A Secondary-Side Semi-Active 3-Phase Interleaved Resonant Converter Employing Multi-Mode Modulation Scheme for Fast EV Charger Applications

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Abstract- **A secondary-side semi-active 3-phase interleaved (SS-SA3PI) resonant converter employing a multi-mode modulation scheme for fast EV charger applications is proposed. A multi-mode modulation scheme is used to widen the output voltage range without dynamic switching problems. By using secondary-side semi-active control, the resonant converter does not need to operate in the $f_s < f_r$ state to achieve boost operation. The minimum switching frequency of the converter is equal to the resonant frequency, which reduces the size of the magnetic components. Thus, the power density of the converter is improved. Compared with the traditional 3-phase interleaved LLC converter controlled by PFM, this converter widens the output voltage range without adding any circuit components. This paper introduces the control mode, working principle and characteristics of the proposed converter in detail. A 10kW experimental prototype is built to realize the output voltage of 100-1000V, which verifies the feasibility and effectiveness of the scheme.**[1]

Index terms- **Secondary-side semi-active 3-phase interleaved (SS-SA3PI) resonant converter, multi-mode modulation scheme, fast EV charger, wide output voltage range, power density.**

I. Introduction

Using electric vehicles (EV) to replace fuel vehicles is one of the effective ways to solve the problem of environmental pollution and greenhouse effect. However, an embarrassing thing about the popularity of EVs is that the charging speed of EVs is slower than the refueling speed of fuel vehicles. For the popularity of EVs, it is necessary to increase the charging power of the charger. The DC/DC part of the charger has become one of the key factors in determining the development of EV [1-2].

Due to the large output power of the DC/DC part, some topologies that can achieve zero voltage switching (ZVS) such as LLC and phase-shifted full-bridge (PSFB) are widely used. Among these topologies, the ability of LLC to achieve soft switching for higher efficiency over the full load range has received a lot of attention [3-7].

The current industrial 20 kW constant power charger has become mainstream, and its common DC/DC topology is a 3-level LLC input series output series/parallel (ISOS/P) structure using Si Mos [8-12]. The advantages of changing the rectifier side series or parallel structure through relays are as follows:

1) wide output voltage range and constant power output voltage range;
2) low voltage and current stress of rectifier diodes;
3) low hardware costs.

However, the implementation of 3-level LLC ISOS/P structure ZVS requires the switching tube body diode to continue the current. The poor Si Mos body diode will cause reverse recovery when the converter is lightly loaded, and the large di/dt will cause EMI problems, making the converter face stability and efficiency challenges. Besides, the complex structure and huge amount of circuit components make the layout of the 3-level LLC ISOS/P converter tight within a limited range, which exacerbates the difficulty of designing the internal air ducts, heat dissipation and circuitry of the charger. Even, it will affect the overall working stability of the converter.

With the application of the third generation semiconductor device SiC in 5G communication, artificial intelligence and EVs, its cost is gradually reducing. SiC power devices have a smaller reverse recovery charge and are suitable for converters where body diodes require current continuation. And the high operating junction and breakdown field strength of SiC tubes

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means lower through-state losses and higher conversion efficiency. The current common 20kW EV charger (250-750V output) using SiC devices is a full-bridge parallel-tube LLC structure [13]. When the power of this structure needs to be designed larger, SiC devices with lower internal resistance need to be used, which means higher hardware costs. For this, the 3-phase interleaved LLC structure provides a solution. Compared with the full-bridge LLC structure, its advantages are as follows:

1) current stress is 2/3 of full-bridge structure;

2) transformer is Y-connected, automatic current equalization between phases;

3) same conversion efficiency as full-bridge structure in 30 kW;

4) low hardware costs.

However, limited control methods limit the voltage regulation capability of the 3-phase interleaved LLC structure. Refer to the classification and discussion of the full-bridge LLC modulation strategy in part [14]. Some modulation strategies that can be extended to the 3-phase interleaved LLC structure are discussed, although these extended applications may not be published. We divide the modulation strategies that can be applied into three categories:

1) primary-side inverter network modulation;

2) resonant network modulation;

3) secondary side rectification network modulation.

For primary-side inverter network modulation, the output voltage is regulated by adjusting the fundamental harmonic of the resonant network input voltage. There exist two freedoms to adjust the fundamental harmonic of resonant tank input voltage: magnitude and frequency. The use of pulse frequency modulation (PFM) is the most common control method for a 3-phase interleaved resonant converter [15]. However, PFM control is considered to be a major drawback of resonant converters [16]. Typically, when a 3-phase interleaved resonant converter using SiC devices has a wide output voltage range, the range of switching frequencies will also be wide, which will result in large magnetic component size and core losses. Besides, the inappropriate design will reduce the ZVS range and bring EMI problems. For magnitude modulation, common approaches are pulse width modulation (PWM) [17], primary side phase shift modulation (PSPSM) [18], asymmetric PWM [19] and variable input voltage control [20]. However, PWM type of control will lead to complex drive design, current and thermal imbalance. And PSPSM control cannot be extended to the 3-phase interleaved resonant converter because the phase angle of each phase is fixed at 120°. The literature [21] extends the method in the literature [17] to a three-phase interleaved structure, which makes the output voltage range of the converter wider. Although PFM control is used, the addition of an additional switching device and its corresponding heat sink cannot be avoided. For variable input voltage control, [22] is similar to the approach implemented in [20], which is particularly suitable for applications with two-stage structure, such as EV chargers, since the output voltage of the PFC is adjustable. This will bring us a positive meaning.

For resonant network modulation, the main control degrees of freedom are: resonant frequency and transformer turn ratio. The main modulation methods based on changing the resonant frequency are: variable inductor control [23], variable capacitor control [24], switched-control capacitor [25] and switched capacitor circuit [26]. And there are switching tubes or relays that are commonly used in industry to change the transformer turn ratio. Although all of these modulation strategies can greatly widen the regulation range of the output voltage, they do not avoid additional detection, resonance and switching components. Moreover, switching-type modulation will bring dynamic stability problems.

For secondary side rectification network modulation, two modulation strategies can widen the output voltage range, one is the secondary side PWM (SSPWM) [26], and another one is the secondary side PSM (SSPSM) [27-28]. In [26], an auxiliary switch is added on the secondary side. The use of PWM control for the auxiliary switches allows the converter to switch from the LLC structure to the voltage doubling rectifier structure. Therefore, the voltage gain can be adjusted from 1 to 2. However, SSPWM still adds an extra switch and the heat sink corresponding to this switch, reducing the power density of the converter. A control method of SSPSM is discussed in [27] and [28], which replaces the two lower diodes on the secondary side with two switches. By using SSPSM control, the converter obtains the boost effect. This modulation strategy adds no additional power components, allowing the converter to maintain a high power density and conversion efficiency. However, if applied to ultra-wide output voltage applications, a single control method will lose some ZVS range and create EMI problems. A time-delay control (TDC) similar to SSPSM is discussed in [29-32], which incorporates PSPFM control to further widen the output voltage regulation range. However, unlike SSPSM, TDC boosts require additional phase detection chips to achieve synchronous rectification. This undoubtedly poses potential stability problems. Moreover, the implementation of boost operation above the resonance point broadens the ZVS range of the converter but also results in a loss of voltage gain. All in all, SSPSM is a very potential regulatory strategy.

