Analysis and Design of a Single-Stage Bridgeless Isolated AC-DC Resonant Converter for Programmable AC Power Source Applications

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ABSTRACT This paper proposes a single-stage bridgeless isolated AC-DC power factor correction (PFC) topology, and this topology can realize the function of step-up and the galvanic isolation for the AC input and DC output voltages through single-stage power conversion process. Firstly, the operation principle of the topology is analyzed in details. Then, by reasonably designing the series resonant parameters, the zero-voltage-switching (ZVS) turn-on for the power switches in the primary-side and zero-current-switching (ZCS) turn-off for the output rectifier diodes in the secondary-side of the transformer can also be achieved under wide load range, and the voltage stress of the power switches are also reduced based on the active clamping technology. The design procedure for the key components of the proposed converter is also derived. Finally, a 3 kW experimental prototype based on SiC power devices with 85 kHz switching frequency is built to verify the correctness and feasibility of the proposed topology, and the experimental results show a good performance. It can operate in a wide input grid voltage range, 95.7% of peak conversion efficiency and 3.9% of total harmonic distortion (THD) value for grid current could be obtained.

INDEX TERMS AC-DC converter, power factor correction, series resonant, zero-voltage-switching, zero-current-switching.

I. INTRODUCTION

In order to reduce the harmonic pollution to the power grid and meet the harmonics regulations and standards of the international electrotechnical commission (IEC) and institute of electrical and electronics engineers (IEEE), active power factor correction (PFC) technique is widely adopted for many AC-DC converters, and these AC-DC equipments are widely used for uninterruptible power supply, communication power supply, electric vehicle on-board charger, programmable AC power source and others [1]–[3].

Isolated AC-DC PFC converters are becoming necessary for many industrial applications [4], [5]. The conventional isolated PFC topology has two stages, the first stage is implemented for rectifier and PFC functions, and the rear stage is designed for output voltage regulation and galvanic isolation [5]–[7]. Many combined topologies are proposed for different industrial applications in pursuit of high conversion efficiency and power density [8]–[10]. However, due to the separated topology structure of the front and rear stages, they have the disadvantages of high components count and complex control strategies, which would lead to high cost, large power losses and low conversion efficiency [5].

Single-stage isolated AC-DC PFC converters process the power conversion only in one stage, and they have the features of simple control structure, few components and high efficiency, their topological structure and control strategy become the research hotspot recently [11]–[16]. A single-stage single-switch isolated converter which combines the boost-like topology and forward topology is proposed in [11] for low voltage and low power application, it is suitable for wide range of line voltage applications with high factor...
values, but it remains the low conversion efficiency and high grid current harmonic content. An improved single-stage boost-flyback with a synchronous rectifier is proposed in [12] for higher power factor and efficiency. A quasi-active PFC scheme is also presented in [13], the input inductor can operate in the discontinuous current mode (DCM), and the performance of lower harmonic content and high power factor could be achieved based on adding an auxiliary winding to the transformer of the cascade flyback stage. In addition, many new single-stage isolated topologies with soft-switching technique have been proposed for light-emitting diodes (LED) driver applications [14]–[16]. However, these single-stage isolated PFC topologies mentioned above are only suitable for low power applications [4], [17]–[20].

For high-power AC-DC application, a 1.6 kW single-phase isolated Cuk PFC converter with digital double-loop control method is proposed in [21], and nearly unity power factor (PF), low THD of grid current and fast dynamic performance under load variation can be achieved, but only 91% of the conversion efficiency can be obtained due to the existence of input diode rectifier bridge and the hard switching operation state of the power devices. A novel bidirectional single-stage 1.5 kVA AC-DC converter for single-phase EV charging application is introduced in [22], it consists of a current-fed half-bridge converter on the primary side and a full-bridge converter on the secondary side of a high-frequency transformer, newly modulation method and control strategy are also presented to guarantee the soft switching condition for power switches in both direction of power flow, and almost 2.5% THD, up to 96.5% conversion efficiency can be achieved for full-load conditions. In [23], a 3 kW single-stage bridge-type AC-DC isolation converter with the current-fed full-bridge topology is proposed for high power PFC applications, zero-current-switching (ZCS) operation is realized for all power switches, but the THD value is over 5% at the rated power, and the input diode bridge still remains, the measured peak efficiency is around 93% to 94%. A combined phase-shift and frequency modulation for the 3.3 kW dual-active-bridge (DAB) AC-DC converter with PFC is presented in [24], it could realize the ZVS over the whole AC mains voltage period, the high power density, a wide output voltage range and a measured efficiency of 95% for 350V output voltage can be obtained, but the drawback of complex control strategy and utilization of two many power devices still remains. A 1.4 kW single-stage AC-DC converter based on dual active bridge is proposed in [25], ZVS operation for the input-side power switches and ZCS operation for the output-side power switches can be achieved under certain load conditions, the maximum conversion efficiency can reach 96%. A 3.3kW bidirectional single-stage AC-DC converter by integrating the interleaved totem-pole PFC converter with the DAB is proposed in [26], the measured efficiency for both forward and reverse modes is almost identical with peak efficiencies of 96.7% and 96.2% respectively. But these topologies mentioned above use too many power switches, which will lead to high hardware costs.

In this paper, a novel single-stage AC-DC bridgeless isolated PFC for programmable AC power source application is presented as shown in Fig.1. ZVS of power switches on the primary side and ZCS of rectifier diodes on the secondary side of the high-frequency (HF) transformer can be achieved based on the resonant soft switching technique. It can operate with high input PF, wide range of input voltage and low THD value of grid current. Moreover, as the reduction of the input diode rectifier bridge and the soft-switching operation for power devices, the remarkable conversion efficiency of 95.7% can be achieved.

