In literature [14], a variable duty cycle soft start control strategy is proposed, but the mode prediction calculation by this method is too complicated. In literature [15], a two-step method is proposed. In the soft startup, the two switches S_3 and S_4 are forcibly turned off to make the converter work on the half-bridge, and the full-bridge working state is restored once the output voltage reaches the steady-state. However, the experimental results show that inrush current still occurs during the startup process, and that the startup frequency is also very high. Reference [16] only proposed a three-step control strategy

FIGURE 3: Soft startup at a low switching frequency.

for a boost PFC rectifier, which has limited application. In literature [17], the optimal trajectory soft startup control strategy based on microcontroller, such as DSP is proposed. This method eliminates the voltage and current surges during the startup process, with its essence being PFM control and the startup frequency being still relatively high. In literature [18–24], the control process is too complex, and the time of setting up the output voltage is too slow. Reference [25] uses analog circuit to realize soft start, and although the starting speed of output voltage is fast and the current inrush is small, it is not desirable for resonant converter with high power density. Reference [26] proposes a three-step soft-shift start modulation technique that allows to limit the inrush current in the DC/DC isolation stage during the DC-link capacitors precharging for a three-stage smart transformer, but it is for nonresonant circuits only.

This paper proposes a phase-shifting soft-start control strategy based on the current-limiting curve. Using the operating mode analysis [27–30], the peak value of the primary current i_p of the converter is limited according to the output voltage and the phase-shifting angle of the switch, and it is not necessary to predict the operating mode of the converter to obtain the current peak, thereby obtaining the limiting current curve in the startup. Based on this current curve, a one-to-one correspondence between output voltage and phase-shifting angle is obtained. The output voltage can be quickly established on the basis of a lower starting frequency, and the resonance current surge can be effectively suppressed.

2. Working Principle of Bidirectional L-LLC-BDC

Figure 1 shows the circuit topology of L-LLC-BDC. The primary side and secondary side of the converter adopt a full-bridge structure, and V_{in} of the converter is the input voltage of the interface unit of the energy storage device. nV_o is the DC bus voltage, L_{m1} is the magnetizing inductance, L_r is the resonant inductance, C_r is the resonant capacitance, and L_{m2} is the additional inductance. The switches S_1 – S_4 are primary side switches and S_5 – S_8 are secondary side switches. V_{Lm1} is the voltage of the magnetizing inductance L_{m1} when the converter works in the forward direction. When the

FIGURE 4: Relationship between startup current and frequency with different output filter parameters.

magnetizing inductance is clamped at the output voltage nV_0 , $U_{Lm1} = nV_0$. When the magnetizing inductance is clamped reversely at the output voltage nV_0 , $U_{Lm1} = -nV_0$. When the converter is in the freewheeling stage, that is, the LLC resonance, $U_{Lm1} < nV_0$. The voltage of the additional inductance L_{m2} is the same as L_{m1} with the reverse power transmission. The forward operation of the converter is the discharge mode, and the reverse operation is the charge mode. If the additional inductance of L_{m2} is equal to the magnetizing inductance of L_{m1} , the forward and reverse working principles of the converter are exactly the same, so this article only analyzes the forward working principle of the converter. The switches S_1 – S_4 form an inverter network, L_{m2} , L_{m1} , L_r , and C_r form a resonant network, and the antiparallel diodes D_{S5} – D_{S8} of the switches S_5 – S_8 form a rectifier network. When the converter is working in the forward direction, S_1 and S_4 form the leading bridge arm, S_2 and S_3 form the lagging bridge arm, and any switching is a complementary drive signal with the same frequency and a 50% duty cycle. The output voltage V_o is modulated by adjusting the phase shift angle between the leading arm and the lagging arm. θ is the phase-shifting angle, $\theta \in [0, \pi]$, and f_r is the resonant frequency. The additional inductance L_{m2} never participates in resonance when the converter is working in the forward direction, but helps the primary side switch to achieve ZVS. In the reverse operation of the converter, the additional inductance L_{m2} forms part of the resonant network, and the magnetizing inductance L_{m1} never participates in resonance, helping the secondary side switch to achieve ZVS.

L-LLC-BDC has multiple operating modes under different phase-shift angles with the phase-shift control strategy. The working principle of the first half cycle and the second half cycle is the same, so only the working state of the

first half cycle is analyzed, without considering the dead time. The operating modes of this converter are as follows.