To meet the future requirements of higher power for EV chargers while achieving a wide output voltage range without reducing the power density of the converter, this paper proposes a SS-SA3PI resonant converter employing a multi-mode modulation scheme to solve the above problems. The main contributions of this paper are as follows:

1) The minimum switching frequency of the proposed SS-SA3PI resonant converter is the resonant frequency, which facilitates the reduction of the size of the magnetic components and thus increases the power density of the converter;

2) The employed multi-mode modulation scheme can enlarge the output voltage range to the greatest extent;

3) The converter also realizes a full range of soft switching, which is beneficial to reducing EMI and improving conversion efficiency. In addition, this paper also simplifies the formula derivation under the boost mode, making the control method

easy to implement in an online controller.

II. Proposed Converter and Working Principle

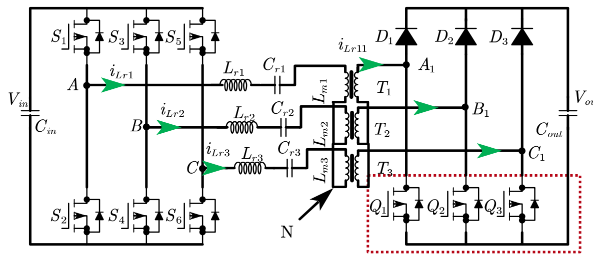


Fig.1 Proposed SS-SA3PI resonant converter.

The proposed SS-SA3PI resonant converter is shown in Fig. 1. A 3-phase bridge high-frequency inverter network consisting of 6 SiC Mos S_1 - S_6 on the primary side of the transformer. Three SiC Mos Q_1 - Q_3 and three diodes D_1-D_3 form a 3-phase semi-active bridge rectification network on the secondary side of the transformer. Two 3-phase bridge networks are connected by three identical sets of high-frequency transformers T and a three-phase staggered resonant network composed of L_r, C_r. The transformers are connected in a Y-shape to balance the currents between the bridge arms of each phase.

In the EV charger applications, it is generally a two-stage structure, and the output DC bus voltage of AC/DC converter is 660-850V at PF=1. Considering the actual application background of the proposed converter, the converter needs to output 100-1000V at the input voltage V_{in} is 660-850V to match different EV battery pack voltages. According to the < Qualification Capability Verification Standard for Electric Vehicle Charging Equipment Suppliers >, the converter is required to be capable of constant maximum power output at an output voltage of 400-750V. Thus, the maximum output current of the proposed converter is limited to 25A. Depending on the different controlled quantities of the converter, we divide the used multi-mode modulation scheme into 3 modes: 1) Boost mode (secondary-side semi-active control); 2) Variable bus voltage control mode; 3) Buck mode.

A. Working Principle of Boost Mode

The typical operating waveform of SS-SA3PI resonant converter under secondary-side semi-active control is shown in Fig. 2. The switching frequency f_s is equal to the resonant frequency f_r. Switching tubes S_1 - S_6, the upper and lower switching tubes of each bridge arm are on complementarily, and the duty cycle is fixed at 0.5. The bridge arms of each phase are staggered by 120° in one period. The basic principle of secondary-side semi-active control in boost mode is that: when S_1, S_4 and S_5 are on, the transformer secondary Q_1 is on and the secondary current flows positively through Q_1 and the parasitic diode along Q_2 short-circuits the transformer T_1, T_2 secondary winding. At this time, the input voltage V_{in} charges to the resonant network. When Q_1 is turned off, the secondary side of the converter continues to rectify the current. During rectification, the input voltage and the resonant network

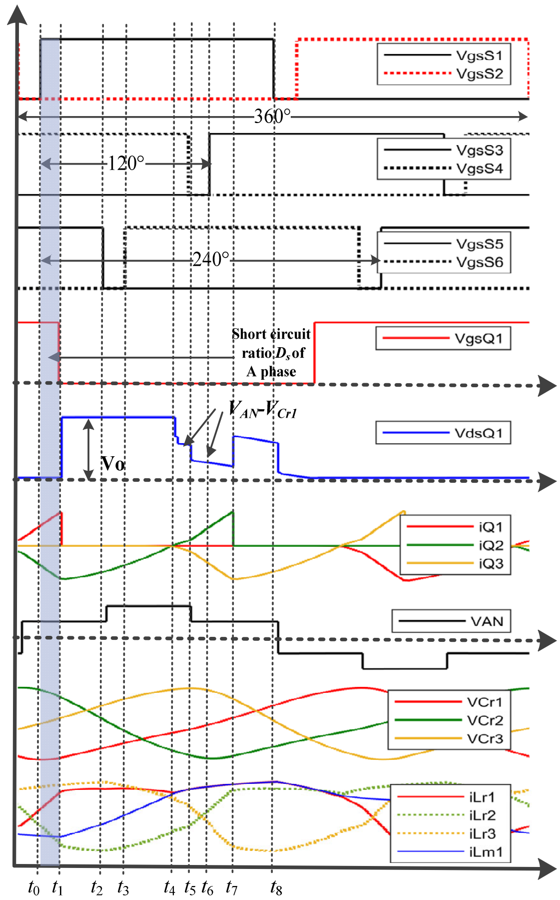
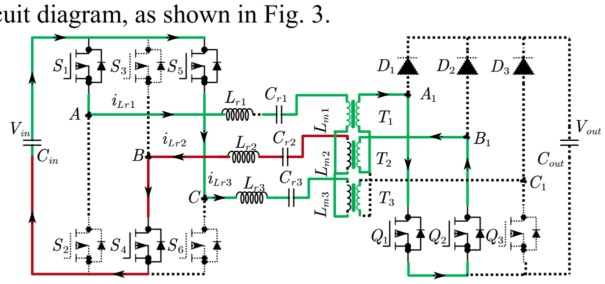
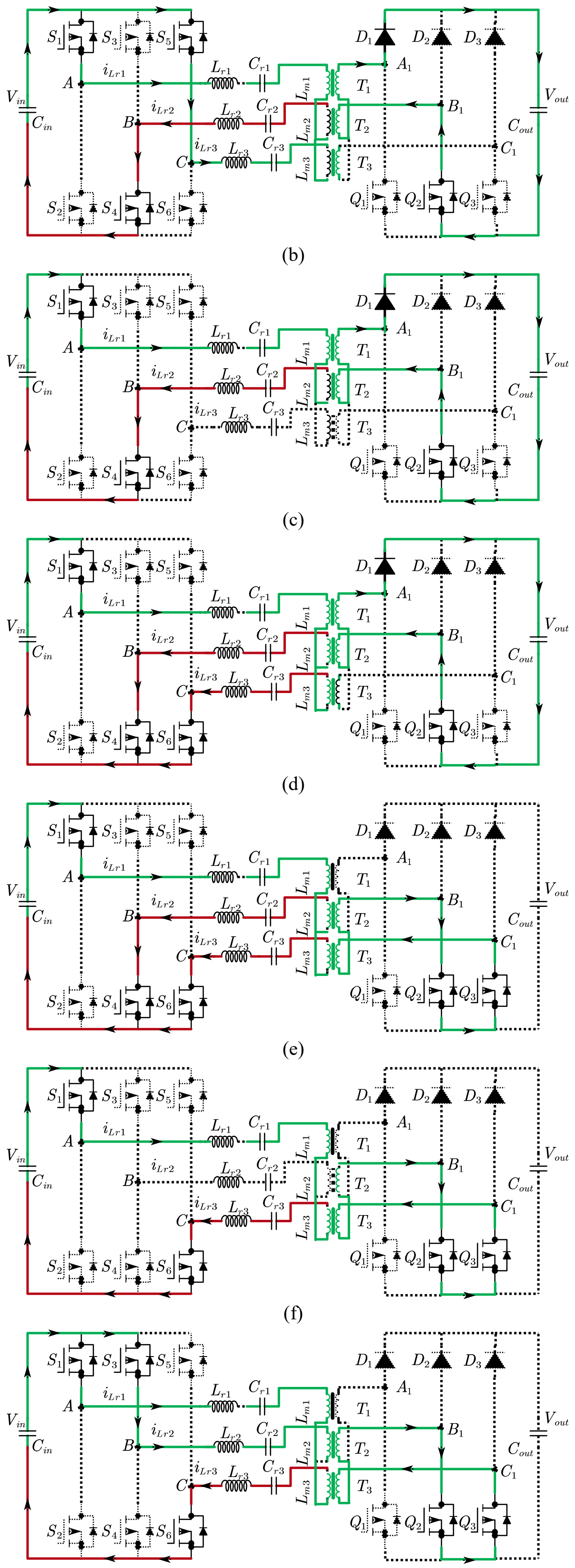


Fig.2 Boost mode waveform under secondary-side semi-active control.

