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Impact on ZVS Operation by Splitting Inductance to Both Sides of Transformer for 1-MHz GaN Based DAB Converter

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Abstract—Zero-voltage-switching (ZVS) of dual active bridge (DAB) converter and its range extension have been studied intensively. However, this paper proves that the external inductance location, i.e. placing on which side of the transformer, does affect the ZVS operation due to transformers' parasitic capacitances, in particular at a high switching frequency of higher than 1 MHz. Thus, this paper, for the first time, gives the mathematical model of the resonant transition during the dead time and thereby analyzes partial hard switching under different modulation schemes with consideration of transformer's parasitic capacitances. From the analysis of ZVS operation, the solution of splitting the interfacing inductance and placing them on both sides of the transformer is proposed and its associated methodology of selecting these inductances is introduced. It extends the ZVS for all the switching devices, therefore, enhance the efficiency or move losses away from the critical semiconductor devices. Finally, the theoretical analysis is verified by the experimental results from a 1 MHz Gallium Nitride (GaN) based DAB converter prototype.

Keywords—dual active bridge, dc-dc converter, GaN, high frequency, parasitic capacitance, soft switching

I. INTRODUCTION

The Dual Active Bridge (DAB) converter, as shown in Fig. 1, was proposed in [1], and has attractive features in terms of low component stress, capability of bidirectional power flow, buck-boost operation, large power capacity, ease of realizing soft-switching, and modular and symmetrical structure [2][3]. In the last decade, much research of the DAB converter focuses on characterizing its zero-voltage-switching (ZVS) operation. Typically, when $V_1 = N \cdot V_2$ (V_1 and V_2 are the bus voltage, and N is the transformer turn ratio, as denoted in Fig. 1), the ac current has trapezoidal shape and ZVS operation can be ensured over an entire load range. In [4]–[6], the ZVS conditions of the DAB converter were analyzed. However, the speed of charging switch junction capacitance was not studied. In [7], the resonant transition during the dead time of the DAB converter was characterized in time domain with consideration of its operation

Fig. 1. Topology of the DAB converter.

modes. In [8], the ZVS conditions were studied with the criteria that the energy stored in the inductor is larger than the demanded energy of fully charging and discharging power switches' junction capacitance. Since the junction capacitance is both nonlinear and voltage-dependent, the voltage-averaged junction capacitance was used to obtain a more accurate calculation. In recent years, high frequency converters (switching frequency ≥ 1 MHz) with wide bandgap devices have attract more attention [9]–[11]. A Gallium Nitride (GaN)-based DAB converter was characterized by analyzing the resonant transition in time domain as well as considering the nonlinearity of GaN transistors' junction capacitance. In [13] and [14], the ZVS performance of a 1 MHz DAB converter was investigated. In [15], the resonant transition of ZVS operation was characterized in frequency domain by using harmonic decomposition of the DAB converter's switching-node waveforms. This method is not affected by a sequence change in the modulation pattern and thereby reduces the analysis complexity. In [14] and [15], the transformer's magnetizing inductance was considered in characterizing the resonant transition. With the above characterization methods, the ZVS range of the DAB converter with single-phase-shift-modulation (SPSM), where a phase shift is added between the two full bridges (FBs) to control the power flow, was found to be narrow [3]; however, the ZVS range is able to be extended by using advanced modulation strategies. In [16], the double-phase-shift-modulation (DPSM) was proposed. In comparison of SPSM, the FB at the high voltage side was modulated to generate a three-level voltage, and thereby DPSM can shape the inductor current to fulfill the conditions for ZVS even at light load with an additional degree of freedom, i.e. the duty ratio of the three-level PWM voltage. In [17], both of the FBs were modulated to generate three-level voltages, and the phase shift between them was used to control the power flow,

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Fig. 2. Equivalent circuit of the DAB converter at resonant transition.

Fig. 3. Simplified equivalent circuit of the DAB converter during resonant transition.

while the duty ratio helped extend the ZVS range. In [18] and [19], the triple-phase-shift-modulation (TPSM) was proposed. The FBs were modulated to generate phase shifted three-level PWM voltages with different duty ratios. Compared to the methods in [16] and [17], one more degree of freedom, i.e. the duty ratio of the other FB was used to extend ZVS regions, which were further studied in [20] under various load conditions.

In the literature of the DAB converters, the resonant transition during the dead time was characterized without considering transformer's parasitic capacitances. Given that and at the same time, assuming magnetizing inductance is large enough, the ac inductor location i.e. to which side of the transformer the inductor L_e , as illustrated in Fig.1 is connected, does not affect the resonant transition. Thus, it has no impacts on ZVS operation of the DAB converter as long as its equivalent inductance is the same. As reported in [21], the transformer parasitic capacitance, in fact, can cause current resonance and finally worsen soft switching under SPSM. The impacts of transformer parasitic capacitances are analyzed by modelling the impedance of the ac link. Based on the analysis in frequency domain, an approach of distributing the external inductance on both sides of the transformer is proposed in order to mitigate the current resonance, which also means the placement of ac inductor does affect the ZVS performance. However, no equations are derived to describe how exactly this current resonance will change the switching behavior.

