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GaN-based ZVS Bridgeless Dual-SEPIC PFC Rectifier with Integrated Inductors

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Abstract- This paper investigates the gallium nitride (GaN) based bridgeless dual single-ended primary inductor converter (SEPIC) power factor correction (PFC) with full input voltage range zero-voltage-switching (ZVS) turn-on for the application of step down AC-DC converter. The operation principle for the bridgeless Dual-SEPIC PFC and the theoretical analysis for the ZVS operation has been investigated. Furthermore, a new magnetic integration has been proposed to assemble all three inductors, including one input inductor and two output side inductors into one E-I-E core. The integrated inductor reduces the total ferrite volume and makes the converter more compact. The inductance design for ZVS SEPIC PFC and the magnetic reluctance modeling for the E-I-E core with the coupled inductor has been analyzed. The effective equivalent inductance of the input inductor can be implemented with a much less number of turns by a carefully designed coupling coefficient. Finally, a 300W GaN-based MHz bridgeless Dual-SEPIC PFC with the integrated inductors is developed and tested with full-range ZVS, 97% peak efficiency based on the ZVS extension strategy.

Index Terms: Bridgeless SEPIC PFC, integrated inductors, ZVS turn-on, E-I-E core.

I. INTRODUCTION

The main power supplies used in telecom, on-board chargers, servers, and industrial PSU systems need to convert the AC line power to a constant DC output voltage. Thus, rectifiers have been employed unavoidably. These systems usually range from a hundred watts to several kilowatts. The major problem for conventional rectifiers is severe harmonic pollution. Therefore, front-end power factor correction (PFC) converters are widely used to shape the input current of the power supply to meet the power factor (PF) standard and the current THD norms such as IEC 61000-3-2 [1]-[3].

As the most efficient and simple PFC circuit, the boost type PFC is extensively adopted. Because of the intrinsic boost characteristic, the output voltage of the boost PFC must be higher than the peak of the AC line voltage, which is typically 400 V [4]-[8]. Thus, in some step-down applications, a second stage is required to convert the output voltage of the boost PFC to a specific voltage lower than the peak input AC voltage. The two-stage converter will degrade the system efficiency as well as the power density. Therefore, the SEPIC PFC is prevailing in those applications. The output voltage of SEPIC PFC can be regulated either higher or lower than the AC input voltage. Thus, SEPIC converters are widely employed in LED lighting systems, wireless power transfer systems, standalone photovoltaic systems, and portable power applications [9],[10].

Fig. 1. (a) Conventional Sepic PFC with the diode bridge rectifier. (b) Type I bridgeless SEPIC PFC. (c) Type II bridgeless Dual SEPIC PFC. (d) modified bridgeless Dual SEPIC PFC.

The conventional SEPIC PFC consists of a bridge rectifier followed by a DC-DC SEPIC converter, as shown in Fig. 1(a) [9]-[13]. The high conduction loss caused by the forward voltage drop of the diode bridge rectifier will degrade the overall system efficiency dramatically, and the heat generated by the bridge rectifier may need an extra heatsink. Hence, to increase the power supply efficiency, the bridgeless SEPIC

Fig. 2. Three operation modes for Sepic PFC. (a) CCM. (b) CRM. (c) DCM.

PFC circuit topologies have been extensively researched [14]-[30]. Fig.1 (b) [19]-[21] and Fig.1(c) [18], [25]-[27] are two typical types bridgeless SEPIC PFC topologies. Without the forward bridge rectifier, it allows the current to flow through a minimum number of semiconductor devices. Thus, the converter conduction losses can be significantly reduced. The Type I bridgeless SEPIC PFC in Fig.1 (b) has two more diodes on the input side that will introduce more conduction loss. The Type II bridgeless SEPIC PFC in Fig.1(c) has two more passive components.

To reduce the volume of the passive component, increasing the switching frequency has been proved as an effective method [7]. The emerging of GaN devices, which have a low on-resistance, fast switching speed, and zero reverse recovery loss, have pushed the switching frequencies to the megahertz range for power electronics equipment [39]. Thus, the size of the magnetic components has been dramatically reduced. Since the state-of-the-art SEPIC PFC prototypes mainly work under discontinuous conduction mode (DCM), the converters are operating in hard switching, which will still cause a substantial turn-on power loss when the switching frequency increased. Hence, the switching frequency of the state-of-the-art SEPIC PFC ranges from tens to hundreds of kHz, it has still big challenge to push the switching frequency to megahertz.

To further increase the switching frequency, the switching loss must be dealt with. In particular, the turn-on loss of GaN FET dominates the power loss and is the major limitation for achieving high efficiency. Conventional SEPIC PFC with the output diode has low efficiency at high switching frequency since the ZVS operation cannot be obtained for all switches. [9],[13]. Thus, the realization of ZVS for the main switch is a key point to design a high-frequency, high-efficiency SEPIC PFC rectifier.

Some researches have been done to implement ZVS for SEPIC PFC [10],[30],[31]. In [30], a soft-switching cell composed of an active switch, a resonant inductor, and a resonant capacitor was applied into the SEPIC PFC to obtain ZVS to promote the converter efficiency. The active clamped SEPIC PFC was studied in [10] and [31]. By adding extra clamping capacitance and MOSFET, partial ZVS can be achieved. However, All the existing studies cannot accomplish the fully ZVS operation without adding any auxiliary components. This paper investigates the bridgeless dual SEPIC PFC with the full range ZVS for all switches, as shown in Fig. 1(d). By replacing the output diode with MOSFET, the

negative current is provided for main switches S1/S2 to discharge the output capacitors. So ZVS is guaranteed in the full range of the input voltage. Thus, without adding any extra components, it can maximize efficiency at a high switching frequency.

The comparison of the three bridgeless SEPIC PFC topologies is shown in Table I. By integrating all the three discrete inductors of the proposed topology in one E-I-E magnetic core, it has 7 components in total, which has the minimum components among the three topologies. In addition, there are no diodes in the line current path. For conventional two types bridgeless SEPIC PFC, the ZVS turn-on cannot be fully achieved, which will limit the switching frequency as well as the efficiency.

Table I
Comparison of three Bridgeless SEPIC topologies

Count	Type I Bridgeless SEPIC PFC	Type II Bridgeless SEPIC PFC	Modified SEPIC PFC
MOSFET	2	2	4
Diode	3	2	0
Inductor	2	3	3 → 1
Capacitor	1	2	2
Total	8	9	9 → 7
ZVS	Partial	Partial	Full range

The operation mode of the conventional PFC can be roughly classified into three categories depending on the waveform of the input inductor current, namely, the continuous conduction mode (CCM), the critical conduction mode (CRM), and the discontinuous conduction mode (DCM) [36]. In CRM mode, ZVS can be achieved with the correct digital control for the boost-type PFC [6]-[8],[38]-[40]. However, when it comes to the SEPIC PFC, the operation mode cannot be classified with the same principle because SEPIC PFC consists of two inductors. In contrast, the operation mode is determined by the sum of two inductor current ($i_{Lin} + i_{Ll}$), which is the current flows through the main switch S_1 when S_1 is ON. When S_1 is OFF, and S_3 is ON, ($i_{Lin} + i_{Ll}$) is output current i_{S3} flow through the synchronous switch S_3 . The input inductor current i_{Lin} commonly designs to work at continuous conduction mode. The inductor current waveforms of SEPIC PFC for three operation modes are depicted in Fig. 2. If the main switch S_1 turns on when i_{S3} is exactly decreased to zero, this is CRM mode, as shown in Fig.