This paper is also organized as follows. The topology and operation principle of the proposed single-stage isolated AC-DC resonant converter are illustrated in section II. The details of the steady-state analysis and design consideration are also introduced in section III and IV respectively. In Section V, experimental results are shown to demonstrate the feasibility of the proposed single-stage topology based on a 3 kW prototype. Finally, the conclusion for this paper is contained in Section VI.

**FIGURE 1. Topology of the programmable AC power supply.**

**II. PROPOSED SINGLE-STAGE AC/DC CONVERTER**

**A. CONFIGURATION OF THE PROPOSED AC-DC CONVERTER**

The topology of the programmable AC power supply is shown in Fig.1. It consists of two part: AC-DC rectifier and DC-AC inverter. The proposed single-phase single-stage isolated bridgeless AC-DC PFC converter is applied to AC-DC rectifier part. For inverter part, two single-phase full-bridge inverters would be connected in series or parallel for high or low output voltage applications because of the wide range AC output voltage. Therefore, the secondary side of the isolated AC-DC converter’s transformer has two windings for two independent output voltages. Since the two windings are identical, just only one winding is taken into consideration to simplify the analysis. The simplified circuit of the proposed single-phase single-stage isolated PFC converter is shown in Fig.2.

The proposed AC-DC topology is comprised of bridgeless rectifier diodes, series resonance circuit, active-clamped circuit, HF transformer and output rectifier diodes circuit. The bridgeless rectifier topology consists of two main power switches $S_1$, $S_2$ and two diodes $D_1$, $D_2$, and the two power
switches are driven by the same logical pulse width modulation (PWM) signal. The auxiliary power switch \( S_3 \) and clamp capacitor \( C_c \) is composed of active-clamped circuit, and it could limit the voltage stress of main power switches. The inductor \( L_r \), capacitor \( C_p \) on the primary side and capacitor \( C_s \) on the secondary side of HP transformer \( T_1 \) constitute the series resonant circuit. By designing the parameters of resonant circuit reasonably, ZVS turn-on of the main switches \( S_1, S_2 \) and auxiliary switch \( S_3 \) and ZCS turn-off of output rectifier diodes could be realized. In order to improve power density and conversion efficiency, SiC MOSFETs are used for power switches. \( D_{Sx} \) and \( C_{Sx} \) represent the anti-parallel diode and parasitic capacitor of the power switch \( S_x \) respectively, where \( x = 1, 2 \) and 3.

### B. OPERATION PRINCIPLE

In order to simplify the theoretical analysis, the following assumptions are made:

(i) As the switching frequency \( f_s \) is much higher than the AC input voltage frequency \( f_g \), it is assumed that the AC input voltage is constant within each switching frequency.

(ii) Input inductor \( L \), resonant capacitors \( C_p \) and \( C_s \), resonant inductor \( L_r \), clamping capacitor \( C_c \) and output capacitor \( C_o \) are ideal components, and \( C_p, C_s \ll C_c, L_r \ll L \).

(iii) The HF transformer is also an ideal electrical component, and the turns ratio of primary and secondary side satisfies \( n_p/n_s = n/1 \).

(iv) Output filter capacitor \( C_o \) and clamping capacitor \( C_c \) are large enough, and the values of output voltage \( U_o \) and clamped capacitor voltage \( U_{Cc} \) almost remain unchanged within each switching period.

(v) The resonant inductor \( L_r \) and equivalent capacitor \( C_r \) constitute the series resonant circuit, where capacitor \( C_p \) and \( C_s \) consists of resonant capacitor \( C_r \), and \( f_r = 1/(2\pi \sqrt{L_r C_r}) \). Resonant frequency \( f_r \) is higher than the switching frequency \( f_s \). In order to facilitate designation of the controller and realize the soft-switching operation of the power devices, the conduction time of the main power switches \( S_1 \) and \( S_2 \) must satisfied \( T_{on} \geq 0.5T_r \), where \( T_r = 1/f_r \).

In order to describe the soft-switching process for the power devices, the operation principle will be analyzed in details when the AC input voltage is in the positive half of voltage period, which is shown in Fig.3. When the AC input voltage is in the negative half of voltage period, the situation is similar. Assuming that the converter operates in continuous conduction mode (CCM), and there are seven different operating modes. The voltage and current of operation waveforms
are also shown in Fig. 4. A detailed description of each operation mode will be given below.

1) **MODE 1 (t0-t1)**

This stage begins when the energy stored by the parasitic capacitor \( C_s \) of the main switch \( S_2 \) is released. At time \( t_0 \), the anti-parallel diode \( D_{s2} \) of the main switch \( S_2 \) starts to conduct, and the bulk diode \( D_{s1} \) of the main switch \( S_1 \) has also conducted, ZVS turn-on for both main switches \( S_1 \) and \( S_2 \) is realized. During this process, auxiliary switch \( S_3 \) is turned off, and the output rectifier diodes \( D_3 \) and \( D_6 \) conducts. At time \( t_1 \), the resonant inductor current \( i_{Lr} \) becomes zero, the resonant capacitor voltage \( U_{Cp} \) reaches the maximum voltage \( U_{Cp(max)} \) and ZCS turn-off for the output rectifier diodes \( D_3 \) and \( D_6 \) could also be realized.

There are two circuit loops on the primary side of the transformer. The first circuit loop consists of AC voltage \( U_{in} \), inductor \( L \), main switch \( S_1 \), resonant capacitor \( C_p \), resonant inductor \( L_r \) and diode \( D_2 \). The other circuit loop consists of main switch \( S_1 \), resonant capacitor \( C_p \), resonant inductor \( L_r \) and the main switch \( S_2 \).