Therefore, this paper characterizes the impacts of transformer parasitic capacitance over ZVS operation of the DAB converters under SPSM, DPSM and TPSM in time domain. A mathematical model is derived to calculate the transition speed of the

switching behavior, which also indicates the ZVS operation region. Then, it is proved experimentally that the location of the external inductance does affect ZVS of the DAB converter due to parasitic capacitances of a real transformer. Moreover, this issue becomes more critical in high frequency ($\geq 1\text{MHz}$) DAB converters equipped with Gallium Nitride (GaN) power transistors because of their smaller output capacitances (dozens to hundreds pico-farad only), which are comparable to transformer parasitic capacitances. In order to ensure ZVS on both the FBs, first, the external inductances need to be split and placed on both sides of the transformer; second, a certain amount of external inductance must be ensured on each side of the transformer, and the corresponding analysis and design is given in this paper.

This paper is organized as follows. After this introduction, Section II derives the mathematical model of the resonant transition with considering transformer parasitic capacitances. Section III compares the derived mathematical model with simulations. Section IV provides the method of using two external inductors to cancel out the influences from the transformer parasitic capacitances, and discusses the inductance selection. Section V presents the experimental results from a 1 MHz GaN-based DAB prototype. Finally, Section VI gives the conclusion.

II. MATHEMATICAL MODEL OF THE RESONANT TRANSITION

A. Equivalent Circuit of the Resonant Transition

During the dead time of the FB connected to V_1 , the equivalent circuit of the DAB converter is given in Fig. 2. C_{oss1} - C_{oss4} represent the switch output capacitances and are assumed the same as C_{oss} . L_{e1} and L_{e2} are the external inductance, L_{k1} and L_{k2} are the transformer leakage inductances, C_{p1} and C_{p2} are the parasitic capacitances of the windings on each side of the ac link, respectively.

As addressed in [22] and [23], the phase-drift phenomenon may happen when the lagging FB starts commutation during the dead time of the leading FB and finishes commutation before the leading FB. As a result, during the dead time of the FB connected to V_1 , an all-inclusive analysis of the resonant transition will treat the FB connected to V_2 as either a bridge composed of four capacitors or a dc voltage source with possible values of V_2 , $-V_2$ and 0. However, it turns out that such all-inclusive analysis is very complicated. In [24], an so-called 'energy deadband' was found, in which the DAB converter's power is uncontrollable. In order to avoid such 'energy deadband', the two FBs must not commute in dead time simultaneously. Therefore, considering both simplicity and no energy deadband, in this paper, we assume the lagging FB always starts commutation after the leading FB's commutation when derive the mathematical model of the resonant transition and carry out the model-based analysis. Then, the FB connected to V_2 is simplified as a voltage source V_{r2} , and it has voltage of V_2 , $-V_2$ or 0, depending on the operation modes of the DAB converter in different modulation schemes.

During the dead time of the FB connected to V_1 , Z_{T2} and Z_{T2}' , denoted in Fig.1, are calculated by (1) and (2), respectively.

Fig. 4. Typical steady-state voltage waveforms of the DAB converter.

$$Z_{T2} = j\omega L_{k2} + \frac{1}{j\omega C_{P2}} \frac{j\omega L_{e2}}{j\omega C_{P2} + j\omega L_{e2}}. \quad (1)$$

$$Z'_{T2} = N^2 Z_{T2} = j\omega N^2 L_{k2} + N^2 \frac{1}{j\omega C_{P2}} \frac{j\omega L_{e2}}{j\omega C_{P2} + j\omega L_{e2}}. \quad (2)$$

Then (2) is rewritten as

$$Z'_{T2} = j\omega N^2 L_{k2} + \frac{\frac{1}{j\omega C_{P2}} j\omega N^2 L_{e2}}{\frac{1}{j\omega C_{P2}} + j\omega N^2 L_{e2}}. \quad (3)$$

Based on (3), the equivalent circuit parameters on the secondary side of the transformer in Fig. 2 can be referred to the primary side as shown in Fig. 3, where

$$V'_{r2} = NV_{r2}, L'_{e2} = N^2 L_{e2}, C'_{P2} = \frac{C_{P2}}{N^2}, L_k = L_{k1} + N^2 L_{k2} \quad (4)$$

B. Label the Switching Patterns

As introduced in Section I, v_{ac1} can be either a two-level or a three-level voltage, depending on the adopted modulation strategy. Even though there are many combinations of two-level or three-level v_{ac1} and V_{r2} voltage of V_2 , $-V_2$ or 0, the analysis can be simplified by using defined switching patterns. Note that the switching patterns defined here do not apply to hard-switching transitions.

Typical waveforms of v_{ac1} and v_{ac2} are given in Fig. 4 and both of them are three-level voltages. v_{ac1} is modulated to have an inner phase shift ϕ_S and lagged behind v_{ac2} by a phase shift ϕ_{P-S} . v_{ac2} is modulated to have an inner phase shift ϕ_P . This typical waveform can be used to describe the operation of DAB converters with SPSM, DPSM and TPSM by assigning different values to ϕ_P , ϕ_S and ϕ_{P-S} . Definitions of the switching-patterns are given in Table I, and the relationships of ϕ_P , ϕ_S and ϕ_{P-S} under different modulation strategies are summarized in Table II. Four switching patterns, i.e. Switching-Pattern #1~#4, are defined, so that investigating the impacts of ac inductor placement on the DAB converters with SPSM, DPSM and TPSM can be simplified as characterizing Switching-Pattern #1~#4.