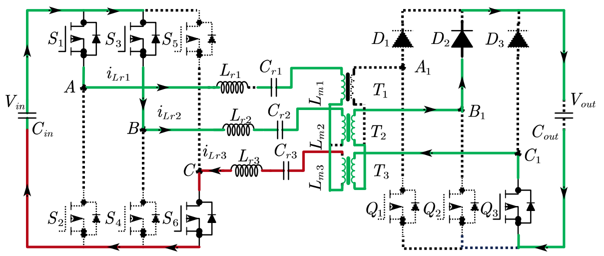
discharge together to the load to obtain a boost effect. The same is true for S_3, S_6, S_1conduction and S_5, S_2, S_3 conduction. We name the time when S_1, S_4, S_5 and Q_1 pulses overlap as the secondary side duty cycle D_s. The adjustment of the D_s can be achieved by adjusting the hysteresis phase of $Q_{1\text{-}3}$ with respect to $S_{2\text{-}6}$. The larger of the D_s, the larger the resonant network energy storage and the higher the boost ratio. In this mode, the bus voltage is fixed at 850V, when the output voltage is 850-1000V, the converter relies on adjusting the D_s to achieve the voltage gain change. For easy understanding, we divided the converter into 8 modes in half a cycle and drew the equivalent circuit diagram, as shown in Fig. 3.



(a)



(b) (c) (d) (e) (f) (g)



(h)

Fig.3 Equivalent circuit of (a)-(h),Mode a-h

Mode a (t_0-t_1): At the moment of t_0, S_1, S_2 commutation ends, A-phase bridge arm resonant current i_{Lr1} is negative, S_1 turn-on signal V_{gss1} arrives, S_1 realizes ZVS-ON. At this time, transformer primary switching tubes S_1, S_4, S_5 conduct, secondary Q_1 conducts, secondary current flows positively through Q_1, flows through Q_2 parasitic diode to short circuit transformer T_1, T_2 secondary windings. V_{in} charges the resonant network, and the resonant currents i_{Lr1}, i_{Lr2} and i_{Lr3} increase linearly. At this time, the resonant network structure is A, C phase in parallel with B phase in series. According to the circuit KVL theorem, the resonant current expression can be obtained as:

$$i_{Lr1}(t) = i_{Lr1}(t_0)\cos[\omega_r(t-t_0)] + \frac{V_{in}-u_{Cr}(t_0)}{Z_r}\sin[\omega_r(t-t_0)] \quad (1)$$

where the resonant angular frequency is ω_r=1/(L_r/C_r)^2, and the characteristic impedance is Z_r=(L_r/C_r)^0.5.

Mode b (t_1-t_2): Q_1 is turned off at t_1, and the secondary windings of transformers T_1 and T_2 are no longer short-circuited, and the secondary current flows through D_1 and Q_2 parasitic diodes for rectification. The primary switching tube action and resonant network structure is the same as the previous mode, at this time, V_{in} and resonant network together discharge to the load. This mode within the resonant network occurs in series resonance, resonant frequency f_r is (2). Where + represents the series connection, the inductor series inductance value is added, the capacitor series capacitance value is the inverse of the capacitance value is added. Similarly, the expression for the resonant current is (3).

$$f_r = \frac{1}{2\pi\sqrt{(L_{r1}//L_{r3}+L_{r2})\times(C_{r1}//C_{r3}+C_{r2})}} = \frac{1}{2\pi\sqrt{L_r\times C_r}} \quad (2)$$

Mode c (t_2-t_3): At the moment of t_2, the switching tubes S_5, S_6 commutate the current. During the dead time, the C-phase bridge arm resonant current i_{Lr3} is positive, i_{Lr3} charges and discharges the S_5, S_6 parasitic capacitors, and i_{Lr3} flows through the S_6 parasitic diode to continue the current. The secondary side of the transformer is the same as the previous mode.

$$i_{Lr1}(t) = i_{Lr1}(t_1)\cos[\omega_r(t-t_1)] + \frac{V_{in}-u_{Cr}(t_1)-nV_o}{Z_r}\sin[\omega_r(t-t_1)] \quad (3)$$

Mode d (t_3-t_4): At the moment of t_3, the S_6 turn-on signal arrives and S_6 achieves ZVS-ON. At this time, the transformer primary switching tubes S_1, S_4 and S_6 are turned on. The resonant network structure is B-phase in parallel with C-phase and then in series with A-phase. The resonant frequency is f_r. The secondary side of the transformer is the same as the previous mode.

Mode e (t_4-t_5): At moment t_4, the secondary i_{Q2} is commutated past the zero point. The secondary current flows positively through Q_2, and flows through Q_3 parasitic diode to short-circuit the transformer T_2, T_3 secondary winding. At this time, V_{in} charges the resonant network and i_{Lr1}, i_{Lr2} and i_{Lr3} increase rapidly. The resonant network structure is the same as the previous mode.

Mode f (t_5-t_6): At the moment of t_5, the switching tubes S_3, S_4 change the current. During the dead time, i_{Lr2} is negative and i_{Lr2} charges and discharges the S_3, S_4 parasitic capacitors. After charging and discharging, the two ends of S_4 are charged to V_{in} and i_{Lr2} flows through the S_3 parasitic diode to renew the current.

Mode g (t_6-t_7): At the moment of $t6$, the S_3 turn-on signal arrives and S_3 achieves ZVS-ON. At this time, the primary switching tubes S_1, S_3, S_6 are turned on and the secondary state is consistent with mode e. The primary resonant network is AB-phase parallel and C-phase series. The primary resonant network is connected in parallel in phase AB and in series in phase C. V_{in} charges the resonant network and i_{Lr1}, i_{Lr2} and i_{Lr3} increase rapidly.

Mode h (t_7-t_8): At the moment of t_7, Q_2 turns off, transformer T_2, T_3 is no longer short-circuited, and the secondary current flows through D_2 and Q_3 parasitic diodes for rectification. The resonant network is consistent with the previous mode, at this time, V_{in} and resonant network together discharge to the load.

B. Working Principle of Variable Bus Voltage Control and Buck Mode

The operating waveform of the SS-SA3PI resonant converter in variable bus voltage mode is shown in Fig. 4. The primary side switches S_1 - S_6 operate in the same mode as Boost mode, and the secondary side Q_1 - Q_3 switch signals are always low. S_1 - S_6 realize ZVS-ON, D_1-D_3 and Q_1 - Q_3 realize ZCS-OFF. In this mode, the input voltage is floating, $f_s = f_r$, the converter voltage gain is constant at 1, and the output voltage (660-850V) is adjusted by adjusting the input voltage V_{in} (660-850V). For example, when the desired output voltage is 750V, we only need to adjust the input voltage to 750V to achieve an output voltage of 750V.