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2(b). If the main switch S_1 turns on before the i_{S3} decrease to zero, this is CCM. Otherwise, it is DCM.

The DCM has been widely used for SEPIC PFC because of the ZCS turn-off for the diode [8]-[29]. However, the ZVS cannot be achieved in DCM mode. The modified bridgeless SEPIC PFC is working under CRM mode (also call boundary conduction mode, BCM) by adding extended conduction time for the synchronous MOSFET device. However, the input inductor still works under continuous conduction mode with a small current ripple even the converter is working in DCM or CRM mode, which may increase the system efficiency and minimize the volume of the EMI filter design.

To further make the converter more compact and reduce the cost, the integrated inductors have been investigated for the bridgeless Dual-SEPIC PFC in this paper. In the DC-DC SEPIC converter, the integrated inductor has been used [33]-[34]. By tightly coupled the input inductor and output inductor, the effective inductance can be double, and thus the current ripple can be reduced. For the AC-DC SEPIC PFC converter, this integrated inductor usually applied to dual-phase interleaved SEPIC PFC with a bridge rectifier [11],[12], the two input inductors, and two output inductors integrated separately with different magnetic cores. In [37], two integrated inductors have been conducted for the four inductors of the bridgeless SEPIC PFC. Still, two magnetics core needed. This paper proposed the integrated inductors, making it possible to assemble all the three inductors of the bridgeless Dual-SEPIC PFC into one E-I-E magnetic core. Based on the equivalent magnetic reluctance model, the theoretical analysis shows that the equivalent inductance can be enlarged several times to the self-inductance. The effective equivalent inductance of the input inductor can be implemented with fewer turns by an appropriately designed coupling coefficient.

In this paper, the full-range ZVS for the bridgeless GaN-based MHz Dual-SEPIC PFC is achieved. The operation principle of the topology and the ZVS control is investigated. Then the integrated inductors are designed to assemble all the magnetic components into one E-I-E magnetic core. The inductance design for the bridgeless SEPIC PFC and the magnetic modeling for the E-I-E core with the coupled inductor has been analyzed.

This paper is organized as follows. The principle of the operation stage and the theoretical analysis of ZVS implementation are presented in Section II. The inductance design and the coupling coefficient selection are presented in Section III. Hardware implementation is introduced in Section IV. Furthermore, a 300 W prototype of the SEPIC PFC is demonstrated, and the experimental results are shown in Section V, and Section VI concludes this paper.

II. ANALYSIS OF THE Dual-SEPIC PFC TOPOLOGY

A. Principle of operation

The bridgeless Dual-SEPIC PFC converter has two identical SEPIC converters, each operating in alternate half-line cycles,

based on the polarity of the input AC voltage. It consists of three inductors: one input inductor L_{in} and two output inductors L_1 , L_2 . L_1 and L_{in} conduct at the positive half-line cycle, and L_2 and L_{in} conduct at the negative half-line cycle. S_2 is always ON, and S_4 is always OFF in the positive half-line cycle. In the negative half-line cycle, S_1 is always ON, and S_3 is always OFF. The analysis and discussions of the positive half-line cycle are similar to the negative half-line cycle, so only the positive half-line cycle has been analyzed in this paper. Assuming the input current $i_{Lin(t)}$ is in phase with the input voltage $V_{in(t)}$, defining the input AC line voltage and input current as:

$$V_{in}(t) = \sqrt{2} \cdot V_{in_rms} \cdot \sin(\omega_{line}t) \quad (1)$$

$$i_{Lin}(t) = \sqrt{2} \cdot I_{in_rms} \cdot \sin(\omega_{line}t) \quad (2)$$

Where V_{in_rms} and I_{in_rms} are the input RMS voltage and RMS current, $\omega_{line} = 2\pi f_{line}$, and f_{line} is the line frequency.

Fig. 3 shows the theoretical current waveforms in the positive half-line cycle. The input inductor L_{in} works under continuous conducting mode with a small current ripple. When S_1 is ON, and S_3 is OFF, the switch current, i_{S1} is the sum of i_{Lin} and i_{L1} . When S_1 is OFF, and S_3 is ON, the output current i_{S3} is the sum of i_{Lin} and i_{L1} . A typical waveform of the inductor current in each switching cycle is shown in Fig. 4. D_{ON} is the turn-on duty cycle for S_1 . According to the operation principle, the instantaneous current of S_1 and S_3 in each switching cycle can be expressed as

$$i_{S1}(t) = \begin{cases} i_{Lin}(t) + i_{L1}(t), & 0 < t < D_{ON} \cdot T_s \\ 0, & D_{ON} \cdot T_s < t < T_s \end{cases} \quad (3)$$

$$i_{S3}(t) = \begin{cases} 0, & 0 < t < D_{ON} \cdot T_s \\ i_{Lin}(t) + i_{L1}(t), & D_{ON} \cdot T_s < t < T_s \end{cases} \quad (4)$$

When S_1 turns OFF, and S_3 turns ON, the average current of S_3 in each switching cycle can be obtained as

$$i_{S3_avg}(t) = \frac{i_{S3_pk} \cdot T_{OFF}}{2 \cdot T_s} = \frac{(i_{Lin_pk} + i_{L1_pk}) \cdot T_{ON} \cdot V_{in}(t)}{2 \cdot T_s \cdot V_o} \quad (5)$$

Then the average of S_3 in the half-line cycle is derived

$$I_{S3_avg} = \frac{1}{\pi} \cdot \int_0^\pi i_{S3_avg}(t) \cdot dt = \frac{V_{in_rms}^2 \cdot (L_1 + L_{in}) \cdot T_{ON}^2}{2 \cdot L_1 \cdot L_{in} \cdot T_s \cdot V_o} \quad (6)$$

Substituting $I_{S3_avg} = P_o / V_o$, based on (1) to (6), T_{ON} and T_{OFF} yields

$$T_{ON} = \frac{V_o}{V_{in_rms}} \cdot \sqrt{\frac{2 \cdot L_1 \cdot L_{in} \cdot T_s}{(L_1 + L_{in}) \cdot R_o}} \quad (7)$$

$$T_{OFF} = \frac{V_{in}(t)}{V_o} \cdot T_{ON} = \frac{V_{in}(t)}{V_{rms}} \cdot \sqrt{\frac{2 \cdot L_1 \cdot L_{in} \cdot T_s}{(L_1 + L_{in}) \cdot R_o}} \quad (8)$$

Since $T_{ON} + T_{OFF} = T_s = 1/f_s$ under CRM mode, hence the switching frequency is

$$f_s(t) = \frac{(L_{in} + L_1) \cdot V_o^2 \cdot V_{in_rms}^2}{2 \cdot P_o \cdot L_{in} \cdot L_1 \cdot (V_{in}(t) + V_o)^2} \quad (9)$$

In Fig. 5, the red line shows the theoretical analysis variable switching frequency for CRM mode of the bridgeless Dual-SEPIC PFC during the half-line cycle. It can be seen that the

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maximum switching frequency occurs in the crossing section of the line cycle, and the minimum switching frequency occurs in the middle of the half-line cycle when the input voltage reaches the peak value.