C. Equations of the Resonant Transition

The aims of this paper are: 1) to raise the issue, which is the impact of ac inductor placement on ZVS operation of the DAB converter, and 2) to provide a solution, i.e. put at least the minimum amount of external inductance on each side of the ac link. Deriving equations of all of the four switching patterns will be the most preferred way of reaching these two aims. However, in Table II, one can conclude that one of the FBs will be in hard

TABLE I. DEFINITIONS OF SWITCHING PATTERNS

Switching pattern number	Definition	Typical waveform, as highlighted in red
#1	During resonant transition, (i) the switching node voltage of the FB in resonant transition reverses polarity; (ii) the switching node voltage of the other FB is of a non-zero value.	
#2	During resonant transition, (i) the switching node voltage of the FB in resonant transition changes from a non-zero value to zero or from zero to a non-zero value; (ii) the switching node voltage of the other FB is of a non-zero value	
#3	During resonant transition, (i) the switching node voltage of the FB in resonant transition reverses polarity; (ii) the switching node voltage of the other FB is zero.	
#4	During resonant transition, (i) the switching node voltage of the FB in resonant transition changes from a non-zero value to zero or from zero to a non-zero value; (ii) the switching node voltage of the other FB is zero.	

TABLE II. MODULATION STRATEGIES AND THE SWITCHING PATTERNS INVOLVED

Modulation Strategy	Relationship of phase shift angle	Switching pattern of v_{ac1}	Switching pattern of v_{ac2}
SPSM	$\phi_P = \phi_S = 0$	Switching-Pattern #1	Switching-Pattern #1
Modulation in [13]	$\phi_P = 0, \phi_S \neq 0$	Switching-Pattern #2	Switching-Pattern #1
Modulation in [13]	$\phi_S = 0, \phi_P \neq 0, \phi_P < \phi_{P-S}$	Switching-Pattern #1	Switching-Pattern #2
Modulation in [13]	$\phi_S = 0, \phi_P \neq 0, \phi_P > \phi_{P-S}$	Switching-Pattern #3	Switching-Pattern #2 + hard switching
DPSM	$\phi_P \neq \phi_S \neq 0, \phi_P < \phi_{P-S}$	Switching-Pattern #2	Switching-Pattern #2
DPSM	$\phi_P \neq \phi_S \neq 0, \phi_P > \phi_{P-S}$	Switching-Pattern #4 + Switching-Pattern #2	Switching-Pattern #2 + hard switching
TPSM	$\phi_P \neq \phi_S \neq 0, \phi_P < \phi_{P-S}$	Switching-Pattern #2	Switching-Pattern #2
TPSM	$\phi_P \neq \phi_S \neq 0, \phi_{P-S} < \phi_P < \phi_{P-S} + \phi_S$	Switching-Pattern #4 + Switching-Pattern #2	Switching-Pattern #2 + hard switching
TPSM	$\phi_P \neq \phi_S \neq 0, \phi_P > \phi_{P-S} + \phi_S$	Switching-Pattern #4	Switching-Pattern #2 + hard switching

switching when either Switching-Pattern #3 or Switching-Pattern #4 is used, and then the mathematical analysis can be simplified as described below.

Due to hard switching, Switching-Pattern #3 and Switching-Pattern #4 cannot be used in practice, so that deriving equations for Switching-Pattern #3 and Switching-Pattern #4 are omitted here and the mathematical work focuses on Switching-Pattern #1 and Switching-Pattern #2. Moreover, only the equations of Switching-Pattern #1 are derived in order to implement theoretical analysis. Nevertheless, experiments under Switching-Pattern #2 are completed to investigate the impact of ac inductor placement on the DAB converter in this mode.

In addition, the derived equations can be directly used for analyzing Switching-Pattern #3 by changing the initial condition of V_{r2}' to 0, as stated in Section IV-C. On the other hand, the derived equations can also be modified to characterize Switching-Pattern #2 and Switching-Pattern #4 by changing the initial voltage across C_{oss1} - C_{oss4} at the beginning of the resonant transition. This modification does not change the order of the final differential equation. Therefore, the numerical solver chosen in this paper can be used to solve newly derived models with high accuracy.