2) **MODE 2 (t1-t2)**

At this stage, main switches \( S_1 \) and \( S_2 \) remain conduction, auxiliary switch \( S_3 \) is still turned-off, and the input inductor \( L \) is linearly charged by the AC input voltage \( U_{in} \). As the direction of resonant inductor current \( i_{Lr} \) changes, the output rectifier diodes \( D_4 \) and \( D_5 \) begin to conduct. The current flowing through the main switch \( S_2 \) equals to the sum of input inductor current \( i_L \) and the resonant inductor current \( i_{Lr} \). The series resonant circuit resonates until the resonant inductor current \( i_{Lr} \) becomes zero at time \( t_2 \). The input inductor \( i_L \), resonant inductor current \( i_{Lr} \) and resonant capacitor voltage \( U_{Cp} \) can be expressed as

\[
i_L(t) = \frac{U_{in}}{L}(t - t_1) + i_L(t_1) \quad (1)
\]
\[
i_{Lr}(t) = i_{Lr(max)} \sin(\omega_r(t - t_1)) \quad (2)
\]
\[
U_{Cp}(t) = i_{Lr(max)} Z_r \cos(\omega_r(t - t_1)) + U_{Cp}(t_1) \quad (3)
\]
\[
i_{S2}(t) = i_L(t) - i_{Lr}(t) \quad (4)
\]

where, \( i_{Lr(max)} \) represents the negative peak value of resonant inductor \( i_{Lr} \), \( Z_r = \sqrt{(L_r/C_r)} \), \( \omega_r = \sqrt{(L_r C_r)} \), \( C_r = C_p C_s / (n^2 C_p + C_s) \).

In this process, resonant capacitor \( C_p \) on the primary side of the transformer begins to discharge. As the conduction time of main switch \( S_2 \) is greater than or equal to half of the resonant period \( T_{on} \geq 0.5 T_r \), resonant inductor \( i_{Lr} \) becomes zero at time \( t_2 \), the resonant capacitor voltage \( U_{Cp} \) drops to the minimum value of \( U_{Cp(min)} \). The operation principle of the resonant capacitor \( C_r \) on the secondary side of transformer is similar to capacitor \( C_p \). The duration of this mode is exactly equal to half of the resonance period \( 0.5 T_r \). The ZCS turn-off for the output rectifier diodes \( D_4 \) and \( D_5 \) also could be realized.

There are also two circuit loops on the primary side of the transformer. The first circuit loop consists of AC input voltage \( U_{in} \) inductor \( L \), main switch \( S_2 \) and diode \( D_2 \). The other circuit loop consists of main switch \( S_1 \), resonant capacitor \( C_p \), resonant inductor current \( i_{Lr} \) and the main switch \( S_2 \).

3) **MODE 3 (t2-t3)**

As shown in Fig. 3(c), main switch \( S_2 \) remains conduction and the inductor \( L \) is still charged by the AC input voltage, inductor current \( i_L \) continues to increase. As the resonant inductor current \( i_{Lr} \) becomes zero, the current flowing through the main switch \( S_2 \) is equal to the inductor current \( i_L \). The output filter capacitor \( C_o \) on the secondary side of the transformer provides energy for the load. This stage ends at the time \( t_3 \), when the main switch \( S_2 \) is turned off. The current flowing through the main switch \( S_2 \) and the capacitor \( U_{Cp} \) can be expressed as

\[
i_{S2}(t) = i_L(t) = \frac{U_{in}}{L}(t - t_2) + i_L(t_2) \quad (5)
\]
\[
U_{Cp}(t) = U_{Cp(t3)} = U_{Cp(min)} \quad (6)
\]

There are only one circuit loop on the primary side of the transformer, and it consists of AC voltage \( U_{in} \), inductor \( L \), main switch \( S_2 \) and diode \( D_2 \). All the diodes of the rectifier bridge on the secondary side of the transformer are in the off-state.

4) **MODE 4 (t3-t4)**

As shown in Fig. 3(d), main switch \( S_2 \) is turned off at time \( t_3 \), and the auxiliary switch \( S_3 \) does not turn on. The inductor current \( i_L \) charges the parasitic capacitor \( C_{S2} \) of the main switch \( S_2 \), and discharges the parasitic capacitor \( C_{S3} \) of the auxiliary switch \( S_3 \). As the value of parasitic capacitor \( C_{S3} \)
is small, the anti-parallel diode $D_{S3}$ will conduct quickly, and the input inductor current $i_L$ also charges the clamping capacitor $C_c$. The series resonant circuit composed of resonant inductor $i_L$ and equivalent capacitor $C_r$ starts to resonate, the resonant inductor current $i_L$ increases from zero, and the output rectifier diodes $D_3$ and $D_6$ start to conduct. The parasitic capacitor voltage $U_{C_{S2}}$ of the main switch $S_2$ rises from zero until it reaches the clamping capacitor voltage $U_{C_c}$. The input inductor current $i_L$, the source voltage $U_{C_{S2}}$ of the main switch $S_2$ and the resonant inductor $i_L$ are obtained from equations (7), (8) and (9) respectively

$$i_L(t) = \frac{U_{in} - U_{C_c}(t_3)}{L}(t - t_3) + i_L(t_3) \quad (7)$$

$$U_{C_{S2}}(t) = i_S(t - t_3) \quad (8)$$

$$i_L(t) = i_{Lr}(\text{max} + \sin(\omega_r(t - t_3))) = i_L(t) - i_{C_r}(t) \quad (9)$$