To achieve ZVS for the main switch S_1 , the negative current of i_{S1} is needed to discharge the output capacitors of S_1 . The conventional SEPIC PFC can not provide the discharged negative current. In order to obtain the negative current, the output diode D_o is replaced by the MOSFET S_3 . Then S_3 turns OFF when i_{S3} declined to the required negative current rather than zero, i.e., $i_{S3} = i_{Lin} + i_{L1} \leq 0$; after S_3 is OFF, the negative current will transfer to i_{S1} to realize ZVS for S_1 .

To limit the high switching frequency at the crossing section of the AC line cycle, the maximum switching frequency is limited to 1 MHz. Moreover, the practical switching frequency is shown as the blue dash line. The practical switching frequency is lower than the theoretical switching frequency because the extended conduction time is added to the synchronous switch S_3 .

Assuming the intermediate capacitors C_1 and C_2 are large enough, the voltage ripples of C_1 and C_2 in each switching cycle are negligible. And the voltage ripple of the AC input voltage $V_{in}(t)$ also can be seemed as a constant value since the switching frequency is much higher than the line frequency, i.e., $V_{C1} = V_{in}(t) = V_{in}(t_0)$. The value is selected at the beginning of each switching cycle. Based on the aforementioned assumptions, the circuit operation can be divided into six stages, separated by time instants ($t_0, t_1 \dots t_6$), as marked in Fig. 4. The equivalent circuit of each stage is given in Fig. 6, and the corresponding analytical model is derived.

Fig. 3. The current waveform during the positive half-line cycle.

Fig. 4. Typical current waveform in each switching cycle.

Fig. 5. Variable switching frequency during half-line cycle.

1) Stage I (t_0-t_1):

As shown in Fig. 4, this stage starts when the switch current i_{S1} crosses zero. The equivalent circuit is shown in Fig. 6(a), L_{in} is charged by the input voltage, and L_1 is charged by the intermediate capacitor C_1 . The voltages of both inductors L_1 and L_{in} equal to $V_{in}(t_0)$. Hence, the inductor current i_{Lin} and i_{L1} starts to rise linearly. Based on the operation mentioned above, the following equations can be obtained:

$$\begin{cases} i_{Lin}(t) = I_{in}(t_0) + \frac{V_{in}(t_0)}{L_{in}}(t - t_0) \\ i_{L1}(t) = -I_{in}(t_0) + \frac{V_{in}(t_0)}{L_1}(t - t_0) \\ i_{S1}(t_1) = i_{L1}(t) + i_{Lin}(t) = \frac{V_{in}(t_0)}{L_{eq}}(t - t_0) \geq 0 \\ i_{S3}(t) = 0 \end{cases} \quad (10)$$

Where $I_{in}(t_0)$ is the average input current, L_{eq} is the equivalent inductor $L_{eq} = L_1 \cdot L_{in} / (L_1 + L_{in})$.

Fig. 6. Six operation stages for the modified SEPIC PFC. (a) Stage I. (b) Stage II. (c) Stage III. (d) Stage IV. (e) Stage V. (f) Stage VI.

2) Stage II (t_1-t_2):

Stage II starts at t_1 when S_1 is OFF, as shown in Fig. 6(b), S_3 is also in the OFF state and ready to turn ON. The two parallel GaNFETs output capacitors C_{oss1} and C_{oss3} , begin to resonate with the two parallel inductors L_{in} and L_1 . During this resonant period, C_{oss1} is charged while C_{oss3} is discharged simultaneously. This stage ends when the voltage of C_{oss1} is charged from 0 to $(V_{in}+V_o)$, and the voltage of C_{oss3} is released from $(V_{in}+V_o)$ to 0. In this short resonant transition period, the currents of i_{Lin} and i_{L1} can be seemed as constant.

3) Stage III (t_2-t_3):

This stage begins by turning S_3 ON at t_2 . Since the C_{oss3} of S_3 is already discharged to zero, thus ZVS of S_3 is realized. The input inductor current i_{Lin} and the output inductor current i_{L1} flow through S_3 to the load, and both inductor voltages are $-V_o$. This stage ends when the current of i_{S3} drops to zero at t_3 . The equations of the currents can be obtained:

$$\begin{cases} i_{Lin}(t) = i_{Lin}(t_2) - \frac{V_o}{L_{in}}(t - t_2) \\ i_{L1}(t) = i_{L1}(t_2) - \frac{V_o}{L_1}(t - t_2) \\ i_{S1}(t) = 0 \\ i_{S3}(t) = i_{L1}(t_2) + i_{Lin}(t_2) - \frac{V_o(t_0)}{L_{eq}}(t - t_2) > 0 \end{cases} \quad (11)$$

4) Stage IV (t_3-t_4):

For the conventional SEPIC PFC shown in Fig. 1(a) to Fig. 1(c), the output diode D_o turns OFF automatically when the diode current i_{D_o} declines to zero. However, the main switch S_3 of the modified Dual-SEPIC PFC remains ON after the switch current i_{S3} reaches zero. i_{S3} continues to decrease by adding an extended conduction time T_{ex} as shown in Fig. 4 from t_3 to t_4 , here $T_{ex} = t_4 - t_3$. Thus, after S_3 is OFF, the negative current is transferred to S_1 . Hence, the negative current is provided to discharge C_{oss1} to achieve ZVS for S_1 . As shown in Fig. 6(d), the equivalent circuit is the same as Stage III, except for the polarity of current i_{S3} . This stage ends when i_{S3} drops to the required negative current value I_{S3_val} for ZVS realization.

$$i_{S3}(t) = i_{L1}(t_2) + i_{Lin}(t_2) - \frac{V_o(t_0)}{L_{eq}}(t - t_2) < 0 \quad (12)$$

Thus, I_{S3_val} can be expressed as

$$I_{S3_val} = -\frac{V_o(t_0)}{L_{eq}}T_{ex} < 0 \quad (13)$$

5) Stage V (t_4-t_5):

This stage is also a resonated stage, which is similar to Stage II. It starts when S_3 is OFF at t_4 . Then the negative current in S_3 is transferred to S_1 . The two parallel GaNFETs output capacitors C_{oss1} and C_{oss3} start to resonate with the two parallel inductors. During this resonant period, the voltage of C_{oss1} is discharging from $(V_{in}+V_o)$ to 0, and the voltage of the C_{oss3} is discharged from 0 to $(V_{in}+V_o)$, as shown in Fig. 6(e). This stage ends when S_1 turns on. Here assuming that $C_{oss1} =$

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$C_{oss3} = C_{oss}$, the input voltage is considered as constant value as $V_{in} = V_{in}(t_0)$. The voltages of C_{oss1} and C_{oss3} are $V_{ds1}(t)$ and $V_{ds3}(t)$. The inductor current i_{S3} and the voltages $V_{ds1}(t)$, $V_{ds3}(t)$ can be expressed by the equations below:

$$\begin{cases} i_{S3}(t) = I_{S3} \cdot \cos(w_0 t) - \frac{V_o}{Z_n} \sin(w_0 t) \\ V_{ds1}(t) = V_{in} + V_o \cdot \cos(w_0 t) + I_{S3} \cdot Z_n \sin(w_0 t) \\ V_{ds2}(t) = V_o - V_o \cdot \cos(w_0 t) - I_{S3} \cdot Z_n \sin(w_0 t) \\ w_0 = 1 / \sqrt{L_{eq} C_{eq}}, Z_n = \sqrt{L_{eq} / C_{eq}}, C_{eq} = C_{oss1} + C_{oss3} \end{cases} \quad (14)$$