From Fig. 3, according to Faraday's law, Kirchhoff's voltage law and Kirchhoff's current law, the following equations can be given.

$$v'_{Le2}(t) = L'_{e2} \frac{di'_{Le2}(t)}{dt} \quad (5)$$

$$v'_{cp2}(t) = v'_{Le2}(t) + V'_{r2} \quad (6)$$

$$i'_{cp2}(t) = C'_{p2} \frac{dv'_{cp2}(t)}{dt} \quad (7)$$

$$i_{Lk}(t) = i'_{cp2}(t) + i'_{Le2}(t) \quad (8)$$

$$v_{Lk}(t) = L_k \frac{di_{Lk}(t)}{dt} \quad (9)$$

$$v_{cp1}(t) = v_{Lk}(t) + v'_{cp2}(t) \quad (10)$$

$$i_{cp1}(t) = C_{p1} \frac{dv_{cp1}(t)}{dt} \quad (11)$$

$$i_{Le1}(t) = i_{cp1}(t) + i_{Lk}(t) \quad (12)$$

$$v_{Le1}(t) = L_{e1} \frac{di_{Le1}(t)}{dt} \quad (13)$$

$$v_{ac1}(t) = v_{Le1}(t) + v_{cp1}(t) \quad (14)$$

$$i_{c4}(t) = C_{oss} \frac{dv_{c4}(t)}{dt} = C_{oss} \frac{d(\frac{V_1 - v_{ac1}(t)}{2})}{dt} \quad (15)$$

$$i_{Le1}(t) = i_{c4}(t) - i_{c3}(t) \quad (16)$$

$$i_{c3}(t) = -i_{c4}(t) \quad (17)$$

$$i_{Le1}(t) = 2i_{c4}(t) = -C_{oss} \frac{dv_{ac1}(t)}{dt} \quad (18)$$

Substituting (5)-(11) into (12), substituting (5)-(17) into (18), and equaling (12) and (18) obtain the following differential equation.

$$\begin{aligned} & -C_{oss}[L_{e1}C_{p1}L_kC'_{p2}L'_{e2} \frac{d^6 i'_{Le2}(t)}{dt^6} + (L_{e1}C_{p1}L_k + L_{e1}C_{p1}L'_{e2} \\ & + L_{e1}C'_{p2}L'_{e2} + L_kC'_{p2}L'_{e2}) \frac{d^4 i'_{Le2}(t)}{dt^4} + (L_{e1} + L_k + L'_{e2}) \frac{d^2 i'_{Le2}(t)}{dt^2}] \\ & = C_{p1}L_kC'_{p2}L'_{e2} \frac{d^4 i'_{Le2}(t)}{dt^4} + (C_{p1}L_k + C_{p1}L'_{e2} + C'_{p2}L'_{e2}) \frac{d^2 i'_{Le2}(t)}{dt^2} + i'_{Le2}(t) \end{aligned} \quad (19)$$

The solution of $i_{Le2}'(t)$ can be obtained by solving (19) numerically. Since all of the circuit states, i.e. $v_{cp2}'(t)$, $i_{cp2}'(t)$, $i_{Lk}(t)$, $v_{cp1}(t)$, $i_{cp1}(t)$, $i_{Le1}(t)$, $v_{ac1}(t)$, etc., are functions of $i_{Le2}'(t)$, the resonant transition is numerically solved in time domain.

TABL III. PARAMETERS USED IN CALCULATIONS AND SIMULATIONS

C_{oss}	17 pF	N	4
C_{p1}	18 pF	L_{e1}	1 μ H-19 μ H
C_{p2}	285 pF	L_{e2}	(20 μ H- L_{e1})/ N^2
V_1	200 V	L_{k1}	300 nH
V_2	50 V	L_{k2}	20 nH

 Fig. 5. Calculated and simulated T_r vs. L_{e1} .

 TABLE IV. $i_{Le2}'(0)$ OBTAINED FROM SIMULATIONS

V_2	50V ($V_1=NV_2$)	40V ($V_1>NV_2$)	60V ($V_1<NV_2$)
$i_{Le2}'(0) @ V_{r2}=V_2$ (A)	-1.4	-1.3	-0.9
$i_{Le2}'(0) @ V_{r2}=0$ (A)	-0.7	-1.0	-0.4

III. SIMULATIONS

Equation (19) is firstly verified by simulations of the DAB converter with SPSM under $V_1=N \cdot V_2$ and a V_1 -to- V_2 phase shift angle of $-6\pi/25$, which is also the phase shift angle used in the experiments. In MATLAB with PLECS Blockset and (19) is numerically solved with ODE-solver-'ode45'. The parameters used in calculations and simulations are given in Table III. These parameters are extracted from the constructed experimental prototype. In the calculations, $i_{Le2}'(0)$ is set to be -1.4, which is obtained from simulations done in PLECS using parameters in Table III. Note that $i_{Le2}'(0)$ could also be calculated by the equation given in (20), where f_s is switching frequency. $i_{Le2}'(0)$ is set to be $(V_1 - V_{r2}')/(L_{e1} + N^2 L_{e2})$, while $d^2 i_{Le2}'(0) \sim d^5 i_{Le2}'(0)$ are set to be 0 because of the linear shape of the ac current before the dead time. With different L_{e1} , the calculated and simulated T_r , which is defined as the time duration of $v_{ac1}(t)$ rising from -200 V to 200 V with $V_{r2}=50$ V, are shown in Fig. 5. L_{e2} is adjusted according to L_{e1} in order to ensure the same power delivered by the DAB converter. Note that T_r is the minimum dead time to achieve ZVS.