2.1. Mode P. $v_{AB} = V_{in}$. Switches S_1 and S_4 are turned on, the primary current i_p continues to flow through D_{S1} and D_{S4} , the output capacitors C_{S1} and C_{S4} of the switch are discharged to zero voltage, and switches S_1 and S_4 realize ZVS. The voltage across the two points A and B which is V_{in} , i_p , i_{Lm1} , and i_{Lm2} begins to increase. The body diodes D_{S5} and D_{S8} of the secondary side switches S_5 and S_8 are turned on with i_p increasing faster. The C-D two-point voltage is clamped at the output voltage V_o , i_p flows through switches S_1 and S_4 to transfer energy to the load. It is worth noting that, in this mode, the primary current $i_p = i_{Lr} + i_{Lm2}$, and the secondary current $i_s = ni_{Lr} - ni_{Lm1}$.

2.2. Mode O. $v_{AB} = V_{in}$. The resonant current i_{Lr} is equal to the magnetizing current i_{Lm1} , and the secondary side current is zero. The body diodes D_{S5} and D_{S8} of switches S_5 and S_8 are naturally turned off because the current is zero, with no reverse recovery loss, and the ZCS of the secondary side switch tube is realized. The output voltage is no longer clamped to the C-D two points, and the output capacitors C_{S5} to C_{S8} of switches S_5 and S_8 participate in resonance. This stage is actually a freewheeling stage.

2.3. Mode P_0 . $v_{AB} = 0$, and the operating mode is similar to the P mode.

2.4. Mode O_0 . $v_{AB} = 0$, and the operating mode is similar to the O mode.

2.5. *Mode N_0* . $v_{AB} = 0$. Switches S_1 and S_4 are turned on, the primary side current i_p continues to flow through D_{S1} and D_{S2} , C_{S1} and C_{S4} discharge to zero voltage, and the switches S_1 and S_4 realize the ZVS. The body diodes D_{S6} and D_{S7} of the secondary side switches S_6 and S_7 are turned on, and the voltage across C and D is clamped to the output voltage $-V_o$. The equivalent circuit of the five modes of the converter is shown in Figure 5.

3. Time-Domain Equation of Bidirectional L-LLC-BDC

The time-domain equations of the above five mode equivalent circuits can be expressed as follows.

3.1. *Mode P*. By normalizing all voltages with voltage factor nV_0 , the normalized all currents with the current factor $I = nV_0/Z_r$.

$$\begin{cases} i_{LrP}(\omega t) = I_{LrP} \sin[\omega(t + t_{P0})], \\ i_{PP}(\omega t) = I_{PP} + \frac{\omega t}{k}, \\ u_{CrP}(\omega t) = -I_{LrP} \cos[\omega(t + t_{P0})] + \frac{1}{M} - 1, \\ i_{Lm1P}(\omega t) = I_{Lm1P} + \frac{\omega t}{kM}, \end{cases} \quad (1)$$

where Z_r is the characteristic impedance, $Z_r = \sqrt{L_r/C_r}$, $k = L_{m1}/L_r$, ω is the resonance angular frequency $\omega = 2\pi f$, M is the voltage gain $M = nV_o/V_{in}$, $0 \leq \omega t \leq \omega t_p$, and I_{LrP} , I_{Lm1P} , I_{PP} , ωt , ωt_{P0} , and M are unknown quantities.

3.2. *Mode O*

$$\begin{cases} i_{LrO}(\omega t) = i_{PO} = I_{LrO} \sin\left[\omega\left(\sqrt{\frac{1}{1+k}}t + t_{O0}\right)\right], \\ i_{Lm1O}(\omega t) = I_{Lm1O} + \frac{\omega t}{kM}, \\ u_{CrO}(\omega t) = -I_{LrO} \cos\left[\omega\left(\sqrt{\frac{1}{1+k}}t + t_{O0}\right)\right] + \frac{1}{M}, \\ u_{Lm1O}(\omega t) = \frac{(1/M) + u_{CrO}(\omega t)}{1+k}, \end{cases} \quad (2)$$

where $0 \leq \omega t \leq \omega t_O$, I_{LrO} , I_{Lm1O} , ωt , ωt_{O0} , and M are unknown quantities.

3.3. *Mode P_0*

$$\begin{cases} i_{LrP_0}(\omega t) = I_{LrP_0} \sin[\omega(t + t_{P_00})], \\ i_{PP_0}(\omega t) = I_{Lm1P_0} + \frac{\omega t}{k}, \\ u_{CrP_0}(\omega t) = -I_{LrP_0} \cos[\omega(t + t_{P_00})] - 1, \\ i_{Lm1P_0}(\omega t) = I_{PP_0}, \end{cases} \quad (3)$$

where $0 \leq \omega t \leq \omega t_{P_0}$, I_{LrP_0} , I_{Lm1P_0} , ωt , ωt_{P_00} , and M are unknown quantities.