Where Z_n is the characteristic impedance in the resonant period. C_{eq} is the equivalent capacitor. w_0 is the resonant angular frequency. $I_{S3} = I_{L1} + I_{Lin}$, where I_{Lin} and I_{L1} are the inductor current value at the beginning of the resonant stage. Based on (14), then the following equation can be obtained:

$$Z_n^2 \cdot i_{S3}(t)^2 + (V_{ds1}(t) - V_{in})^2 = I_{S3}^2 \cdot Z_n^2 + V_o^2 \quad (15)$$

It shows that the trajectory of the drain-source voltage V_{ds1} at this resonant stage is a circle, with the center at $(0, V_{in})$, and the radius is $\sqrt{I_{S3}^2 \cdot Z_n^2 + V_o^2}$, as shown in Fig. 7(c). For the conventional DCM SEPIC PFC with the diode, the $I_{D0} = I_{Lin} + I_{L1} = 0$. Therefore the radius is V_o when the input voltage exceeds the output voltage ($V_{in} > V_o$). With a similar analysis, it also can be obtained that the radius is V_{in} when $V_{in} < V_o$ for the conventional SEPIC PFC. The state plane trajectories for the conventional DCM SEPIC PFC are also shown in Fig. 7(a) and Fig. 7(b), respectively.

6) Stage VI ($t_5 - t_6$):

This stage starts by turning S_1 ON at t_5 . Since the voltage of output capacitor C_{oss1} is already discharged to 0 in Stage V,

hence the ZVS operation of S_1 is obtained. Both the two inductor currents flow through S_1 , and the voltages across the two inductors are V_{in} . This stage ends when the current of i_{S1} rises to zero. Then another switching cycle starts as Stage I.

B. Conditions for ZVS implementation

Compared with the modified Dual-SEPIC PFC, the conventional DCM SEPIC PFC with the diode rectifier only has four stages because no negative output current existed. When the diode current i_{D0} is decreased to zero, as shown in Fig. 7(d) ($V_{in} < V_o$) and Fig. 7(e) ($V_{in} > V_o$), the two parallel inductors start to resonate with the two parallel capacitors, including the output capacitor of S_1 and the parasitic capacitors of the diode D_o . The valley switching can be applied with the correct turn-on instant time control. Thus, the minimum value of V_{ds1} is achieved at the valley switching is $(V_{in} - V_o)$. Hence the ZVS for the conventional SEPIC PFC can only be achieved when the input voltage V_{in} is lower than the output voltage V_o as shown in Fig. 7(d). The state plane trajectories for conventional DCM SEPIC PFC shown in Fig. 7(a) ($V_{in} < V_o$) and Fig. 7(b) ($V_{in} > V_o$) also indicate that the ZVS operation can only be achieved when $V_{in} < V_o$. As shown in Fig. 7(e), when the input voltage exceeds the output voltage, the minimum value of drain-source voltage V_{ds1} can only resonate to a valley point $(V_{in} - V_o)$. Therefore, at least $0.5C_{eq}(V_{in} - V_o)^2$ power loss occurs when S_1 starts turning on.

In order to achieve the fully ZVS in the whole line cycle, the diode D_o is replaced by the high-speed GaN device S_3 , as shown in Fig. 1(d). And then the ZVS extension strategy explained in [6],[8],[39]–[41] is used. Then the negative current can be provided to guarantee the ZVS for the main

Fig. 7. (a). State plane trajectory for conventional SEPIC PFC when $V_{in} \leq 0.5V_o$. (b). State plane trajectory for conventional SEPIC PFC when $V_{in} \geq 0.5V_o$. (c). State plane trajectory for the modified SEPIC PFC. (d). Operation waveforms for conventional SEPIC PFC when $V_{in} \leq 0.5V_o$. (e). Operation waveforms for conventional SEPIC PFC when $V_{in} \geq 0.5V_o$. (f). Operation waveforms for the modified SEPIC PFC.

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switch S_1 . Thus, instead of automatically turning OFF the diode D_O in DCM mode when the diode current decreases to zero, a short extended conduction time T_{ex} is added on purpose for S_3 shown in Fig. 7(f) *Stage IV* (t_3-t_4). Then the negative current flows through S_1 is provided to discharge the output capacitor C_{oss1} to help to achieve ZVS for the main switch S_1 . It also can be seen from the state plane trajectory in Fig. 7(c), the negative current i_{S3} enlarges the radius from V_o to $\sqrt{I_{S3}^2 \cdot Z_n^2 + V_o^2}$. Hence the voltage V_{ds1} can drop to zero. Based on the state plane trajectory as shown in Fig. 7(c) and (14), the minimum required negative current i_{S3_val} , and the extended conduction time T_{ex} to achieve ZVS operation for S_1 in the half-line cycle is calculated as (16) and (17).

$$i_{S3_val}(t) = -\sqrt{\frac{V_{in}(t)^2 - V_o^2}{Z_n^2}} \quad (16)$$

$$T_{ex}(t) = \frac{|i_{S3_val}(t)| \cdot L_{eq}}{V_o} \quad (17)$$

Finally, the theoretical analysis for the required negative current i_{S3_val} and the required extended conduction time T_{ex} are plotted in Fig. 8. Here the input voltage V_{in} is 220 Vac, and the output voltage is 100 Vdc.

Fig. 8. Required negative current and required extended conduction time for S_3 in the positive half-line cycle.

III. INTEGRATED INDUCTORS DESIGN

The specification of the bridgeless Dual-SEPIC PFC is list in the Table II:

TABLE II SPECIFICATION OF BRIDGELESS DUAL-SEPIC PFC	
Circuit parameters	Value
Input voltage	220 Vac
Nominal Output voltage	100 Vdc
Input current ripple	20% * I_{in}
Output voltage ripple	10% * V_o
Power rating	300 W

A. Inductance design for the SEPIC PFC

There are three constraints for the inductance design of the integrated inductors:

1. The ZVS turn-on in the full range of the input voltage;
2. Input inductor current ripple limitation;
3. The output inductor current ripple limitation.

For the bridgeless dual-SEPIC PFC converter, based on (7) and (8), the duty cycle can be derived:

$$\begin{cases} D_{ON} = \frac{V_o}{V_{in_rms}} \sqrt{\frac{2L_1 \cdot L_{in}}{(L_1 + L_{in})R_0 T_s}} \\ D_{OFF} = \sqrt{\frac{2L_1 \cdot L_{in}}{(L_1 + L_{in})R_0 T_s}} \end{cases} \quad (18)$$

As mentioned before, D_{ON} and D_{OFF} are the turn-on duty cycle and turn-off duty cycle for the main switches S_1 and S_2 . To make sure the output current i_{S3} can drop below zero, the sum of D_{ON} and D_{OFF} should always small than 1. Thus, the first constraint can be obtained:

$$D_{ON} + D_{OFF} \leq 1 \Rightarrow \frac{2L_1 \cdot L_{in}}{(L_1 + L_{in})R_0 T_s} \leq \frac{(\sqrt{2}V_{in_rms})^2}{(\sqrt{2}V_{in_rms} + V_o)^2} \quad (19)$$

The relationship between L_{in} and L_1 for the inequality is shown in Fig. 9 as the red line. It can be seen that if the equivalent inductance L_{in} becomes large to obtain a smaller input current ripple, then the equivalent inductance of L_1 should be limited to a smaller value. This limitation comes from the output current $i_{S3} = i_{Lin} + i_{L1} \leq 0$ when S_3 turns OFF, the more negative current is needed for i_{L3} since the input current ripple becomes small.