In Buck mode, the bus voltage is fixed at 660V. The converter primary side switching tubes are complementary, each phase bridge arm is lagged 120° in turn, and the secondary side drive signal is often low. When the output voltage is 100-660V, $f_s > f_r$, the switching frequency is adjusted to reduce the voltage gain of the converter. In this mode, the primary-side switches S_1-S_6 achieve ZVS-ON and the secondary-side switches do not achieve ZCS, but considering that the secondary-side switch tubes and diodes

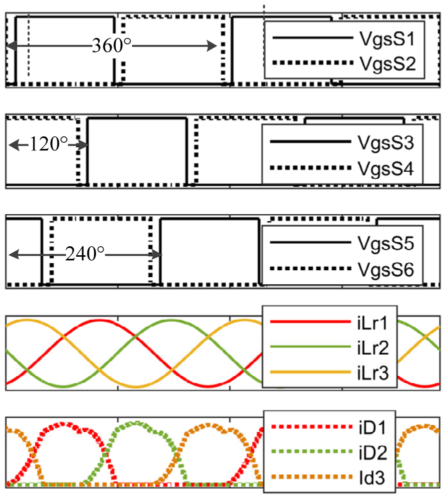


Fig.4 Variable bus voltage mode: f_s=f_r.

are of SiC structure, the secondary-side reverse recovery losses can be ignored. The operating mode in variable bus voltage and buck mode can be referred as the operating mode in boost mode, except that the secondary-side semi-active control is correspondingly absent.

III. Characterization And Control

A. Conditions For ZVS Implementation

Under multi-mode modulation, the implementation of the primary side switching ZVS needs to rely on the resonant current to fully charge and discharge the switching tube junction capacitance during the dead time, which needs to meet:

$$L_{m\max} < \frac{t_{dead}}{16C_{oss}f_s}. \tag{4}$$

In (4), t_{dead} is the dead time and C_{oss} is the ZVS equivalent charge/discharge capacitance. When the converter works in boost mode, the implementation of the transformer's secondary switching tubes Q_1 - Q_3 ZVS is the simplest, because the rectifier current does not flow through the switching tubes before the secondary switching tubes turn on, and the switching tube DS voltage is always 0.

B. Voltage Gain

1): Buck and variable bus voltage control mode: In these mode, the DC gain of the SS-SA3PI resonant converter is equation (5)[10]. In (5), n_p is the number of turns of the primary winding of the transformer, n_s is the number of turns of the secondary winding of the transformer. k is the inductance ratio, f_n is the normalized switching frequency, and Q is the quality factor of the converter. R_{ac} is the equivalent AC equivalent load to the primary side of the transformer. R is the output load equivalent resistance value.

Considering the limit working condition, the minimum output voltage at constant power in this mode is 400V, the minimum output voltage at non-constant power is 100V, and the current limit is 25A. From (5), Fig. 5 can be made. From

Fig. 5, at $f_n = 1.4$, the full load $V_o = 400$V; at $f_n = 1.63$, the maximum output current $I_{omax} = 25$A, $V_o = 100$V; at $f_n = 2.88$, the light load, $V_o = 100$V, the voltage gain meets the requirement.

$$M = \frac{V_{out}}{V_{in}} = \frac{1}{n}\frac{1}{\sqrt{(1+k-\frac{k}{f_n^2})^2 + Q^2(f_n - \frac{1}{f_n})^2}} \tag{5}$$

$$k = \frac{L_r}{L_m}, Q = \frac{\sqrt{L_r / C_r}}{n^2 R_{ac}}, n = \frac{n_p}{n_s}, f_n = \frac{f_s}{f_r},\quad R_{ac} = \frac{6}{\pi^2} R n^2$$

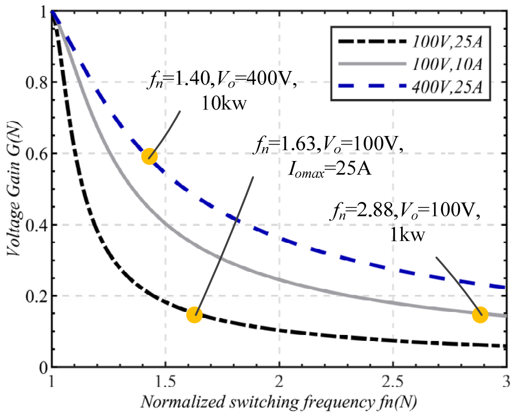


Fig.5 Buck mode: voltage gain curve.

2): Boost mode: In this mode, the switching frequency $f_s = f_r$, then the converter AC gain is 1. According to the above modal analysis and waveform, it is known that the excitation current i_{Lm1} rises to i_{Lr1} at the moment t_4 and the secondary side current i_{Lr11} is intermittent as shown in Fig. 6. From the inductance voltage $V_{in} = L_r \times d_{iLr11}/d_t$ we know that:

$$\Delta I_L = \frac{V_{in} D_s}{L_r f_s} \tag{6}$$

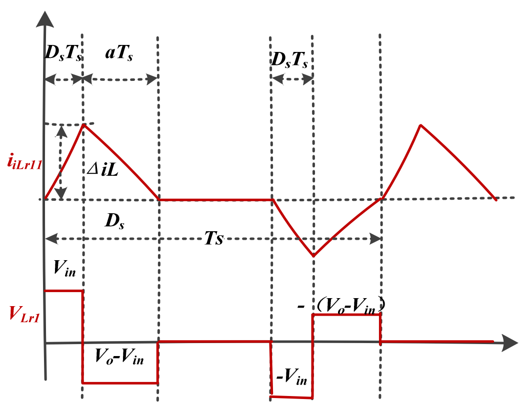


Fig.6 i_{Lr11} and V_{Lr1} curve.

According to the volt-sec balance relation of the inductor voltage under steady-state conditions, it is obtained that:

$$V_{in} D_s T_s = (V_o - V_{in}) a T_s \tag{7}$$

The diode current switching cycle average and load current are:

$$I_D = \frac{1}{2}\Delta I_L a, I_o = \frac{V_o}{R} \tag{8}$$

In steady state, the average value of diode current switching cycles is equal to the load current. Based on this, for i_{Lr1}, which is short-circuited twice per cycle, the output voltage gain M_{boost} can be expressed as :

$$M_{boost}(f_s, L_r, D_s, R) = \frac{V_{out}}{V_{in}} = \frac{1}{2} + \sqrt{\frac{1}{4} + \frac{2D_s^2 R}{L_r f_s}} \tag{9}$$

Equation (9) can be further expressed as:

$$V_o = f(D_s) = \frac{V_{in} P L_r f_s}{P L_r f_s - 2D_s^2 V_{in}^2} \tag{10}$$

When operating in Boost mode, the converter has a constant maximum power output, then V_{in}, P, L_r, f_s are constants in (10), and (9) is reduced to a univariate function $V_o = f(D_s)$, which is easy to implement in the online controller. Equation (9) and (10) are verified in Matlab/Simulink to obtain Fig. 7, which shows that the theoretical calculation matches the simulated gain. Since the input voltage V_{in} = 850V and the maximum output voltage V_{omax} = 1000V in Boost mode, the circuit requires the maximum gain $M_{boostmax}$ = 1.1764. Considering the dead time of the converter, the gain loss caused by the rise/fall and turn-on/off delay time of the switching tubes, the designed boost gain range is far more than 0.15 at D_{smax} = 0.15. The designed boost gain range meets the requirements.