3.4. *Mode N_0*

$$\begin{cases} i_{LrN_0}(\omega t) = I_{LrN_0} \sin[\omega(t + t_{N_00})], \\ i_{PN_0}(\omega t) = I_{Lm1N_0} - \frac{\omega t}{k}, \\ u_{CrN_0}(\omega t) = -I_{LrN_0} \cos[\omega(t + t_{N_00})] + 1, \\ i_{PN_0}(\omega t) = I_{PN_0}, \end{cases} \quad (4)$$

where $0 \leq \omega t \leq \omega t_{N_0}$, I_{LrN_0} , I_{Lm1N_0} , ωt , ωt_{N_00} , and M are unknown quantities.

3.5. *Mode O_0*

$$\begin{cases} i_{LrO_0}(\omega t) = i_{PO_0} = I_{LrO_0} \sin\left[\omega\left(\sqrt{\frac{1}{1+k}}t + t_{O_00}\right)\right], \\ u_{Lm1O_0}(\omega t) = \frac{-u_{Cr}(\omega t)}{1+k}, \\ u_{CrO_0}(\omega t) = -I_{LrO_0} \cos\left[\omega\left(\sqrt{\frac{1}{1+k}}t + t_{O_00}\right)\right], \\ i_{Lm1O_0}(\omega t) = I_{Lm1O_0}, \end{cases} \quad (5)$$

where $0 \leq \omega t \leq \omega t_{O_0}$, I_{LrO_0} , I_{Lm1O_0} , ωt , ωt_{O_00} , and M are unknown quantities.

L-LLC-BDC has 6 operating states formed by connecting these five modes in different orders with the increase of phase-shift angle θ , with the operating states being $PP_0O_0N_0$, PP_0O_0 , OPP_0O_0 , $OPOO_0$, P , and OPO . The time-domain waveform is shown in Figure 6.

FIGURE 5: Simplified equivalent circuits with L-LLC-BDC under different modes. (a) Mode P. (b) Mode O. (c) Mode P₀. (d) Mode O₀. (e) Mode N₀.

3.5.1. State PP₀O₀N₀. The converter enters the PP₀O₀N₀ state when the phase-shift angle θ is small. As shown in Figure 6(a), starting with mode P, the converter enters P₀ when the voltage of input is zero. The secondary diodes are turned off and enter to the O₀ mode when the resonance current i_{Lr} is equal to the magnetizing current i_{Lm1} . When mode O₀ ends, L_{m1} falls down to $-V_o$, diodes D_{S7} and D_{S6} of the secondary side are turned on, and L_{m1} is clamped at $-V_o$. The positive half cycle ends when the resonant inductance and magnetizing inductance are equal again.

3.5.2. State PP₀O₀. Increasing the load on the basis of the PP₀O₀N₀ state, or further increasing the phase-shift angle, the converter starts the PP₀O₀ state in Figure 6(b). The PP₀O₀ state is the same as the first three modes of the PP₀O₀N₀ state.

3.5.3. State OPP₀O₀. In Figure 6(c), with the increase of phase-shift angle θ , the converter enters the OPP₀O₀ state, but this state only exists under light load conditions. The magnetizing inductance voltage V_{Lm1} does not have enough

FIGURE 6: Continued.

FIGURE 6: Time-domain waveforms for L-LLC-BDC in different states. (a) $PP_0O_0N_0$. (b) PP_0O_0 . (c) OPP_0O_0 . (d) $OPOO_0$. (e) P. (f) OPO.

capacity to start the P mode, and therefore, the converter starts with the O mode and then enters the P mode. After the end of the P mode, the converter enters the P_0 mode when the voltage of input is zero and ends in the O_0 mode.

3.5.4. State $OPOO_0$. With the increase of phase-shift angle θ , the converter enters the $OPOO_0$ state as shown in Figure 6(d), and this state only exists under light load conditions too, which is similar to the first two modes of the OPP_0O_0 state in Figure 6(c). After the end of the P mode, the converter enters the O mode and ends in the O_0 mode when the voltage of input is zero.

3.5.5. States P and OPO. When the phase-shift angles θ increases to the maximum value π , the converter has two operating states, respectively, as shown in Figures 6(e) and 6(f). The P state works at the resonant frequency point, which is the most ideal working state after the converter enters the steady state. The OPO operating state exists under the light load condition when the phase shift angle is π . At this time, the magnetizing inductance voltage V_{Lm1} does not have enough capacity to start the P mode, and therefore, the converter works in mode O. After mode P, the magnetizing inductance voltage V_{Lm1} is not low enough to enter mode N, which ends in mode O.