Denoting K_I is the peak input current ripple Δi_{Lin_pk} over the peak input current I_{in_pek} , based on (7), since $\Delta i_{Lin_pk} = V_{in}(t)T_{on}/L_{in}$, then K_I can be deduced by the following equation

$$K_I = \frac{\Delta i_{Lin_pk}}{\sqrt{2} \cdot I_{in_rms}} = \frac{1}{L_{in}} \sqrt{\frac{2L_1 \cdot L_{in} \cdot T_s}{(L_1 + L_{in})P_o}} \sqrt{2}V_{in_rms} \leq 0.2 \quad (20)$$

K_I is determined by the output power, the inductance values, and the switching frequency. According to Table II, here, $K_I \leq 0.2$ is selected, and this inequality is shown in Fig. 9 as the blue line.

Fig. 9. Inductance design by three constraints.

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As mention before, the voltages crossed the input inductor L_{in} and output side inductor L_I are always the same in each operation stage. Hence, the ratio of the output side inductor ripple Δi_{L_I} and the input inductor current ripple $\Delta i_{L_{in}}$ equals to the ratio of the inductance L_{in} to L_I shown in (21). To limit the ripple current of L_I , the inductance ratio K_2 between L_{in} and L_I should also keep as low as possible. K_2 is selected smaller than 35, and this inequality is shown in Fig. 9 as the black line.

$$\frac{\Delta i_{L_I}}{\Delta i_{L_{in}}} = \frac{L_{in}}{L_I} \leq K_2 = 35 \quad (21)$$

Based on the three constraints aforementioned in (19) ~ (21), the available values for L_I and L_{in} can be selected from Fig. 9 in the yellow area as bellows:

$$\begin{cases} L_{in_eq} = 120 \sim 190 \mu\text{H} \\ L_{1_eq} = 4 \sim 6 \mu\text{H} \\ L_{2_eq} = 4 \sim 6 \mu\text{H} \end{cases} \quad (22)$$

B. E-I-E core with a specified coupling coefficient

For general three coupled inductors, the relationships among the voltage of each winding, the self-inductance, and the mutual inductance are given by the equations below, respectively:

$$\begin{bmatrix} V_{L_{in}} \\ V_{L_I} \\ V_{L_2} \end{bmatrix} = \begin{bmatrix} L_{in_self}, \pm M_{in_1}, \pm M_{in_2} \\ L_{1_self}, \pm M_{1_in}, \pm M_{1_2} \\ L_{2_self}, \pm M_{2_in}, \pm M_{2_1} \end{bmatrix} \begin{bmatrix} \frac{di_{L_{in}}}{dt} \\ \frac{di_{L_I}}{dt} \\ \frac{di_{L_2}}{dt} \end{bmatrix} \quad (23)$$

The inductors can be coupled directly or inversely. The sign before M is "+" or "-" depended on the inductors are coupled directly or inversely. $V_{L_{in}}$, V_{L_I} , and V_{L_2} are the voltages across the three inductors, respectively. L_{in_self} , L_{1_self} , and L_{2_self} are the self-inductance value of the windings. M is the mutual inductance between inductors, where $M_{1_in}=M_{in_1}$, $M_{2_in}=M_{in_2}$, $M_{1_2}=M_{2_1}$.

For the three inductors in the proposed topology, the output inductors L_I only works in the positive half-line cycle, and L_2 only works in the negative half-line cycle. Therefore, L_I and L_2 are directly coupled with the input inductor L_{in} separately. L_I and L_2 need to be always decoupled from each other. In this paper, the symmetric E-I-E core geometry is selected. The magnetics structure is shown in Fig. 10. The output side inductors L_I and L_2 are wound on the center legs of the E-cores individually. The input inductor L_{in} is wound in the middle I-core with two identical parts in series, where each part has the same number of turns. Due to the low reluctance path provided by the middle I-core, the flux induced by L_I and L_2 will return from the middle I-core rather than the E-core. Thus, L_I can be decoupled with L_2 by the middle I-core.

Fig. 10. Proposed E-I-E core structure for the coupled inductor.

Since L_I is always decoupled with L_2 , M_{1_2} is equal to 0. Taking the positive half-line cycle as an example, the identical voltages are applied to L_{in} and L_I at all times throughout the switching period, which means $V_{L_{in}}$ is equal to V_{L_I} . Since L_2 does not work at the positive line cycle, the inductor voltage is zero, $V_{L_2}=V_{C2} \approx 0$. Then (23) can be rewritten

$$\begin{cases} V_{L_{in}} = L_{in_self} \frac{di_{L_{in}}}{dt} + M_{in_1} \frac{di_{L_I}}{dt} - M_{in_2} \frac{di_{L_2}}{dt} \\ V_{L_I} = L_{1_self} \frac{di_{L_I}}{dt} + M_{1_in} \frac{di_{L_{in}}}{dt} \\ 0 = L_{2_self} \frac{di_{L_2}}{dt} - M_{2_in} \frac{di_{L_{in}}}{dt} \end{cases} \quad (24)$$

Assuming $M_{in_1}=M_{in_2}=M$, $L_{1_self}=L_{2_self}$, the equivalent inductance of each inductor can be obtained

$$\begin{cases} L_{in_eq} = \frac{L_{in_self} L_{in_self} - 2M^2}{L_{1_self} - M} = \frac{(1-2k_c^2)L_{in_self}}{1-k_c \sqrt{L_{in_self} / L_{1_self}}} \\ L_{1_eq} = \frac{L_{1_self} L_{in_self} - M^2}{L_{in_self} - M} = \frac{(1-k_c^2)L_{1_self} L_{in_self}}{L_{in_self} - k_c \sqrt{L_{1_self} L_{in_self}}} \\ L_{2_eq} = \frac{L_{2_self} L_{in_self} - M^2}{L_{in_self} - M} = \frac{(1-k_c^2)L_{2_self} L_{in_self}}{L_{in_self} - k_c \sqrt{L_{2_self} L_{in_self}}} \\ M = k_c \sqrt{L_{in_self} L_{1_self}} \end{cases} \quad (25)$$

Based on (25), it shows that the equivalent inductance of the input inductor L_{in_eq} can be enlarged several times over the self-inductance L_{in_self} when the self-inductance L_{1_self} and L_{2_self} are very close to the mutual inductance M . The required equivalent inductance of the input inductor L_{in_eq} can be achieved by much smaller self-inductance L_{in_self} by appropriate coupling coefficient (k_c).

The magnetic flux flows in the proposed E-I-E core in the positive half-line cycle are shown in Fig. 11. Here i_{L_2} is almost zero, only L_{in} and L_I coupled with each other. N_{in} , N_I , N_2 are the number of turns for L_{in} , L_I , and L_2 , respectively. Φ_1 , Φ_2 , Φ_3 , and Φ_4 are the flux flow in four outer legs of the E-I-E core geometry, respectively. Fig. 11(a) shows an equivalent magnetic reluctance model of the proposed E-I-E core module.