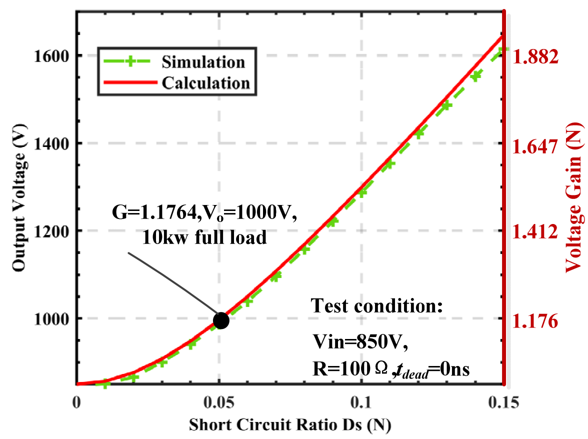


Fig.7 Boost mode: voltage gain curve.

C. Multi-Mode Modulation

From the above analysis, we can see that the multi-mode modulation strategy uses three different control methods, and we name the three different control methods used as three modes. The selection of the mode is based on the output voltage as can be seen in Fig. 8. Also, Fig. 8 shows an example of the output voltage varying from 100V to 1000V, with the transition and control variables changing for each mode:

At the output voltage range of 100-660V, the input voltage V_{in} is fixed at 660V, D_s=0, the switching frequency f_s is gradually reduced to improve the voltage gain, until f_s reaches f_{smin}, f_s=f_r , V_{out}=660V. At this time the voltage gain is 1,

V_{out}=V_{in}. When the output voltage needs to continue to increase, the converter enters the next mode.

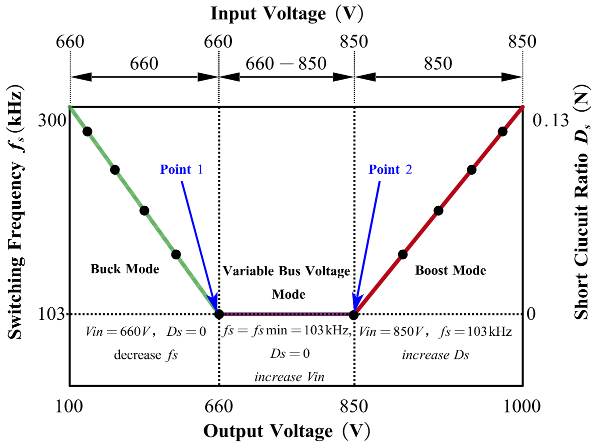


Fig. 8 Example of output voltage regulation.

At the output voltage range of 660-850V, f_s=f_r, D_s=0, the voltage gain is fixed to 1. When the output voltage needs to continue to increase from 660V, the converter only needs to gradually increase the input voltage V_{in} to achieve the output voltage regulation. Until the output voltage reaches 850V, the switching frequency and input voltage regulation will no longer increase the gain of the converter, then the converter enters the next mode.

At the output voltage range of 850-1000V, f_s=f_r, V_{in}=850V. When the output voltage needs to continue to increase from 850V, the converter only needs to gradually increase D_s to achieve the output voltage regulation. When D_s increases to 0.13, the converter reaches its maximum voltage gain point and achieves an output of 1000V.

From this example we can see that the converter goes through three modes when the output voltage changes from 100V to 1000V. Among them, the converter mainly involves two mode transition points: Point 1 and Point 2. They correspond to output voltages of 660V and 850V, respectively. And according to the above voltage gain analysis, the voltage gain of the two transition points Point 1 and Point 2 is 1. In addition, the converter has a constant 10kW output when the output voltage is 400V or more. Therefore, the mode transition of the converter does not involve a change in voltage gain or power. Thus, the problem of dynamic transition stability is eliminated.

The control block diagram of this paper is given in Fig. 9. When the desired output voltage is lower than 660V, the MCU of the converter fixes the bus voltage of the PFC to 660V through SCI communication, and then controls the D_s to 0 to reduce the voltage gain. Finally, the switching frequency is adjusted to achieve the desired output voltage. When the desired output voltage range is between 660-850V, the PI regulator output f_s is saturated. At this point, adjust V_{in} to obtain the desired output voltage. When the desired output voltage is higher than 850V, both the controlled quantity input voltage V_{in} and the switching frequency f_s are saturated. At this time, increase D_s to increase the voltage gain of the converter to achieve the desired output voltage.

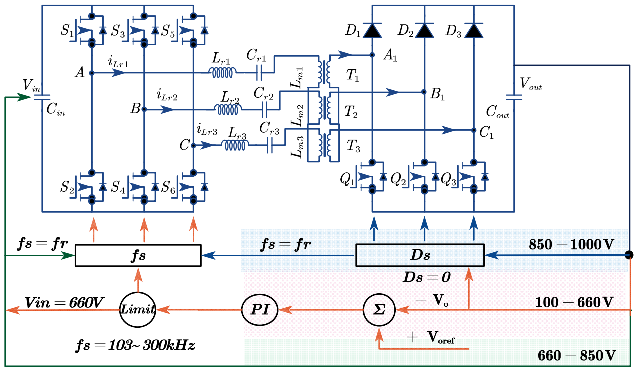


Fig. 9 Control block diagram.

IV. Design Considerations

A. Selection of the Frequency

Core losses account for a large component of the losses in high power isolation converters. We choose Mn-Zn ferrite made of PC95 as the core of the magnetic element. This material has a large reduction in core loss in the range of 25° to 100° compared to PC40 material. Combining the loss of the magnetic core datasheet and considering the converter heat dissipation mode, power density, control mode and stability, the resonant frequency and the maximum switching frequency in the constant power range are designated as 100kHz and 140kHz. Also, in the non-constant power voltage range (V_{outmin}=100V), the maximum switching frequency is designed to be 3 times the resonant frequency, i.e., 300 kHz, to obtain a large voltage step-down effect.

B. Resonant Parameter Design

According to equation (4), we only need to determine t_{dead} and C_{oss} to obtain the L_{mmax} value. Considering the turn-on/ off delay time of the SiC switch and the safety margin, the t_{dead} was determined to be 200ns. In order to reduce the dv/dt of the switching tubes and the midpoint of the bridge arm of each phase, so as to achieve the reduction of switching turn-off losses and improve EMI, we generally in parallel with an additional capacitor C_{DS}=1nF between the switching tubes DS. And C_{oss} in equation (4) should be C_{DS} + C_{ossmos}. Considering the nonlinearity of the switching tube C_{ossmos} and the ZVS margin we design the L_m value as 50% of L_{mmax}, so L_m=50uH.

As mentioned before, based on loss considerations, we set the maximum switching frequency of the converter at 140kHz for the constant power case, which is 660V input and 400V 10kW output. According to equation (2) and (5), we decouple the formula of Q. The decoupled Q can be expressed as:

$$Q=\frac{2\pi f_r L_r}{n^2 R_{ac}} \tag{11}$$

And the parameter solutions of L_r can be calculated by equation (11) and (5) when f_n, f_r, L_m and R_{ac} are known. And C_r can be calculated according to equation (2).