The converter operates at the above six working states, and its time domain equations must satisfy the following stepwise continuous conditions and half-period symmetry conditions. AB is the two modes connected in a sequence in

a certain operating state. Hh is a certain operating state at the beginning and end of the first half of the cycle, as shown in equations (6) and (7).

$$\begin{cases} i_{LrA}[\omega(t+t_A)] = i_{LrB}(0), \\ i_{Lm1A}[\omega(t+t_A)] = i_{Lm1B}(0), \\ u_{CrA}[\omega(t+t_A)] = u_{CrB}(0), \\ i_{pA}[\omega(t+t_A)] = i_{pB}(0), \end{cases} \quad (6)$$

$$\begin{cases} -i_{LrH}[\omega(t+t_H)] = i_{Lrh}(0), \\ -i_{Lm1H}[\omega(t+t_H)] = i_{Lm1h}(0), \\ -u_{CrH}[\omega(t+t_H)] = u_{Crh}(0), \\ -i_{pH}[\omega(t+t_H)] = i_{ph}(0). \end{cases} \quad (7)$$

According to the mode analysis, the converter only transfers energy to the load in the P, P_0 , and N_0 stages. In mode O and mode O_0 , the secondary side discontinues, and there is no energy transfer. Therefore, the power balance conditions are

$$\begin{aligned} \frac{Z_r P_O}{n^2 V_O^2} = & \frac{1}{\pi} \left[\int_0^{\theta_P} (i_{LrP} - i_{Lm1P}) d\theta + \int_0^{\theta_{P_0}} (i_{LrP_0} - i_{Lm1P_0}) d\theta \right. \\ & \left. + \int_0^{\theta_{N_0}} (i_{Lm1N_0} - i_{LrN_0}) d\theta \right]. \end{aligned} \quad (8)$$

It can be seen from the running state of the converter that mode O exists before and after mode P, and we get

$$u_{nV_O}(\theta_O) = 1. \quad (9)$$

Mode O_0 is connected to mode N_0 , and at the end of mode O_0 , we get

$$u_{nV_O}(\theta_{O0}) = -1. \quad (10)$$

When the input voltage is $V_{in} = V_{AB}$, the phase-shifting angle θ of mode P and mode O is the same as the operating angle of the input voltage. It yields

$$\theta_P + \theta_O = \theta. \quad (11)$$

The operating angle of each stage in the first half of the cycle is the sum of the operating angles of each stage. It yields

$$\theta_H = \pi. \quad (12)$$

Given power P and phase-shifting angle θ , each mode equation system can be solved according to the above conditions. For the P and OPO states, the phase-shifting angle at this time is π , and the frequency is the resonant frequency. Therefore, the above equations can be used to describe the operating state of L-LLC-BDC accurately during the soft startup.

According to the equations and constraints of the six operating states of the converter, the voltage gain characteristics of the converter can be obtained, as shown in Figure 7. At a fixed transmission power, when the phase-shifting angle θ increases from 0 to π , the voltage gain M of the converter will also increase monotonously.

Since there is a one-to-one correspondence between phase-shifting angle θ and the voltage gain M , the limiting current is obtained by a numerical algorithm during soft startup, as shown in Figure 8. V_s is the steady-state output voltage. An analysis of Figure 6 shows that the current value of the P mode is always the largest in each operating state of the converter. Therefore, it is only necessary to calculate the peak resonance current i_p of mode P corresponding to the output voltage V_o and the phase-shifting angle θ in the six operating states. A series of data of the phase-shifting angle θ corresponding to the output voltage V_o are solidified in the DSP, with 1024 points collected. An appropriate phase-shifting angle θ can be found according to the output voltage V_o by looking up the table during the soft startup. In actual engineering, a reasonable limiting current value i_{lim} can be designed. Figure 9 indicates the relationship between output voltage and phase-shifting angle when the limit current is 17 A.

In order to analyze the dynamic process between hard start and soft start, the current variation curves of the converter in hard startup and soft startup are obtained by simulation, as shown in Figure 10. At 0.1 s, it can be seen that the current at the primary side of the converter increases instantaneously during the hard startup. The primary side current has reached about 25 A at 0.2 s when the phase-

shifting soft startup starts at 0.1 s. Therefore, both hard startup and phase-shifting soft startup are not the best choice for converter. In this paper, limiting current is added on the basis of phase-shifting soft startup. The limiting current is 1.1 times of the maximum peak current, and it is up to 17 A at 0.2 s. This control strategy enables the converter to establish the output voltage quickly and smoothly on the basis of small starting current.