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Where R_{l1} and R_{l2} are the reluctance of E-core, R_{lin} is the reluctance of I-core, here $R_{l1}=R_{lin}$. R_{g1} and R_{g2} represent the reluctance of the air gap between the I-core, and two E-cores, The air-gap of g_1 is also identical to g_2 , i.e., $R_g=R_{g1}=R_{g2}$. By defining $R_l=R_{l1}+R_{l2}+R_g$, $R_2=(R_{l2}+R_g)/2$, the magnetic reluctance model can be simplified to Fig. 11(b).

Fig. 11. The magnetic reluctance model of the E-I-E core. (a) The original model. (b) The simplified model.

Based on the equivalent magnetic reluctance model, four reluctance equations in each sub-loop can be obtained:

$$\begin{cases} \frac{N_{in}i_{Lin}}{2} + N_l i_{L1} = \Phi_1 \cdot R_l + (\Phi_1 + \Phi_2) \cdot R_2 + (\Phi_1 + \Phi_3)R_{l1} \\ \frac{N_{in}i_{Lin}}{2} + N_2 i_{L2} = \Phi_2 \cdot R_l + (\Phi_1 + \Phi_2) \cdot R_2 + (\Phi_2 + \Phi_4)R_{l1} \\ \frac{N_{in}i_{Lin}}{2} = \Phi_3 \cdot R_l + (\Phi_3 + \Phi_4) \cdot R_2 + (\Phi_1 + \Phi_3)R_{l1} \\ \frac{N_{in}i_{Lin}}{2} = \Phi_4 \cdot R_l + (\Phi_3 + \Phi_4) \cdot R_2 + (\Phi_2 + \Phi_4)R_{l1} \end{cases} \quad (26)$$

Since the flux Φ_1 is identical to Φ_2 , and Φ_3 is identical to Φ_4 because of the symmetrical core structure and windings, then (26) can be simplified into two equations:

$$\begin{cases} \frac{N_{in}i_{Lin}}{2} + N_l i_{L1} = \Phi_1 \cdot R_l + \Phi_1 \cdot 2 \cdot R_2 + (\Phi_1 + \Phi_3)R_{l1} \\ \frac{N_{in}i_{Lin}}{2} = \Phi_3 \cdot R_l + \Phi_3 \cdot 2 \cdot R_2 + (\Phi_1 + \Phi_3)R_{l1} \end{cases} \quad (27)$$

Solving (27), then the flux in each loop is obtained:

$$\begin{cases} \Phi_1 = \Phi_2 = \frac{(R_l + 2R_2)N_{in}i_{Lin} + 4R_{l1}N_l i_{L1}}{2(R_l + 2R_2)(2R_l + R_{l1})} \\ \Phi_3 = \Phi_4 = \frac{(R_l + 2R_2)N_{in}i_{Lin} - 2R_{l1}N_l i_{L1}}{2(R_l + 2R_2)(2R_l + R_{l1})} \end{cases} \quad (28)$$

The total flux flows through the winding L_l is $\Phi_1 + \Phi_2 = 2\Phi_1$, and the total flux flows through L_{in} is $\Phi_1 + \Phi_2 + \Phi_3 + \Phi_4 = 2(\Phi_1 + \Phi_3)$. Taking the partial derivative of the total flux in each winding:

$$\begin{cases} V_{L1} = N_l \cdot \frac{d(\Phi_1 + \Phi_2)}{dt} = \frac{4N_l^2 \cdot R_l}{\Delta} \cdot \frac{di_{L1}}{dt} + \frac{N_{in} \cdot N_l}{2R_l + R_{l1}} \cdot \frac{di_{Lin}}{dt} \\ V_{Lin} = N_{in} \cdot \frac{d(\Phi_1 + \Phi_3)}{dt} = \frac{N_{in}^2}{2R_l + R_{l1}} \cdot \frac{di_{Lin}}{dt} + \frac{N_{in} \cdot N_l}{2R_l + R_{l1}} \cdot \frac{di_{L1}}{dt} \end{cases} \quad (29)$$

The equation (29) can be rewritten

$$\begin{bmatrix} V_{L1} \\ V_{Lin} \end{bmatrix} = \begin{bmatrix} \frac{4N_l^2 \cdot R_l}{\Delta} & \frac{N_{in} \cdot N_l}{2R_l + R_{l1}} \\ \frac{N_{in} \cdot N_l}{2R_l + R_{l1}} & \frac{N_{in}^2}{2R_l + R_{l1}} \end{bmatrix} \begin{bmatrix} \frac{di_{L1}}{dt} \\ \frac{di_{Lin}}{dt} \end{bmatrix} = \begin{bmatrix} L_{1_self} & M \\ M & L_{in_self} \end{bmatrix} \begin{bmatrix} \frac{di_{L1}}{dt} \\ \frac{di_{Lin}}{dt} \end{bmatrix} \quad (30)$$

Here $\Delta = (2R_l + R_{l1})(R_l + 2R_2)$, then self-inductances and mutual inductances can be obtained:

$$\begin{cases} L_{1_self} = \frac{4N_l^2 \cdot R_l}{(2R_l + R_{l1})(R_l + 2R_2)} \\ L_{in_self} = \frac{N_{in}^2}{2R_l + R_{l1}} \\ M = \frac{N_{in} \cdot N_l}{2R_l + R_{l1}} \end{cases} \quad (31)$$

Finally, the theoretical coupling coefficient can be derived:

$$k_c = \frac{M}{\sqrt{L_{1_self} \cdot L_{in_self}}} = \sqrt{\frac{R_l + 2R_2}{4R_l}} \approx \sqrt{\frac{1}{2}} = 0.707 \quad (32)$$

Here it can be seen that the coupling coefficient is related to the core structure, the air-gap, and the material of the magnetic core. R_g is much bigger than R_{l1} and R_{l2} due to the air-gaps between the E-core and I-core. Hence, the theoretically calculated value of the coupling coefficient is around 0.7. It has also been verified in the 2D finite-element analysis (FEA) simulation.

IV. HARDWARE IMPLEMENTATION

A 300 W GaN-based MHz bridgeless Dual-SEPIC PFC prototype is built to verify the ZVS control method with the integrated E-I-E inductors. The integrated inductor design for the Dual-SEPIC PFC, and control scheme are described as follows.

A. Integrated inductor design

Ferrite material with 3F36 is selected since it is suitable for high frequency from 300 kHz to 1 MHz. The detailed parameters of the integrated E-I-E core are listed in Table III.

TABLE III
PARAMETERS OF E-I-E CORE

Circuit parameters	Value
Ferrite core material	3F36 (300 kHz to 1 MHz)
Initial permeability (μ_0)	1600
E-core size	E43/10/28
I-core size	PLT43/28/4.1

Defining K_{Lin} , and K_{Ll} are the ratios of the equivalent inductance over the self-inductance for the input inductor L_{in} and the output side inductor L_l , respectively. According to (25) and (31), K_{Lin} , and K_{Ll} can be obtained

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$$K_{L1} = \frac{L_{1_eq}}{L_{1_self}} = \frac{1-k_c^2}{1-k_c\sqrt{L_{1_self}/L_{in_self}}} = \frac{1-k_c^2}{1-\sqrt{2}k_c\frac{N_1}{N_{in}}} \quad (33)$$

$$K_{Lin} = \frac{L_{in_eq}}{L_{in_self}} = \frac{1-2k_c^2}{1-k_c\sqrt{L_{in_self}/L_{1_self}}} = \frac{1-2k_c^2}{1-\frac{1}{\sqrt{2}}k_c\frac{N_{in}}{N_1}} \quad (34)$$

It can be seen that K_{Lin} and K_{L1} are determined by the number of winding turns and coupling coefficient k_c . According to the equivalent inductance requirement based on (22) and the flux saturation limitation, here, $N_{in} = 12$, $N_1 = N_2 = 5$ are selected. The relationship between K_{Lin} , K_{L1} with the coupling coefficient k_c is shown in Fig. 12. Fig. 12(a) shows that the equivalent inductance L_{in_eq} can be enlarged several times to self-inductance L_{in_self} . On the other side, the ratio K_{L1} for L_1 is around 1, as shown in Fig. 12(b), meaning that the self-inductance L_{1_self} is almost equal to the inductance L_{1_eq} .