During this interval, the charging time of parasitic capacitor $C_{S2}$ and discharging time of parasitic capacitor $C_{S3}$ are very short, and the anti-parallel diode $D_{S3}$ of auxiliary switching $S_3$ starts conduction. As the value of clamping capacitor $C_c$ is much higher than parasitic capacitor $C_{S2}$ and $C_{S3}$, the capacitor voltage $U_{C_c}$ almost remains constant. The time interval of this stage can be obtained as

$$\Delta t_{34} = t_4 - t_3 = \frac{U_{C_c}C_{S2}}{i_{S2}(t_3)} \quad (10)$$

In order to realize the ZVS turn-on condition for auxiliary switch $S_3$, the energy stored in inductor $L$ must be greater than that stored in parasitic capacitors $C_{S2}$ and $C_{S3}$, that is, it must be satisfied

$$L > \frac{(C_{S2} + C_{S3})(U_{C_c}(t_3))^2}{i_L^2(t_3)} \quad (11)$$

There are three circuit loops on the primary side of the transformer. The first circuit loop consists of AC voltage $U_{in}$, inductor $L$, main switch $S_2$ and diode $D_2$. The secondary circuit loop consists of AC voltage $U_{in}$, inductor $L$, main switch $S_1$, auxiliary switch $S_3$, clamping capacitor $C_e$ and diode $D_2$. The third circuit loop consists of AC voltage $U_{in}$, inductor $L$, main switch $S_1$, resonant capacitor $C_p$, resonant inductor current $i_{Lr}$ and diode $D_2$.

5) MODE 5 ($t_4$-$t_5$)

As shown in Fig.3(e), the driver signal becomes high for auxiliary switch $S_3$ at time $t_4$, the ZVS turn-on for auxiliary switch $S_3$ could be realized because of the conduction of anti-parallel diode $D_{S3}$. The input inductor current $i_L$ decreases with the slope of $(U_{in} - U_{C_c})/L$, and the clamping capacitor current $i_{C_c}$ becomes zero at time $t_5$. Since the resonant inductor current $i_{Lr}$ keeps charging the resonant capacitor $C_p$, the resonant capacitor voltage $U_{C_p}$ increases. The input inductor current $i_L$ can be expressed as

$$i_L(t) = \frac{U_{in} - U_{C_c}(t_3)}{L}(t - t_4) + i_L(t_4) \quad (12)$$

6) MODE 6 ($t_5$-$t_6$)

As shown in Fig.3(f), when the resonant inductor current $i_{Lr}$ is equal to input inductor current $i_L$. The current $i_{S3}$ becomes zero and starts to flow from the drain to the source. At time $t_6$, the auxiliary switch $S_3$ is turned off, and this stage ends. During the interval, the direction of resonant inductor $i_{Lr}$ is always positive, and the resonant capacitor $C_p$ is always charged. The resonant capacitor voltage $U_{C_p}$ increases, and the diodes of $D_3$ and $D_6$ on the secondary side of the transformer conducts.

There are also two circuit loops on the primary side of the transformer. The first circuit loop consists of AC voltage $U_{in}$, inductor $L$, main switch $S_1$, resonant capacitor $C_p$, resonant inductor $i_{Lr}$ and diode $D_2$. The other circuit loop consists of AC voltage $U_{in}$, inductor $L$, main switch $S_1$, resonant capacitor $C_p$, resonant inductor $i_{Lr}$ and diode $D_2$.

7) MODE 7 ($t_6$-$t_7$)

As shown in Fig.3(g), at time $t_6$, auxiliary switch $S_3$ is turned off, and the resonant inductor current $i_{Lr}$ discharges the parasitic capacitor $C_{S2}$ of main switch $S_2$ and charges the parasitic capacitor $C_{S3}$ of the auxiliary switch $S_3$. Resonant inductor current $i_{Lr}$ begins to decrease, but the direction remains positive. In order to ensure the ZVS turn-on for the main switch $S_2$, the parasitic capacitor voltage $U_{C_{S2}}$ needs to be zero before the end of this stage to enable the conductions of the anti-parallel diode $D_{S2}$. Therefore, the electric energy stored in the resonant inductor $L_r$ must be larger than that stored in the parasitic capacitor $C_{S2}$, so that the energy stored in the parasitic capacitor $C_{S2}$ is completely released. The resonant inductor $L_r$ must meet the following requirements

$$L_r > \frac{(C_{S2} + C_{S3})(U_{C_c}(t_6))^2}{i_{Lr}^2(t_6)} \quad (13)$$

At time $t_7$, as the conduction of anti-parallel diode $D_{S2}$, ZVS turn-on condition can be achieved for main switch $S_2$ if the driver signal becomes high level, and the next switch period begins.

There are also two circuit loops on the primary side of the transformer. The first circuit loop consists of AC voltage $U_{in}$, inductor $L$, main switch $S_1$, resonant capacitor $C_p$, resonant inductor $i_{Lr}$ and diode $D_2$. The other circuit loop consists of auxiliary switch $S_1$, resonant capacitor $C_p$, resonant inductor $i_{Lr}$ and main switch $S_2$.

It should be noted that when the input grid voltage or load changes, some of the power devices may not work in the soft-switching condition, and the corresponding working mode would be slightly different, but overall the operation process is similar. In order to avoid repetition, it will not be described.
in details here. The soft-switching operation condition for the power switches will be introduced in the following section.