In Fig. 5, the calculations and the simulations show the same trend: when L_{e1} is small, e.g. 1 μ H, which means that most of the external inductance is contributed by L_{e2} , it takes approximately 20 ns for $v_{ac1}(t)$ to rise from -200 V to 200 V. When L_{e1} is increased to approximately 2 μ H, which means that additional 1 μ H is assigned to L_{e1} , it only takes approximately 5 ns for $v_{ac1}(t)$ to rise from -200 V to 200 V. When L_{e1} is increased further, the required time to achieve ZVS remains at around 5 ns. The reason for this discontinuity of T_r when increasing L_{e1} will be given in Section IV. There are differences between the calculations and the simulations, which are caused by the inherent errors of the numerical-solutions used. Since the simulations and calculations shows the same trend, and the calculated maximum T_r and the calculated minimum T_r matches the simulated ones, the derived

(a). Calculated $v_{ac1}(t)$ at $L_{e1}=1\mu\text{H}$ (b). Calculated $v_{ac1}(t)$ at $L_{e1}=3\mu\text{H}$ (c). Calculated $v_{ac1}(t)$ at $L_{e1}=10\mu\text{H}$ (d). Calculated $v_{ac1}(t)$ at $L_{e1}=15\mu\text{H}$ Fig. 6. Calculated $v_{ac1}(t)$ at different L_{e1} .

mathematical model is adopted to carry out the analysis in Section IV.

$$i'_{Le2}(0) = -\frac{\Phi p-s(-2NV_2)+V_1+NV_2}{\pi 4Nf_s(L_{e1}+N^2L_{e2})} \quad (20)$$

IV. ANALYSIS OF THE RESONANT TRANSITION AND ZVS

In this section, the resonant transition during the dead time of the DAB converter under $V_1=N\cdot V_2$ with $V_{r2}=V_2$, i.e. Switching-Pattern #1 with $V_1=N\cdot V_2$, is analyzed in detail. After this detailed analysis, the calculated T_r vs. L_{e1} under $V_1=N\cdot V_2$, $V_1>N\cdot V_2$, $V_1<N\cdot V_2$ with V_{r2} equals to V_2 and 0, i.e. Switching-Pattern #1 and Switching-Pattern #3 with $V_1=N\cdot V_2$, $V_1>N\cdot V_2$, $V_1<N\cdot V_2$, are directly given. Because the DAB converter's waveforms are symmetrical in every half-switching cycle, only the analysis in the positive half cycle is given. Since T_r is the minimum dead time to achieve ZVS, the analysis of T_r is critical to study the impacts of ac inductance placement.

A. Influences of Transformer Parasitic Capacitances on T_r

Assume that the resonance is not interrupted by the actions of the switches, and by using the parameters listed in Table III, the calculated waveforms of $v_{ac1}(t)$ with L_{e1} of 1 μH , 3 μH , 10 μH or 15 μH are plotted in Fig. 6, where $V_1=N\cdot V_2$, $V_{r2}=V_2$, $i_{Le2}'(0)=-1.4$, $di_{Le2}'(0)=(V_1-V_{r2})/(L_{e1}+N^2L_{e2})$, and $d^2i_{Le2}'(0)\sim d^2i_{Le2}'(0)=0$.

The resulted $v_{ac1}(t)$ is no longer pure sinusoidal, and therefore, T_r or the minimum dead time for achieving ZVS calculated in [7] may cause errors. The expression of $v_{ac1}(t)$ derived in [7] is given as (21), where definitions of Z_0 and ω_0 are defined in (22) and (23). As shown in Fig. 6(a), the non-sinusoidal nature of $v_{ac1}(t)$ requires the T_r to be around 20 ns, which is 3 times larger than the 4.9 ns calculated by the equations in [7], where C_{p1} and C_{p2} are neglected. Thus, if the dead time is set based on the equations in [7], the DAB converter will lose ZVS.

$$v_{ac1}(t) = -NV_2 - i_{Le1}(0)Z_0 \sin(\omega_0 t) + (-V_1 + NV_2)\cos(\omega_0 t) \quad (21)$$

$$Z_0 = \sqrt{\frac{L_e}{C_{oss}}} \quad (22)$$

$$\omega_0 = \frac{1}{Z_0} \quad (23)$$

B. Placing External Inductances to Shorten T_r

The existence of C_{p1} and C_{p2} leads to a non-sinusoidal $v_{ac1}(t)$, and it is possible to make $v_{ac1}(t)$ quasi-sinusoidal by adjusting L_{e1} . As shown in Fig. 6(b)-(d), if L_{e1} increases, $v_{ac1}(t)$ will become less non-sinusoidal. And with a quasi-sinusoidal $v_{ac1}(t)$, T_r can be reduced. With a reduced T_r , ZVS operation can be guaranteed even with a small dead time.

In Fig. 7, the influences of L_{e1} on T_r under different C_{p1} are given, where C_{p2} is assigned to be N^2 times of C_{p1} . As shown, when L_{e1} is small, T_r increases when C_{p1} and C_{p2} are increased. In order to keep T_r small even though C_{p1} and C_{p2} are increased, L_{e1} needs to be increased accordingly.