C. Resonant Component Design

The design of magnetic components is concerned with only two things: whether the core will saturate (considering losses at high frequencies) and whether the windings will overheat. And the core evaluation equation for transformers and inductors can be expressed as:

$$B_{Lr\max}=\frac{i_{Lr\max}L_r}{A_e N_{Lr}},B_{T\max}=\frac{E_{T\max}}{4f_s A_e N_p} \tag{12}$$

where i_{Lrmax} can be calculated according to equation (1), and the maximum value of the average voltage on the transformer is:

$$E_{T\max}=\frac{2}{T}(\int_0^{t_2}V_{AN}dt+\int_{t_2}^{t_4}V_{AN}{}^2dt+\int_{t_4}^{t_8}V_{AN}{}^2dt)=\frac{4}{9}V_{in\max} \tag{13}$$

D. Discussion of Voltage and Current Stress

Once the resonance parameters are determined, we can discuss the voltage and current stresses on the switching tubes of the converter. In the soft switching converter, the voltage stress of the primary side switch tube is V_{inmax}=850V, and the voltage stress of the secondary side switch tube is V_{outmax}=1000V. Considering the resonance caused by the distribution parameters during the turn-on and turn-off of the converter, the withstand voltage of the switching tubes should have a margin of 20%.

According to the operating principle of the converter, we know that when the input voltage is 660V and the output voltage is 400V, the current stress of the converter is maximum. The RMS value of the current on the primary side of the converter can be expressed as:

$$i_{pri\_rms}=\frac{\pi}{3\sqrt{2}}\times\frac{P_{out\max}N_s}{V_{out}\eta N_p} \tag{14}$$

Where η is the efficiency of the converter, we take 0.95 in the calculation of stress. The excitation current of the transformer can be expressed as:

$$i_{Lm\_rms}=\frac{E_{Trms}}{2\pi f_r L_m} \tag{15}$$

Where the transformer voltage RMS value E_{Trms} is:

$$E_{Trms}=\sqrt{(\frac{V_{in}}{3})^2\times\frac{2}{3}+(\frac{2V_{in}}{3})^2\times\frac{1}{3}} \tag{16}$$

According to equation (13), (14) and (15), we can know that the primary side current stress is also the peak resonant current is:

$$i_{Lr\_\max}=\sqrt{2}\times\sqrt{\left(i_{Lm\_rms}\right)^2+\left(i_{pri\_\max}\right)^2} \tag{17}$$

The current stress on the secondary side is the same as the primary side because the ratio of the transformer is 1.

E. Parameter Design Process

The parameter design process is shown in Fig. 10. The hardware parameters can be initially determined according to the parametric design flow, and then the design flow can be gradually adjusted according to the loss analysis and hardware layout.

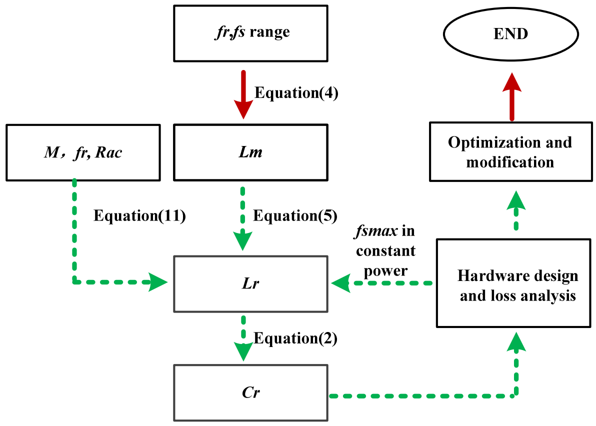


Fig.10 Parameter design process.

V. Experimental Verification And Discussion

To evaluate the performance of the SS-SA3PI resonant converter, an experimental prototype was built. The pictures and parameters of the prototype are shown in Fig. 11 and Table I. The experimental prototype has a size of 250*210*36.6 mm and achieves a power density of 5.2kW/L. The cooling of the experimental prototype is forced air cooling, provided by two fans. The green arrow in Fig. 11 represents the fan wind direction, and the prototype leaves a slight gap in each component to improve heat dissipation.

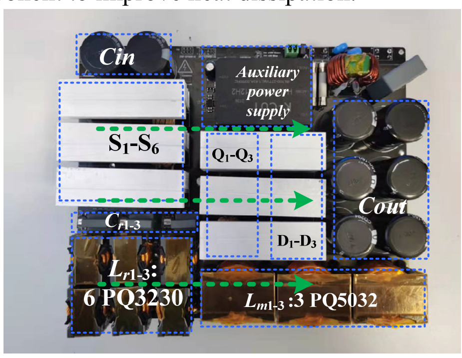


Fig.11 Experimental prototype

TABLE I
THE DETAILED PARAMETERS OF THE HARDWARE

Symbol	Detail	Symbol	Detail
f_r	103kHz	S_1-S_6	UF3C120040K4S
f_s	103-300kHz	Q_1-Q_3	UF3C120040K4S
L_r	24uH	D_1-D_3	FFSH40120ADN
C_r	99nF	V_{in}	660-850V
L_m	50uH	V_{out}	100-1000V
n_p:n_s	10:10	I_{omax}	25A
C_{out}	270uF	P_{omax}	10kW
L*W*H:250*210*36.6mm		Power dentist：5.2kW/L	
Constant maximum power at output voltage 400-1000V			

A. Experimental Performance

Fig. 12 shows the waveform for 400V 25A in buck mode with an input voltage of 660V. From Fig. 12, we can see that i_{Lr1}>0 when S_2 is turned on, the resonant current flows through the S_2 parasitic diode let S_2 achieves ZVS-ON, and similarly, S_1-S_6 are also ZVS-ON. Besides, when S_1 turn-off, V_{dsS1} and V_{gsS1} crossover is small, so the subsequent loss analysis ignores the turn-off loss.

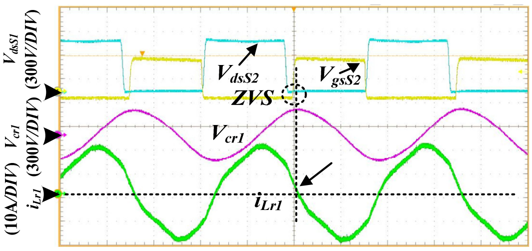
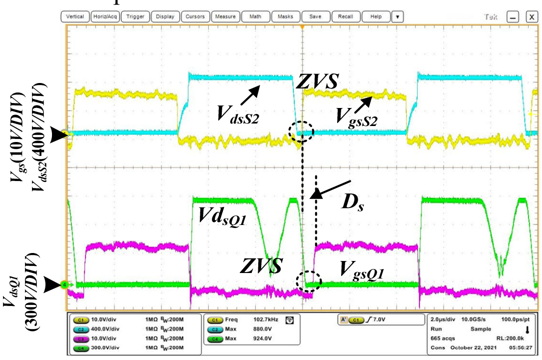
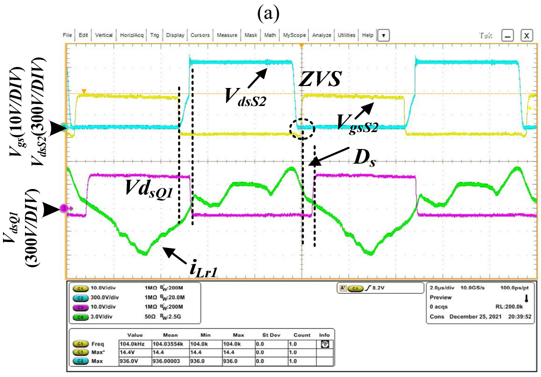


Fig.12 ZVS realization waveform: V_{in}=660V, V_{out}=400V I_{out}=25A, f_s=143kHz at 10kW.