III. STEADY-STATE ANALYSIS

A. VOLTAGE GAIN

In this section, the voltage gain of the proposed topology will be discussed for details. When the AC voltage $U_{in}$ is positive, the equivalent circuit of the proposed topology could be simplified as shown in Fig.5. Fig.5(a) and (b) present the equivalent circuit of the main switch $S_2$ in on-state or off-state respectively. When the main switch $S_2$ is turned off, as the clamping capacitor $C_c$ is relatively large, the clamped capacitor voltage $U_C$ is almost constant during this interval, and the current $i_{C_c}$ is close to zero, so the clamped circuit can be ignored for simplicity. When the main switch $S_2$ is in the conduction state as shown in Fig.5(a), the voltage at both ends of the inductor $L$ is $U_{in}$, and the products of volt-seconds is $U_{in}DT_s$. Where, $D$ and $T_s$ represent the duty ratio of the main switch $S_2$ and switch period respectively. When the main switch $S_2$ is in the off state as shown in Fig.5(b), the resonant inductor $L_r$ is in series with $C_r$, and the average voltage at both end of the resonant inductor $i_{L_r}$ is zero. Similarly, the product of the volt-seconds at both ends of the inductor $L$ can be expressed as $(2nU_0 - U_{in})(1 - D)T_s$. Based on the principle of volt-seconds balance of the input inductor $L$, equation (14) can be obtained as

$$U_{in}DT_s = (2nU_0 - U_{in})(1 - D)T_s$$

and the ratio of output voltage $U_o$ to input voltage $U_{in}$ can be obtained as

$$\frac{U_o}{U_{in}} = \frac{1}{2n(1 - D)}$$

When the AC voltage $U_{in}$ is negative, the same expression of voltage gain can be obtained based on the similar analytical method, and the voltage gain $M$ of the proposed topology is expressed as

$$M = \frac{U_o}{|U_{in}|} = \frac{1}{2n(1 - D)}$$

B. ZVS AND ZCS CONDITION

Based on the analysis of the aforementioned section II, in order to realize the ZVS turn-on of the main switch $S_2$, the energy stored in resonant $L_r$ should fully charge the parasitic capacitor $C_{S2}$ of the main switch $S_2$ and discharge the parasitic capacitor $C_{S3}$ of switch $S_3$, and enable the anti-parallel diode $D_{S2}$ of the main switch $S_2$ to conduct. In this case, the current flowing through the main switch $S_2$ should meet the requirement of $i_{S2}(t_0) = i_L(t_0) - i_{L_{r}}(t_0) < 0$, which satisfies equation (17) and (18).

$$i_{S2}(t_0) = i_L(t_0) - i_{L_{r}}(t_0) = \frac{P_o}{U_{in}} \frac{DT_s U_{in}}{2L} = \frac{P_o}{U_{in}} \frac{DT_s U_{in}}{2L} - \frac{I_o}{4nC_{c}sZ_{r}} \sin((\omega_s(1 - D)T_s)$$

$$\frac{1}{2}(L_{r}i_{L_{r}}^{2}(max) - L_{r}^{2}i_{L_{r}}^{2}(min)) \geq \frac{1}{2} C_{S2} + C_{S3}U_{C_{c}}^{2}$$

The situation for switch $S_1$ is similar when input AC voltage $U_{in}$ is negative. In order to realize the ZVS turn-on condition for auxiliary switch $S_3$, the current $i_{L_{r}}$ stored in the inductor $L$ discharges the parasitic capacitor $C_{S3}$ of the auxiliary switch $S_3$ and charges the parasitic capacitor $C_{S2}$ of the main switch $S_2$, and enables the anti-parallel diode $D_{S3}$ of the auxiliary switch $S_3$ to conduct, which satisfies equation (19) and (20).

$$i_{S3}(t_3) = i_L(t_3) - i_{L_{r}}(t_3) = i_{L_{r}}(max) = \frac{P_o}{U_{in}} + \frac{DT_s U_{in}}{2L}$$

$$\frac{1}{2} L_{r}i_{L_{r}}^{2}(max) \geq \frac{1}{2}(C_{S2} + C_{S3})U_{C_{c}}^{2}$$

In fact, at time $t_3$, resonant inductor current $i_{L_{r}}$ is zero, and equation (20) can be easily satisfied with large input inductor $L$ and small parasitic capacitor $C_{S2}, C_{S3}$. The ZVS condition for auxiliary $S_3$ can be realized nearly in overall load range. equation (18) may not be satisfied with a small resonant inductor $L_r$ or light load. With large resonant inductor $L_r$ and small input inductor $L$ may extend the ZVS range, but it would lead to large ripple current of $i_{L_{r}}$ and it should be chosen based on the tradeoff between power losses and current ripple of the whole system.

The ZCS condition for output rectifier diodes can be easily deduced according to mode 1 and 2 in section II. When the nonconduction time of the switch $S_1$ or $S_2$ is greater than half of the resonance period, $D_3$ and $D_6$ can work in ZCS.
condition, which satisfies

$$T_{on\_min} = D_{max}T_s = \frac{|U_{in}|}{nU_o}T_s \geq \frac{1}{2}T_r$$  \hspace{1cm} (21)

Similarly, the ZCS operation condition of diodes $D_4$ and $D_5$ can be obtained based on equation (22)

$$T_{on\_min} = D_{min}T_s = \frac{|U_{in}|}{nU_o}T_s \geq \frac{1}{2}T_r$$  \hspace{1cm} (22)

where, $D$, $T_s$, and $T_r$ represent the duty ratio of switch $S_1$ or $S_2$, switching period and resonant period respectively.