As shown in Fig. 12(a), when the coupling coefficient k_c is large than 0.59, the value of equivalent inductance L_{in_eq} is negative, which means the rate of the flux change and the rate of the current change, are negative as shown in Fig. 13 in the dashed line. To obtain the same negative current I_{S3_val} when S_3 is OFF to achieve ZVS for S_1 , more current ripples will be introduced for L_1 when $k_c > 0.59$. Therefore, to minimize the inductor ripple, the coupling coefficient should be designed to less than 0.59.

Fig. 12. Relationship between k_c and ratios of equivalent-inductance over self-inductance. (a) K_{Lin} (b) K_{L1}

Fig. 13. Current waveform when k_c is large than 0.59.

Since the self-inductance L_{in_self} is around 15 μH and L_{1_self} is around 5 μH , to satisfy the equivalent inductance requirement (22), the ratio $K_{Lin} = L_{in_eq} / L_{in_self}$ is designed between 8 to 10. Thus, according to Fig. 14, the coupling coefficient k_c ranges from 0.562 to 0.568.

Fig. 14. Relationship between k_c and K_{Lin}

To verify the equivalent magnetic reluctance model and select the air-gap for the E-I-E integrated inductors, 2D finite-element analysis (FEA) simulation has been conducted. As shown in Fig. 15(a), the flux is symmetrical in the E-I-E core. Fig. 15(b) shows the integrated inductors with four windings for the E-I-E core. The practical value of the coupling coefficient has been obtained based on the decoupled T model of the coupled inductor for the proposed E-I-E core geometry.

Finally, the relationship between the air-gap and coupling coefficient both for FEA simulation and practical test are shown in Fig. 16. It can be seen that the coupling coefficient k_c is decreased when the length of the air-gap increases. Both of them are smaller than the theoretical analysis value 0.7 due to the leakage flux and fringing effect [41][42].

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Fig. 15. The proposed E-I-E core. (a) Flux distribution for 2D-FAE simulation. (b) the practical E-I-E core structure.

Fig. 16. Coupling coefficient value with different air-gap under 2D FEA simulation and practical test.

According to the aforementioned analysis, the air-gap $g_1=g_2=1.3$ mm is selected for the integrated E-I-E core inductors. According to $L_{eq}=V_{in}T_{on}/\Delta I_L$, the practice equivalent inductance can be obtained by the testing inductor current waveform. The self-inductances are directly obtained by the precision impedance analyzer Agilent 4294A. Finally, Table IV illustrates the actual inductance of the E-I-E integrated inductors.

Fig.17 shows the size comparison between the proposed E-I-E integrated inductors and the conventional discrete inductors. Here it shows that the integrated inductors can make the magnetic components more compact and reduce the volume of the inductor size.

TABLE IV
DESIGN INDUCTANCE OF THE INTEGRATED INDUCTOR

E-I-E core parameters		Value (μH)
Self-inductance	L_{in_self}	14.14
	L_{1_self}	5.18
	L_{2_self}	5.30
Equivalent inductance	L_{in_eq}	140.27
	L_{1_eq}	5.81
	L_{2_eq}	5.63
Mutual inductance	M_{1_in}	5.6
	M_{2_in}	5.4

Fig. 17. Size comparison between the integrated inductors and three isolated inductors

B. Control scheme

Fig. 18 shows the simplified control diagram with the MCU-based digital controller for the bridgeless Dual-SEPIC PFC. The voltage-mode control method is selected since it does not require a high-speed current sensor and is easy to implement than the current-mode control. To attenuate the double line voltage ripple on the DC bus, the bandwidth of the voltage loop is designed around 10 Hz. The difference between the output voltage reference V_{o_ref} and the real output voltage V_{o_real} is sent to the voltage PI controller. The PI controller determines the ON-time (T_{on}) of the main switches S_1 and S_2 in different half-line cycles.

The turn-off instant for the synchronous switches S_3 and S_4 are triggered by the zero current detection (ZCD) signal. The ZCD signal is obtained by the shunt resistor R_{ZCD} placed before the output capacitor. After the ZCD signal is sent to the digital controller, the extended conduction time T_{ex} for S_3 and S_4 is added based on (16) and (17) to achieve the required negative current i_{S3_val} . The gate driver signals for S_2 and S_4 in the positive half-line cycle are determined by the input voltage and threshold voltage V_{th} .

Fig. 18. The simplified control diagram for the SEPIC PFC.

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V. EXPERIMENTAL SETUP AND MEASUREMENT

A. Experimental prototype

Fig. 19 shows the photo of the 300 W prototype converter. The size of this prototype is 14.7 cm × 5.3 cm × 2.8 cm; the parameters of the key components are listed in Table V. The intermediate capacitor is 940 nF with two film capacitors paralleled. The DC-link bulky capacitors are 1320 uF with four electrolytic capacitors in parallel. The film capacitor has been used for the intermediate capacitors C_1 and C_2 . But the size of the intermediate capacitors can further be reduced by using few high voltage ceramic capacitors in parallel (such as CKC33C224JCGACTU, C0G, 500 V, 0.22 uF) since the capacitance is not very high.

Fig. 19. Prototype of the proposed SEPIC PFC.

TABLE V
PARAMETERS OF KEY COMPONENTS

Circuit parameters	Value
$S_1 \sim S_4$	GS66508T, 650 V, 50 mΩ
Isolate driver	SI8271AB-IS
Digital isolator	ADUM1100
Optically Isolated Amplifiers	ACPL-C87
C_1/C_2 , Film Capacitor	MKP385E44740JKI2B0, 4*470 nF
Output electrolytic capacitor	UCY2C331MHD6, 4*330 uF
Shunt resistor for ZCD circuit	WSHM2818R0150FEA, 15 mΩ
Comparator for ZCD circuit	LMV7219
Switching frequency	400 kHz~1 MHz
Digital controller	TMS320F28379D, 200 MHz

B. Measurement

Fig. 20(a) and Fig. 20(b) show the two different situations for DCM conventional SEPIC PFC with the output diode. After the current of the diode i_{Do} decreases to zero, the resonance happened between V_{ds} and i_{Do} . In Fig. 20(a), when $V_{in} \leq V_o$, the ZVS operation for S_1 can be achieved since V_{ds1} can resonate to zero. However, when $V_{in} > V_o$, only partial ZVS turn-on can be achieved at the valley switching, as shown in Fig. 20(b). Fig. 20(c) shows the waveform of ZVS for the modified SEPIC PFC with an extended conduction time T_{ex} added for the synchronous GaNFETs S_3 . The negative current i_{S3} transfer to the main switch S_1 to help achieve ZVS for S_1 after the S_3 turns OFF. The drain-source voltage V_{ds1} is discharged to zero before the main switch S_1 turns ON, thus the ZVS is guaranteed in the full range of the input voltage.