Fig. 13 shows the waveform of the SS-SA3PI resonant converter in boost mode under secondary-side semi-active control in heavy load with $f_s = f_r$. From Fig. 13(a), it can be seen that the specific implementation of the semi-active boost control on the secondary side is achieved by adjusting the hysteresis phase shift angle of $Q_{1,\ 2,\ 3}$ with respect to $S_{2,\ 4,\ 6}$. Under this control, G > 1, the converter achieves boost operation. At the same time, ZVS-ON is achieved for both the primary and secondary side switching tubes. Fig. 13(b) shows the waveforms of V_{dsS2}, V_{gsS2}, V_{gsQ1} and i_{Lr1} for a full-load input voltage of 850V. Based on this, we can analyze that the resonant current i_{Lr1} is short-circuited in the secondary side winding of transformer T_1T_2 after S_2 is turned off and S_1 is turned on, and i_{Lr1} rises rapidly and linearly, proving the charging process of the input voltage to the resonant network, thus, the boost operation of the converter is realized.

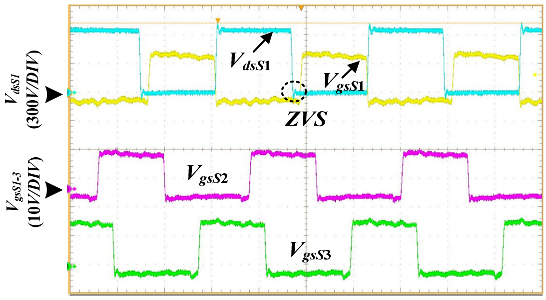


(a)

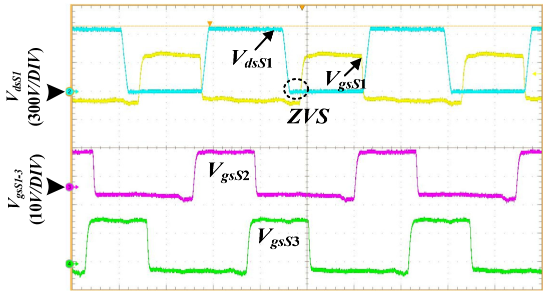


(b)

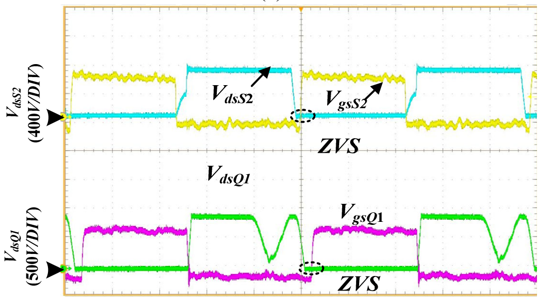
Fig.13 Boost mode: secondary-side semi-active control. (a) f_s=f_r, V_{in} = 800V, V_{out}=924V, 8.5kW, D_s=0.07, G=1.155. (b) f_s=f_r, V_{in} = 847V, V_{out}=977V, 9.6kW, D_s=0.07, G=1.155.



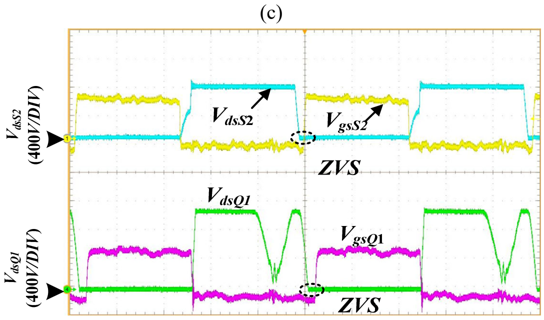
(a)



(b)



(c)



(d)

Fig.14 ZVS implementation waveform. (a) f_s=155kHz, V_{in} = 660V, V_{out}=100V, 2.5kW. (b) f_s=290kHz, V_{in} = 660V, V_{out}=100V, 1kW. (c) f_s=f_r, V_{in} = 850V, V_{out}=900V, 1kW. (c) f_s=f_r, V_{in} = 850V, V_{out}=1000V, 1kW.

Fig. 14 shows the ZVS implementation waveform for the converter at different output voltages, especially at light loads. Fig. 14(a) shows the waveform diagram for a switching frequency of 155kHz with an input voltage of 660V and an output of 100V 25A. We can see that S_1 achieves ZVS-ON, and S_1, S_2, and S_3 drive signals are each lagged by 120°, it can be inferred that $S_{1\text{-}6}$ all achieve ZVS-ON, the converter works well. The most difficult situation for the converter to achieve

ZVS is when the converter output voltage and output power are at their lowest, which is shown in Fig. 14(b). The switching frequency is 290kHz with an input voltage of 660V and an output of 100V 10A. We can see that the converter realizes ZVS. Also, Fig. 14(c) and (d) show the ZVS cases when the converter output conditions are 900V 1kW and 1000V 1kW. From all this, we can infer that the converter realizes ZVS soft switching under all working conditions.

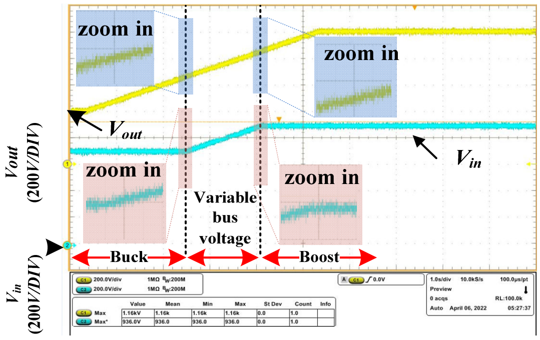
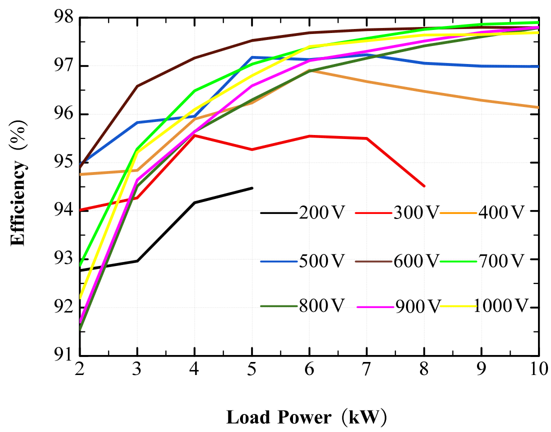


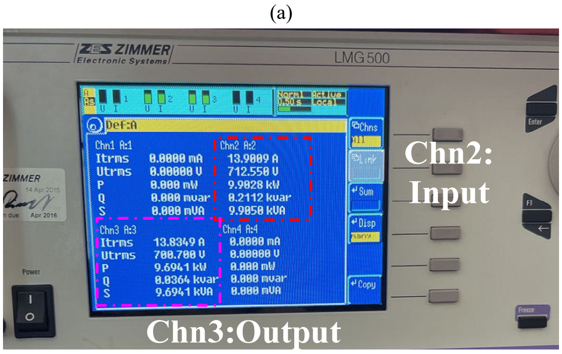
Fig.15 Mode transition: output voltage is from 411V to 1003V.

The mode transition waveforms are shown in Fig. 15. From the figure, we can see that the converter experiences three modes when the output voltage changes from 411V to 1003V. Moreover, there is no obvious overshoot of the output voltage at the 2 mode transition points. The converter employs the multi-mode modulation scheme achieves smooth mode conversion.