![FIGURE 6. Equivalent circuit with the secondary quantities are referred to the primary.](image)

**C. VOLTAGE STRESS ANALYSIS**

In order to simplify the analysis, the winding on the secondary side of the HF frequency is referred to the primary side, and the equivalent circuit is obtained as shown in Fig.6. The maximum voltage stress of the main switch can be expressed as

$$U_{S1\_max} = U_{S2\_max} = 2nU_o + U_{C_r\_max}$$  \hspace{1cm} (23)

When the system works stably, the current flowing through the diodes of $D_4$ and $D_5$ satisfies that

$$I_{D4\_avg} = I_{D5\_avg} = \frac{I_o}{2n}$$

$$= \frac{1}{T_s}\int_{t_1}^{t_5} \frac{1}{Z_r}U_{C_r\_max}\sin(o_r t)dt$$

$$= \frac{1}{T_s}\int_{0}^{1/T_r} \frac{C_r}{L_r}U_{C_r\_max}\sin(o_r t)dt$$  \hspace{1cm} (24)

And $U_{C_r\_max}$ also can be deduced from equation (24)

$$U_{C_r\_max} = \frac{I_o}{4nC_ff_s}$$  \hspace{1cm} (25)

It can be easily analyzed that the maximum voltage stress of auxiliary switch $S_3$ satisfies

$$U_{S3\_max} = U_{C_r\_max} = \frac{|U_{in\_max}|}{1-D}$$  \hspace{1cm} (26)

The maximum voltage stress of diode $D_1$ and $D_2$ also can be derived as the equation (26). It is also easy to conclude that the voltage stress of rectifier diodes $D_3$ and $D_6$ on the secondary side of the transformer are always equal the output voltage $U_o$ regardless of the duty ratio $D$.

$$U_{D1\_max} = U_{D2\_max} = U_{C_r\_max}$$  \hspace{1cm} (27)

$$U_{D3\_max} = U_{D5\_max} = U_o$$  \hspace{1cm} (28)

**IV. DESIGN CONSIDERATION**

In this section, the design consideration of the proposed single-stage PFC converter is presented, and the detailed specifications of the bridgeless AC-DC PFC converter is shown in Table 1.

| Parameter | Value |
|-----------|-------|
| AC input voltage range, $U_{in}$ | 175~265V |
| AC input frequency range, $f_o$ | 47~53Hz |
| Input current ripple, $f_r$ | 20% |
| Output voltage ripple, $\delta$% | 5% |
| Turns ratio of the transformer, $n$ | 6:1 |
| DC output voltage, $U_o$ | 300 V |
| Rated power, $P_o$ | 3000W |

**A. CHOSEN OF KEY COMPONENT**

1) INPUT INDUCTOR $L$ AND OUTPUT CAPACITOR $C_o$  
Assuming that the current ripple $1\%$ of the inductor is $20\%$ and the converter operates in CCM mode. Ignoring the power losses and based on the equality of AC input power and DC output power, it is easy to calculate the inductance value as follows:

$$L = \frac{U_{in\_min} \cdot D_{min} \cdot f_r \cdot I_{peak\_max}}{f_s \cdot U_o^2 \cdot \delta} = 441uH$$  \hspace{1cm} (29)

If the output voltage ripple is $\delta = 5\%$, the capacitance value of the output filter capacitor $C_o$ can be easily calculated according the equation(30)

$$C_o = \frac{P_o}{4f_sU_o^2 \cdot \delta} = 3333uF$$  \hspace{1cm} (30)

Finally, standard values of the inductor $L = 450uH$ and capacitor $C_o = 3300uF$ are chosen for the experimental prototype.

2) RESONANT INDUCTOR $L_r$, RESONANT CAPACITOR $C_r$ AND CLAMPING CAPACITOR $C_c$  
The resonant circuit is the core of the whole converter, which is composed of resonant inductor $L_r$, resonant capacitor $C_r$ and $C_c$. Reasonable design of resonant parameters can not only ensure the normal operation of the main circuit, but also enable the power devices of the converter to work in soft switching condition.

The selection of resonant inductor $L_r$ should ensure that the main switch $S_1$ and $S_2$ can realize ZVS turn-on operation, which satisfies equation (18) and (20). Finally, $L_r = 5uH$ and $C_r = 0.52uF$ are chosen for the resonant circuit, where $C_r = \frac{C_p}{n^2C_p + C_s}$, $C_p = 2.2uF$, $C_s = 0.5uF$ and $n = 6/7$. The resonant frequency can be calculated as $f_r = \frac{1}{2\pi \sqrt{L_r C_r}} = 98.7kHz$. Meanwhile, as the conduction time of the main switch $S_2$ or $S_1$ satisfies $T_{on} \geq 0.5T_r$, and the soft switching condition for the power components could be realized.

About the selection of clamping capacitor $C_c$, it is expected that the value of the clamping capacitor would be as large as possible, so that the voltage $U_{C_c}$ of clamping capacitor will remain basically unchanged during the switch period $T_s$.  

**TABLE 1. Design specifications.**
However, the clamping capacitor with an excessive capacity will make the volume of the converter too large. In addition, based on the operation mode 6 mentioned in section II, resonant inductor $L_r$, resonant capacitor $C_r$ and clamping capacitor $C_c$ will actually resonate when the auxiliary switch $S_3$ turns off. Without considering the effect of clamping capacitor $C_c$ for the main circuit, the value of the clamping capacitor $C_c$ must be much greater than that of resonant capacitor $C_r$, so $C_c = 50\mu F$ is chosen for the clamping circuit. As a result of $C_c \gg C_r$, the voltage $U_{Cc}$ of clamping capacitor $C_c$ keeps almost constant during the switch period.