In Fig. 8, the influences of L_{e1} on T_r under different $i_{Le2}'(0)$ are given. Since $i_{Le2}'(0)$ is related to the input/output power, i.e. the higher the input/output power, the larger the absolute value of $i_{Le2}'(0)$, Fig. 8 actually indicates the influences of L_{e1} on T_r under different power. As shown, when L_{e1} is small, T_r increases when $i_{Le2}'(0)$ is increased. In order to keep T_r small even though $i_{Le2}'(0)$ is increased, L_{e1} needs to be increased accordingly. The smallest T_r under different $i_{Le2}'(0)$ increases when $i_{Le2}'(0)$ is increased.

Fig. 9 shows the realization of ZVS under different combinations of L_{e1} and dead time T_d . The combinations with which ZVS can be achieved are marked as grey area. Based on

 Fig. 7. Calculated T_r vs. L_{e1} under different C_{p1} , $C_{p2}=N^2C_{p1}$ @ $i_{Le2}'(0)=-1.4A$.

 Fig. 8. Calculated T_r vs. L_{e1} under different $i_{Le2}'(0)$, @ $C_{p1}=18$ pF.

 Fig. 9. Realization of ZVS, @ $i_{Le2}'(0)=-1.4A$, $C_{p1}=18$ pF.

 Fig. 10. Calculated T_r vs. L_{e1} under $V_1=N\cdot V_2$, $V_1>N\cdot V_2$, $V_1<N\cdot V_2$ & $V_{r2}=V_2$ and 0.

analysis on Fig. 7 and 8, the influences of power and transformer parasitic capacitance on realization of ZVS is summarized in Fig. 9 as well. As shown, the ZVS area is enlarged when power is increased or transformer parasitic capacitance is decreased, and vice versa.

Fig. 6 also explains the discontinuity of T_r in Figs. 5, 7 and 8. As shown in Fig. 6(a), the first wave crest of $v_{ac1}(t)$ is smaller than 200 V, which is the DC voltage connected to the FB. $v_{ac1}(t)$ reaches 200 V after the first wave trough. As a result, T_r must be larger than the time when $v_{ac1}(t)$ is at the first wave trough. However, once the first wave crest of $v_{ac1}(t)$ is increased to 200 V with increasing L_{e1} , T_r will be directly shortened to be the time when $v_{ac1}(t)$ is at the first wave crest.

C. T_r versus L_{e1} Curves of Switching-Patterns #1 and #3

The calculated T_r versus L_{e1} curves with $V_1=N\cdot V_2$, $V_1>N\cdot V_2$, $V_1<N\cdot V_2$ & $V_{r2}=V_2$ and 0 are given in Fig. 10. $i_{Le2}'(0)$ used in the calculations is obtained from simulations done in PLECS and are given in Table IV. Note that $i_{Le2}'(0)$ could also be calculated by the equation given in (24) [25]. These simulations are based on parameters in Table III, except V_2 , which is changed accordingly in order to make V_1 be either larger/smaller than or equal to NV_2 . The ϕ_{p-s} used in simulations is $-6\pi/25$, and an additional inner phase shift angle ϕ_p of $8\pi/25$ is added to the FB connected to V_2 in $V_{r2}=0$ cases.

$$i_{Le2}'(0) = -\frac{V_1\left(1-\frac{\phi_p}{\pi}\right)+NV_2\frac{2|\phi_{p-s}|+\phi_s-\pi}{\pi}}{2f_s(L_{e1}+N^2L_{e2})} \quad (24)$$

As shown in Fig. 10, in all of the considered cases where $V_1=N\cdot V_2$, $V_1>N\cdot V_2$, $V_1<N\cdot V_2$ & $V_{r2}=V_2$ and 0, T_r decreases sharply at certain L_{e1} . Therefore, the impact of ac inductor placement on the resonant transition does exist. Moreover, the following patterns and the corresponding characterizations can be found:

- The solid lines are not overlapped with each other, neither do the dashed lines. Therefore, the relationship between V_1 and NV_2 affects the resonant transition.
- All of the dashed lines are on the up-right side of the solid lines with the same color. Therefore, with the same relationship between V_1 and NV_2 , 1) T_r under $V_{r2}=0$ is always larger than T_r under $V_{r2}=V_2$; 2) L_{e1} with which T_r starting to drastically decrease under $V_{r2}=0$ is larger than that under $V_{r2}=V_2$.
- As shown in the T_r vs. L_{e1} curve under $V_1<N\cdot V_2$ & $V_{r2}=0$, T_r drops drastically twice, i.e. T_r firstly steps down from 50 ns to 35 ns at $L_{e1}=1$ μ H, then it steps down again from 50 ns to 20 ns at $L_{e1}=8$ μ H. This double stepping-down in T_r indicates that under $V_1<N\cdot V_2$ & $V_{r2}=0$, if L_{e1} is smaller than 1 μ H and dead time is set to be a value larger than 50 ns in order to achieve ZVS, v_{ac1} during dead time will have double wave crests and double wave troughs. These multi-resonances in the switching-node voltage of one of the FBs will cause distortions in the ac current, which flows in the other FB as well. As a result, the switching-node voltage of the other FB will be distorted. To conclude, prolonging the dead time does help realize ZVS even with small external inductance. However, this approach will cause distortions in the switching-node voltages and the ac current.