4. Experimental Verification

In order to verify the soft start-up control strategy proposed in this paper, the experimental device of L-LLC resonant bidirectional DC-DC converter is designed, as shown in Figure 11.

The output voltage V_o is 380 V, the device varies from 630 V to 890 V in input voltage, the rated input voltage of the device V_{in} is 760 V, and the resonance frequency f_s is 100 kHz. The operating frequency range is 80–120 kHz. The model parameters for the L-LLC-BDC converter are shown in Table 1. The comparison between hard start, phase-shifting soft start, and phase-shifting soft start control strategy based on limiting current is shown in Figures 12–16.

Due to the presence of the transient process of hard start in experiment, the input voltage is set to 200 V. We can see that the current spike of the converter under no-load has reached nearly 55 A as shown in Figure 12. The higher the voltage, the greater the current inrush.

The peak value of the resonance current at the resonance operating point is

$$i_p = \frac{V_O}{4nR_L} \sqrt{\frac{n^4 R_L^2 T^2}{L_{m1}^2} + 4\pi^2} + \frac{\sqrt{2}TnV_O}{4\sqrt{3}L_{m2}} = 15.2(A). \quad (13)$$

The peak value of the resonance current under full-load conditions is calculated by each operating state. The peak value of the resonance current is obtained by equation (13), which is 15.2 A. Resonant current is limited according to 1.1 times of peak current, and the value of resonant current is 17 A. Figures 13 and 14 show the comparison between phase-shifting soft start and phase-shifting soft start control strategy based on limiting current under full load conditions when the input voltage is 760 V. The current of the primary side i_p is limited to 17 A in Figure 13, and it is up to 25 A in Figure 14. The starting current is reduced significantly by the phase-shifted soft-start based on limiting current control strategy, and the time of setting up the output voltage V_O is within 12 ms. Figures 15 and 16 are the waveforms of the current i_p and output voltage V_O under no-load condition. Similarly, the resonant current is effectively suppressed, and the startup time is shorter in Figure 15. However, the problem of output voltage floating high will appear during no-load soft startup, it can be solved by burst-mode control.

FIGURE 7: Voltage gain characteristics of L-LLC-BDC.

FIGURE 8: Numerical algorithm for the current-limiting curve of L-LLC-BDC.

FIGURE 9: Current limit curve during soft startup.

FIGURE 10: The current curve of I_p under three conditions.

FIGURE 11: Physical photo of test circuit topology.

TABLE 1: The model parameters of the converter.

Symbol	Quantity	Parameter
S	SiC MOSFET	C2M0080120D
V_{in}	Primary side rated voltage	760 V
V_o	BDC secondary side rated voltage	380 V
P	BDC rated power	6 kW
L_{m1}	BDC magnetic inductance	0.614 mH
n	BDC transformer ratio	2 : 1
L_r	BDC resonant inductance	0.087 mH
C_r	BDC resonant capacitance	29.1 nF
L_{m2}	BDC additional inductance	0.614 mH
f	Resonant frequency	100 kHz

FIGURE 12: Waveforms of hard startup.

FIGURE 13: The experiment waveforms of phase-shift soft startup based on current-limiting under full load. (a) The waveforms of output voltage and primary current. (b) The detail waveforms of output voltage and primary current.

FIGURE 14: The experiment waveforms of phase-shift soft startup under full load. (a) The waveforms of output voltage and primary current. (b) The detail waveforms of output voltage and primary current.

FIGURE 15: The experiment waveforms of phase-shift soft startup based on current-limiting under no load. (a) The waveforms of output voltage and primary current. (b) The detail waveforms of output voltage and primary current.

FIGURE 16: The experiment waveforms of phase-shift soft startup under no load. (a) The waveforms of output voltage and primary current. (b) The detail waveforms of output voltage and primary current.

5. Conclusion

L-LLC-BDC will generate a large resonance current and voltage surge during the startup process, posing a threat to the safe operation of power devices. Although a very high starting frequency can effectively suppress the impact, it will also increase the output current demand of the ICs. This paper presents a phase-shifting soft-start control strategy based on the current-limiting curve. At the resonant frequency point, the peak value of the startup current is obtained according to numerical algorithm during the startup process. Through this current curve, a one-to-one correspondence between the output voltage and the phase shift angle of the switch is obtained. Experiments show that the control strategy can effectively suppress the resonant current, and the startup time is also shorter at the resonant frequency point.

Data Availability

The data used to support the findings of this study are available from the corresponding author upon request.

Conflicts of Interest

The authors declare that they have no conflicts of interest.

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