3) HF TRANSFORMER
The transformer must be designed to transfer the instantaneous input power of 3kW at the conditions of 85kHz switching frequency and 300V output voltage. Based on the analysis mentioned in section II, as the existence of capacitors $C_p$ and $C_s$ in the primary and secondary side of the transformer, there is no DC voltage component, it would greatly reduce the volume of the transformer. The transformer is composed of EE55 magnetic core, and two magnetic core are stacked to increase the core area as shown in Fig.7. The copper foil is used for the primary and secondary-side windings, and the size is $0.1\text{mm}$ (thickness)$\times 27\text{mm}$ (width). The secondary side of the transformer has two windings, and they are connected in series when the output AC voltage of the corresponding inverter is between 150V and 300V, when output voltage of AC source is lower than 150V, these two wings are connected in parallel. The turns of primary and secondary-side are 6 and 7 respectively. Some key components for the PFC converter are shown in Table.2.

![HF transformer](image1)

**FIGURE 7.** HF transformer. (a) Winding diagram. (b) Physical model.

| Parameter                      | Value |
|-------------------------------|-------|
| Input inductor, $L_r$         | 450mH |
| Resonant inductor, $L_r$      | 5uH   |
| Primary resonant capacitor, $C_p$ | 2.2uF |
| Secondary resonant capacitor, $C_s$ | 0.5uF |
| Output filter capacitor, $C_o$| 2200uF |
| SiC Power MOSFETs, $S_1 \sim S_3$ | CSM006509J |
| Diodes, $D_1 \sim D_2$        | C4D90120D |
| Diodes, $D_3 \sim D_6$        | RMRF1560 |

### TABLE 2. Key components for the PFC converter.

**B. CONTROL STRATEGY**
The control block diagram of the proposed bridgeless isolated PFC rectifier is shown in Fig.8, and the three electrical parameters of input AC voltage $U_{in}$, inductor current $i_L$ and output voltage $U_o$ are measured respectively, it is not necessary to sense the positive or negative AC voltage cycle. The driver logic signals for main switches of $S_1$ and $S_2$ are same, and the driver signal of auxiliary switch $S_3$ is complementary to them. A standard PFC IC UC3854 could be chosen for the controller, which uses average current-mode control to accomplish fixed-frequency current control with stability and low distortion of grid current.

**FIGURE 8.** Control block diagram of the single-stage PFC converter.

**FIGURE 9.** Experimental prototype of the single-stage PFC converter.

3) HF TRANSFORMER
The transformer must be designed to transfer the instantaneous input power of 3kW at the conditions of 85kHz switching frequency and 300V output voltage. Based on the analysis mentioned in section II, as the existence of capacitors $C_p$ and $C_s$ in the primary and secondary side of the transformer, there is no DC voltage component, it would greatly reduce the volume of the transformer. The transformer is composed of EE55 magnetic core, and two magnetic core are stacked to increase the core area as shown in Fig.7. The copper foil is used for the primary and secondary-side windings, and the size is $0.1\text{mm}$ (thickness)$\times 27\text{mm}$ (width). The secondary side of the transformer has two windings, and they are connected in series when the output AC voltage of the corresponding inverter is between 150V and 300V, when output voltage of AC source is lower than 150V, these two wings are connected in parallel. The turns of primary and secondary-side are 6 and 7 respectively. Some key components for the PFC converter are shown in Table.2.

**TABLE 2. Key components for the PFC converter.**

| Parameter                      | Value |
|-------------------------------|-------|
| Input inductor, $L_r$         | 450mH |
| Resonant inductor, $L_r$      | 5uH   |
| Primary resonant capacitor, $C_p$ | 2.2uF |
| Secondary resonant capacitor, $C_s$ | 0.5uF |
| Output filter capacitor, $C_o$| 2200uF |
| SiC Power MOSFETs, $S_1 \sim S_3$ | CSM006509J |
| Diodes, $D_1 \sim D_2$        | C4D90120D |
| Diodes, $D_3 \sim D_6$        | RMRF1560 |

**V. EXPERIMENTAL RESULTS**
In order to verify the correctness and feasibility of the proposed sing-stage isolated bridgeless PFC converter topology and its operation principle, a 3kW experimental prototype based on SiC power device is built as shown in Fig.9. The HF transformer has two wings with rectifier diodes on the secondary side and they are connected in parallel. Due to the high switch frequency of the SiC power devices, the standard low-cost IC of UC3854 is chosen as the controller.

Experimental waveforms are measured at rated input AC voltage of 220V and output power of 3kW resistance load. Fig.10 shows the measured experimental waveform of input
AC voltage $U_{in}$, input inductor current $i_L$ and output capacitor voltage $U_o$, it can be observed that the input inductor current is sinusoidal and in phase with the input voltage, and the output capacitor voltage is stable at 300V. The input power factor and the THD value of grid current are measured by analyzer Tek PA3000, the PF value is as high as 0.995, and the THD value is 3.9%, it also can meet IEC 61000-3-2 standard.

The voltage of resonant capacitor $C_p$ in the primary side and the capacitor $C_s$ in the secondary side of HF transformer is measured as shown in Fig.11. Capacitor voltage $U_{Cp}$ and $U_{Cs}$ have the same phase angles, and the phase and amplitude of the capacitor voltage $U_{Cp}$ is almost same with the rectifier AC input voltage $|U_{in}|$, which is consistent with the analysis of equivalent circuit in steady-state operation as shown in Fig.5.