Fig. 11. Selection of the external inductance.

 Fig. 12. $Index_{sym}$ vs. L_{e1} & C_{p1}

Considering the DAB converter with SPSM, due to its symmetrical topology, the same conclusion can be applied to both of the FBs. Therefore, the external inductance needs to be placed on both sides of the transformer. By doing so the ZVS in both of the FBs can be achieved within a short dead time. This inductance assignment needs to ensure that the external inductances on both sides are larger than the calculated minimum values, with which v_{ac1} and v_{ac2} during dead time are quasi-sinusoidal, respectively. As shown in Figs. 7 and 8, the minimum values of the required external inductances are affected by the transformer parasitic capacitance, the relationship between V_1 and NV_2 , and the modulation strategy.

D. Selection of External Inductance

Without considering ZVS performance, the constraint conditions of the external inductance L_{e1} and L_{e2} are:

$$\begin{cases} L_{e1} \geq 0 \\ L_{e2} \geq 0 \\ L_{e1} + N^2 L_{e2} = L_e \end{cases} \quad (25)$$

Plot (25) and yield the solid lines in Fig. 11, where the selection of external inductance is visualized.

With achieving ZVS as one of the criteria, the dead time T_d need to be determined at first. Using (19), in the SPSM

controlled DAB converter with the parameters in Table III, the rising time of v_{ac1} will be shortened from 20 ns to 5 ns approximately once L_{e1} exceeds 2 μH , and the rising time of v_{ac2} will also be shortened from 20 ns to 6 ns approximately once L_{e2} exceeds 100 nH. When T_d is selected to be 6ns, in order to achieve ZVS, L_{e1} can be any value between 2 μH and 18.4 μH (at $L_{e1}=18.4 \mu\text{H}$, L_{e2} reaches 100 nH), L_{e2} can be any value between 100 nH and 1.125 μH (at $L_{e2}=1.125 \mu\text{H}$, L_{e1} reaches 2 μH). The candidates of L_{e1} and L_{e2} for realizing ZVS under T_d of 6ns are visualized as the grey area in Fig. 11.

In order to specify the inductances of the two external inductors, at least one criteria in addition to achieving ZVS is needed. The additional criteria used in this manuscript is to keep the ac link of the DAB converter as symmetrical as possible. As introduced in Section I, one of the attracting feature of the DAB converter is its symmetrical structure, which makes the DAB converter a promising topology for applications requiring bidirectional power flow. In bidirectional converters, the index of performance, e.g. efficiency, temperature rise of critical components, noises, at one direction of power flow is desired to be the same to those at the other direction. A detailed modeling of the effects of selecting the external inductances on the mismatch of index of performance is difficult. The degree of symmetry of the ac link can be estimated by calculating its input impedance in both directions of power flow. When power flows from v_{ac1} to v_{ac2} , the input impedance of the ac link $Z_{in_ac_12}$ can be calculated as,

$$Z_{in_ac_12} = j\omega L_{e1} + \frac{j\omega(N^2 L_{e2} + L_k) - j\omega^3 L_{e2} C_{p2} L_k}{Denom_{12}} \quad (26)$$

where

$$Denom_{12} = 1 - \omega^2 L_{e2} N^2 C_p + L_k \omega^2 C_{p1} (\omega^2 L_{e2} C_{p2} - 1) \quad (27)$$

$$C_p = C_{p1} + \frac{C_{p2}}{N^2} \quad (28)$$

On the other hand, when power flows from v_{ac2} to v_{ac1} , the input impedance of the ac link $Z_{in_ac_21}$ can be calculated as,

$$Z_{in_ac_21} = j\omega L_{e2} + \frac{\frac{1}{N^2} [j\omega(L_{e1} + L_k) - j\omega^3 L_{e1} C_{p1} L_k]}{Denom_{21}} \quad (29)$$

where

$$Denom_{21} = 1 - \omega^2 L_{e1} C_p + L_k \omega^2 \frac{C_{p2}}{N^2} (\omega^2 L_{e1} C_{p1} - 1) \quad (30)$$

The mismatch between $Z_{in_ac_12}$ and $Z_{in_ac_21}$ is defined as:

$$\delta Z_{in_ac} = Z_{in_ac_12} - N^2 Z_{in_ac_21} \quad (31)$$

Substituting (26) and (29) into (31), δZ_{in_ac} can be expressed as,

$$\delta Z_{in_ac} = j\omega \delta L_e + \frac{k_{Le} \delta L_e + k_{Cp} \delta C_p + k_{CL} \delta C_L + k_{CCL} \delta C_{CL}}{Denom_{12} Denom_{21}} \quad (32)$$

where $Denom_{12}$, $Denom_{21}$, k_{Le} , δL_e , k_{Cp} , δC_p , k_{CL} , δC_L , k_{CCL} and δC_{CL} is defined as