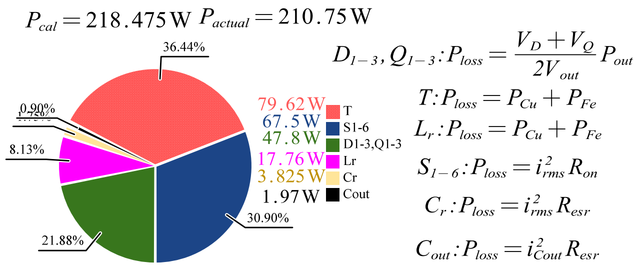
B. Measured Efficiency



(a)



(b)



(c)

Fig.16 Efficiency of SS-SA3PI resonant converter.

Fig. 16(a) shows the efficiency curve of the converter for 100-1000V. The maximum efficiency of the converter is 97.8925%, which is measured by the power analyzer LMG500, as shown in Fig. 16(b). And the loss distribution and calculation at this measurement efficiency is shown in Fig. 16(c), where the iron loss of the magnetic element is calculated by numerical fitting method. The rectification losses of diodes and switching tubes are mainly considered as on-state voltage drop losses (V_D and V_Q are the average values at the highest and lowest temperatures), and S_1-S_6 are mainly considered as conduction losses (R_{on} is the average value at the highest and lowest temperatures). Since the parameters of the switching tubes and diodes vary greatly with temperature and device consistency, there is a 7.7W difference between the loss calculation and the actual loss. Through the loss distribution chart we can see that the loss of the transformer at the highest efficiency is the main loss, which is because at this time the transformer B_{max} is 0.23 and the iron loss is higher. Also, this explains why the converter has the highest efficiency at 700V output, because when the output voltage is higher, the transformer has a higher iron loss.

C. Comprehensive Comparison with Similar Structures

We have made a comprehensive comparison of some similar structures. Among them, the structure of reference [33] is a three-level interleaved LLC. The common denominator of the compared structures is the presence of three symmetrical sets of resonant elements. However, the sizes and power density of the experimental prototypes are not described in detail. Some detailed comparison information is given in Table II and Fig. 17. In this case, the maximum voltage conversion ratio in Fig. 17(b) ignores the transformer ratio and compares only the voltage conversion ratio provided by the control method and the converter structure. It can be seen that the SS-SA3PI resonant converter using multi-mode modulation proposed in this paper has obvious advantages.

For a fuller comparison, we have compared our experimental prototypes with industrial products in detail. The comparison results are shown in Table III. It should be noted that the AC/DC and DC/DC converters are included in the off-board EV charger product. The overall structure of the product is that the AC/DC and DC/DC converter are mounted in a top and bottom pair. The length and width of the two converters are the same(420*213mm), but because the number of components in the AD/DC converter is small, it allows the

TABLE II
COMPARISON WITH SIMILAR STRUCTURES

	This paper	[3]	[4]	[21]	[33]
Structure	SS-SA3PI	3-phase LLC	3-phase interleaved LLC	3-phase interleaved LLC	3-level 3-phase interleaved SRC
Rated power	10 kW	5 kW	10 kW	3 kW	2 kW
Input voltage	660-850V	600V	360-400V	400V	440-590V
Output voltage	100-1000V	600V	300V	32-100V	54-400V
Max efficiency	97.89%	97.6%	97.88%	96%	96.5%
Additional components	No	No	No	Yes	No
Switching tubes	9Mosfet+3Diode	6Mosfet+6Diode	12Mosfet+6Diode	6Mosfet+8Diode	6Mosfet+8Diode
Power density	High	Low	Medium	Not given	Not given

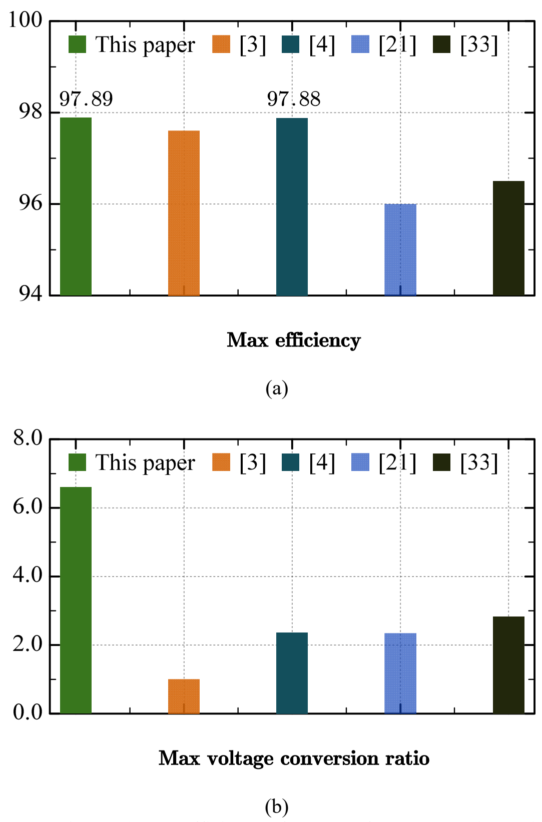


Fig.17 Comparison. (a) max efficiency. (b) max voltage conversion ratio.

components of the DC/DC converter to protrude(transformer). The volume of this part of the protruding element was also added to Table III to calculate the detailed power density. From the comparison results in Table III, we can see that the output voltage range of this paper is wider for the same input voltage and no additional components are added. However, one of the drawbacks of this paper is that it has the same power density as the industrial product, but with slightly lower efficiency. This is mainly due to the following reasons:

1) The height of the transformer has been reduced for the sake of uniformity in the height of all components. This makes the transformer window area smaller and the number of turns on the primary side smaller, resulting in a larger transformer B-value and reduced conversion efficiency.
2) The experimental prototype in this paper contains the auxiliary power supply, while the auxiliary power supply for DC/DC of industrial products is located in the AC/DC converter.
3) The number of custom components also creates a disadvantage.

TABLE III
COMPARISON WITH INDUSTRIAL PRODUCT

	This paper	[8]
Structure	SS-SA3PI	3-level, 3-phase interleaved LLC with ISOP/S
Rated power	10kW	20kW
Input voltage	660-850V	660-850V
Output voltage	100-1000V	250-1000V
Max efficiency	97.89%	98.1%
Size L*W*H	250*210*36.6mm	420*213*40mm + 108*73*33mm
Power density	5.2kW/L	5.2kW/L
Additional components	No	Yes
Switching tube number	9Mosfet+3Diode	12Mosfet+12Diode

VI. Conclusion

The proposed SS-SA3PI resonant converter employing a multi-mode modulation scheme is well suited for high-power

EV charger applications. This modulation scheme can greatly widen the output voltage regulation range of the converter without adding additional circuit components. It also avoids the problem of dynamic stability during mode transition. The use of secondary-side semi-active control in boost mode also reduces the design difficulty of resonance parameters and magnetic components. This all help to reduce the hardware cost and improve the power density of the converter. Moreover, the converter realizes soft switching under all working conditions and obtains high conversion efficiency. The experimental results show that the converter achieves an output voltage of 100-1000V with a maximum conversion efficiency of 97.89% at a power level of 10kW.

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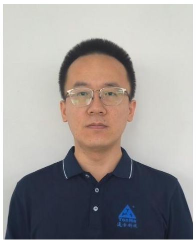

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