The experimental waveforms of ZVS turn-on operation for main switches $S_1$, $S_2$ are shown in Fig.12. When the driver signal of the main switches become high, the voltage between the drains and sources have been reduced to zero due to the conduction of anti-parallel diodes, thus the ZVS turn-on operation of the main switches can be realized. The soft-switch operation principle of the auxiliary $S_3$ is similar, and the ZVS turn-on waveform is shown in Fig.13.

The soft switching waveform of output rectifier diodes $D_3$ and $D_5$ in the secondary side of HF transformer is shown in Fig.14. The voltage waveforms of anode and cathode and the conduction current waveform for the rectifier diodes $D_3$ and $D_5$ are shown in Fig.14(a) and Fig.14(b) respectively. Due to the change of direction of resonant inductor $i_L$, within a switching cycle, the ZCS turn-off operation of the output rectifier diodes $D_3 \sim D_6$ can be achieved.

The conversion efficiency of the proposed single-stage isolated converter is measured by Tektronix PA3000. As the two-stage PFC topology proposed in [5] is widely
implemented in programmable AC source applications at present, and it is compared with the proposed single-stage isolated converter. Meanwhile, the 3kW single-stage isolated ZCS current-fed full-bridge converter recently proposed in [23] is also chosen for comparison as shown in Fig.15. The conversion efficiency of the proposed single-stage topology is significantly higher than the other two topologies, and a peak efficiency of 95.7% could be achieved.

The measured efficiency curve and PF value with different input voltage at 3kW load condition are provided in Fig.16. Nearly unity PF can be achieved and a high performance of conversion efficiency is obtained over the entire range of input voltage.

Fig.17 shows the power loss analysis of the components at rated power when input voltage is 220V, and the loss calculation method for each component is routine [23], [24], [27], [28]. All the results are at 85kHz switching frequency, and the power losses will be lower with higher switching frequency. As the ZVS turning-on condition for power MOSFETs, the switching losses can be minimized, and with the ZCS turning-off condition for output rectifier diodes, the snubber circuit is small or not necessary, this also improves the conversion efficiency of the converter. The majority of the power losses comes from the transformer, diodes and power switches.

Table 3 shows the comparison between the proposed single-stage bridgeless isolated PFC converter with the presented isolated AC-DC converter in [5], [21], [22], [23], [24], [25] and [26], and the comparison results show that the

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**TABLE 3. Performance comparison.**

| Reference | Input voltage range | Output voltage range | Output Power | Switch frequency | Number of switches | Number of magnetic components | Peak efficiency | Control complexity | Cost |
|-----------|---------------------|----------------------|--------------|-----------------|-------------------|-------------------------------|----------------|------------------|------|
| Ref.[5]   | 85~265V             | 200~450V             | 3.3kW        | 200kHz          | 6 MOSFETs, 11 diodes | 4                            | 93.6%          | Complexity      | High |
| Ref.[21]  | 220V                | 240V                 | 1.6kW        | 25kHz           | 1 MOSFETs, 5 diodes | 3                            | 91%            | Simple          | Low  |
| Ref.[22]  | 120V                | 220~336V             | 1.5kVA       | 100kHz          | 8 MOSFETs         | 4                            | 96.5%          | Simple          | Medium |
| Ref.[23]  | 208V                | 400V                 | 3kW          | 10kHz           | 5 MOSFETs, 13 diodes | 2                            | 94%            | Simple          | High  |
| Ref.[24]  | 207~253V            | 280~420V             | 3.3kW        | 20~120kHz       | 8 MOSFETs         | 3                            | 95%            | Complexity      | Medium |
| Ref.[25]  | 220V                | 60V                  | 1.4kW        | 50kHz           | 8 MOSFETs, 4 diodes | 3                            | 96%            | Simple          | Medium |
| Ref.[26]  | 220V                | 320~400V             | 3.3kW        | 100kHz          | 10 MOSFETs        | 4                            | 96.7%          | Complexity      | High  |
| Proposed  | 175~265V            | 300V                 | 3kW          | 85kHz           | 3 MOSFETs, 4 diodes | 3                            | 95.7%          | Simple          | Low  |

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**FIGURE 15.** Comparison of conversion efficiency with different topologies at rated input voltage.

**FIGURE 16.** Measured efficiency curve and PF with different input voltage at 3 kW load condition.

**FIGURE 17.** Loss analysis of the proposed PFC converter at rated power.
topology proposed in this paper demonstrates a good performance for single-phase high-power applications based on the tradeoff of the cost, control complexity and the conversion efficiency.

VI. CONCLUSION

In this paper, a single-stage bridgeless isolated AC-DC converter topology with PFC is proposed for programmable AC power source applications. The ZVS turn-on of main and auxiliary switches and ZCS turn-off of output rectifier diodes can be realized based on LC series-resonant technology and active-clamping method under certain load conditions. Finally, a 3kW experimental prototype based on SiC power devices is built to verify the correctness and feasibility of the topology, 95.7% of peak conversion efficiency and 3.9% of grid THD current can be achieved. Some features of the proposed single-stage isolated converter can be summarized as follows:

1) Low cost and high energy conversion efficiency of the proposed converter can be achieved because of few components needed, elimination of the diode rectifier bridge, ZVS turn-on for power switches and ZCS turn-off for output diode rectifier bridge.

2) There is no need to detect the positive and negative half cycle of AC input voltage as the two main switches are driven by the same logic PWM signals.

3) The control circuit is simple and it could be easily implemented by low cost MCU or available analog IC (UC3854).

The single-stage bridgeless converter topology proposed in this paper can provide constant output voltage, low THD value of grid current and nearly unity PF for programmable AC power source applications with galvanic isolation, and it can also be used in other applications, such as battery charges, uninterruptable power supply, etc.

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