Fig. 21 shows the waveform of the drain-source voltage V_{ds1} , V_{ds3} , and the gate driver signal in one switching cycle in the

positive half-line cycle. Here it can be seen that the ZVS for both the main switch and synchronous switch is achieved.

Fig. 20. waveform for ZVS turn-on and Non ZVS turn-on. (a) ZVS turn-on for conventional SEPIC PFC with synchronous diode when $V_{in} \leq V_o$. (b) Non ZVS turn-on for conventional SEPIC PFC with synchronous diode when $V_{in} > V_o$. (c) ZVS turn-on for the modified SEPIC PFC.

Fig. 21. The waveform of ZVS turn on for S_1 and S_3

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Fig. 22. waveform of inductor current and input voltage

Fig. 23. waveform of inductor current and input voltage

Fig. 22 shows the waveform of three inductor currents i_{L1} , i_{L2} , and the output current before the DC bulk capacitors i_o for the bridgeless Dual-SEPIC PFC. The output current i_o is equal to the sum of the three inductor currents in the whole line cycle. It can be seen that the input current i_{Lin} is working in CCM mode, and the input current ripple is much smaller than the output current ripple.

Fig. 23 shows the detailed waveform of the three inductor currents in different switching period with variable switching frequencies. The top two sub-figures show the switching cycle in the positive half-line cycle. Here it can be seen that the

output side inductor current i_{L2} is almost 0. The output current i_o is equal to i_{S3} , which is the sum of i_{Lin} and i_{L1} . Identically the bottom two sub-figures show the inductor currents in the negative half-line cycle. Here, the inductor current i_{L1} is almost 0, and i_o is equal to the sum of i_{Lin} and i_{L2} . The switching frequency of the SEPIC PFC ranges from 400 kHz in the middle half-line cycle to 1 MHz in the crossing section of the line-cycle.

Fig. 24 and Fig. 25 show the experimental results for the transient performance of the proposed integrated inductors. The output voltage reference is changed from 60 Vdc to 100 Vdc. Here it shows the integrated inductors work well under the transient change without any overshoot or undershoot.

Fig. 26 shows the testing waveform of the input current, input voltage, and DC-link voltage. The DC-link voltage is maintained at around 100 Vdc. Fig. 27 shows that the voltage of the intermediate capacitor V_{C1} and V_{C2} equal to the input voltage V_{in} in the positive half-line cycle and negative half-line cycle, respectively.

Fig.28 shows the input current THD of the bridgeless Dual-SEPIC PFC. Fig.29 shows the input current of each harmonic order for the bridgeless Dual-SEPIC PFC is under the limitation of the international standard EN61000-3-2 Class-A. Due to the negative current in the switching cycle and the no power transfer mode near the zero-crossing input voltage, some distorted input current has happened around the zero-crossing section of the input voltage [39],[44].

Fig. 24. dynamic response from $V_{o_ref}=60$ Vdc to 100 Vdc

Fig. 25. Three inductor current waveforms of the integrated inductors under dynamic response

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Fig. 26. waveform of input current, input voltage, and DC-link voltage

Fig. 27. the waveform of input voltage V_{in} , intermediate capacitors voltage V_{C1} , V_{C2} .

Fig. 28. Measured THD of the SEPIC PFC ($V_{in} = 220$ Vac)

The system efficiency (including the DC-link bulky capacitors) of the prototype from 10% to 100% of full load under the $V_{in} = 220$ Vac, $V_o = 100$ Vdc, is shown in Fig. 30. The peak efficiency is 97% under full load.

Fig. 29. Harmonics of Dual SEPIC PFC and Class-A equipment harmonic current limit standard

Fig. 30. Measured efficiency curve of the SEPIC PFC prototype

Table VI lists the comparison of the state-of-the-art SEPIC PFC with the modified SEPIC PFC. It can be seen that the

TABLE VI
Comparison of different SEPIC PFC

Topology	Power (W)	Work Mode	Switching frequency(kHz)	ZVS	Peak Efficiency	Input rectifier	Switch device information	Output side switches
Sepic PFC with diode bridge[13]	1000	DCM	100	No	<94%	1*Rectifier: GBPC3506W 600V, Vf: 1.1 V	2*Si: FCH041N60F 600V, 36 mΩ	1*Diode: VS-60APU04 400 V, Vf: 0.87 V
Interleaved Sepic PFC with diode bridge[11]	500	DCM	150	No	96%	1*Rectifier: unknown	2*Si: unknown	2 * Diode: unknown
Active Clamped Isolated Sepic PFC[10]	2000	DCM	200	Partial ZVS	96.56%	1*Rectifier: unknown	2*SiC: SCT3040KL 1200V, 40 mΩ	2*SiC Diode: 4D40120D 1200 V, Vf: 1.5 V
Bridgeless Sepic PFC[22]	60	DCM	50	No	96%	No	2*Si: FQA48N20 200V, 50 mΩ	1*Clamp Diode: RURP8100 1000 V, Vf:1.8 V
Bridgeless Sepic PFC with Power decoupling[21]	500	CCM	20	No	95.6%	No	2*Si: IPW60R099P6 600V, 99 mΩ	2*Diode: MUR840 400 V, Vf: 1.3 V
Bridgeless Sepic PFC with output diode[26]	760	CCM	20	No	92%	No	2*Diode: DSEI60-06A 600V, Vf:1.8 V	1*Diode: DSEI60-06A 600V, Vf:1.8 V
Improved Bridgeless SEPIC Converter[16]	100	DCM	65	No	94%	No	2*Si: IPP60R125C6 600V, 125 mΩ	2 * Diode: unknown
Proposed converter	300	CRM	400~1000	Yes	97%	No	2*Diode: SF20L60U 600 V,Vf: 3 V	2*Diode: SF20L60U 600 V, Vf: 3 V
							2*GaN: GS66508T 650V, 50 mΩ	2*GaN: GS66508T 650V, 50 mΩ

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GaN-based MHz Dual-SEPIC PFC developed by the author's group is the only one that can achieve the ZVS in the full range of the input voltage. It achieves the highest peak efficiency in a high switching frequency.

VI. CONCLUSION

This paper presents a bridgeless GaN-based Dual-SEPIC PFC with ZVS operation for the application of a step-down AC-DC converter. The input current is working in CCM mode, and the sum of the input current and output current is working in CRM mode. To achieve the ZVS operation for the main switch, the extra negative current is provided by adding the extended conduction time for the synchronous switch after the current of the synchronous switch decrease to zero. Moreover, The E-I-E integrated inductors have been proposed for the bridgeless Dual-SEPIC PFC. All the inductors, including one input inductor and two output inductors, have been integrated into the E-I-E core, reducing the total ferrite volume and making the converter more compact. The inductance design and magnetic modeling for the coupled inductor have been analyzed. With a carefully designed coupling coefficient, the equivalent input inductance can be implemented with a much smaller number of winding turns. A 300W GaN-based MHz bridgeless Dual-SEPIC PFC with the integrated inductors is developed to verify the theoretical analysis and design. The experimental results show that full-range ZVS is realized, and the peak efficiency is 97%.

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