$$k_{Le} = -j\omega - j\omega^3 L_k C_p + \frac{j\omega^5 L_k^2 C_{p1} C_{p2}}{N^2} \quad (33)$$

$$\delta L_e = L_{e1} - N^2 L_{e2} \quad (34)$$

$$k_{Cp} = j\omega^3 L_k^2 - j\omega^5 N^2 L_{e1} L_{e2} L_k C_p + j\omega^7 L_{e1} L_{e2} C_{p1} C_{p2} L_k^2 \quad (35)$$

$$\delta C_p = C_{p1} - \frac{C_{p2}}{N^2} \quad (36)$$

$$k_{CL} = j\omega^3 L_k \quad (37)$$

$$\delta C_L = L_{e1} C_{p1} - L_{e2} C_{p2} \quad (38)$$

$$k_{CCL} = -j\omega^5 L_k^2 \quad (39)$$

$$\delta C_{CL} = L_{e1} C_{p1} C_{p1} - L_{e2} C_{p2} \frac{C_{p2}}{N^2} \quad (40)$$

The index of symmetry $Index_{sym}$, which is defined in (41), is used to estimate the influences of δZ_{in_ac} on the degree of symmetry of bidirectional power flow in the DAB converter with SPSM.

$$Index_{sym} = \sum_{n=1,3,5,\dots} \frac{1}{n} |\delta Z_{in_ac}| \big|_{\omega=2n\pi f_s} \quad (41)$$

n is defined as odd integers because that, ideally, in the DAB converter with SPSM, voltage and current flowing through the ac link are composed by the fundamental component and the odd order harmonics [26].

As expressed in (32), there are several contributors, i.e. δL_e , δC_p , δCL , and δCCL , to δZ_{in_ac} . Therefore, in order to minimize $Index_{sym}$, the selection of external inductances depends on the extent to which C_{p1} deviates from C_{p2}/N^2 .

Fig. 12 plots $Index_{sym}$ under varying L_{e1} and C_{p1} ($Index_{sym}$ is calculated by up to 7th order of odd harmonics). Note that when C_{p1} is 18 pF as in Table III, δC_p is 0.2 pF, which is only 1% of C_{p1} . As shown, with the parameters in Table III, which is extracted from the transformer used in the experiments, $Index_{sym}$ is minimized when δL_e reaches zero. Therefore, with the objective of keeping the ac link as symmetrical as possible, the external inductances should be selected so that L_{e1} is equal to $N^2 L_{e2}$. This constraint yields the dot line in Fig. 11. Resultantly, as shown in Fig. 11, the star, which is the crossing of the dot line and the top solid line, are the final selections of the external inductances when $L_e=20 \mu\text{H}$. Since the star is inside of the grey area, this final selection will guarantee the ZVS of all of the switches.

In the experiments, the total high-voltage-side referred inductance is 13 μH . With this inductance the converter works at rated power of 250 W. The ZVS area with $L_e=13 \mu\text{H}$ & $T_d=6$ ns is shown as the grey area bounded by dashed lines in Fig. 11. Analysis on $Index_{sym}$ indicates that $Index_{sym}$ is minimized when δL_e reaches zero. Therefore, the final selection of external inductance is the triangle as marked in Fig. 11. The total high-voltage-side referred inductance is kept approximately the same for the three setups, i.e. 13.2 μH in the single-high-voltage-side configuration, 12.6 μH in the single-low-voltage-side configuration and 12.3 μH in the configuration with two external inductors. In the configuration with two external inductors, the high-voltage-side-referred L_{e2} is 5.4 μH , L_{e1} is 6.9 μH . Note that considering the approximate 70-90 nH inductance of the PCB traces in the ac link on the low voltage side, the high-voltage-side-referred ac inductance on the low voltage side is 6.5-6.8 μH , which is close to L_{e1} .

V. EXPERIMENTS

A 1 MHz DAB converter prototype is built with the specifications listed in Table V to verify the theoretical analysis. The photos of the constructed prototype and the corresponding test setup are given in Fig. 13. In Fig. 13 (b), the transformer and inductors, which have been shown in Fig. 13(a), are removed in order to show the items of the prototype under the magnetic components. The parameters extracted from the prototype are given in Table III. v_{ac1} and v_{ac2} are measured by differential probes with bandwidth (BW) of 25 MHz, which is too low to monitor those multiple resonances in v_{ac1} and v_{ac2} . Therefore, the multiple resonances are monitored by measuring the drain-

source voltage (v_{ds}) of the switches. v_{ds} is measured by Lecroy PPE4kV with 6 pF capacitance and BW of 400 MHz. i_{Le1} and i_{Le2} are measured by Lecroy CP030 and HIOKI3273-50 current probes, which have the same BW of 50MHz, respectively. In the following waveforms, v_{ds} of EPC2016C is measured from Drain-to-Source of S_6 , and v_{ds} of GS66502B is measured from Drain-to-Source of S_4 . N4L PPA5530 power analyzer measures

TABLE V. SPECIFICATIONS OF THE DAB PROTOTYPE

	Description	Parameter
Specifications of the prototype	Switching Frequency	1 MHz
	DC voltage on the HV side	200 Vdc
	DC voltage on the LV side	50 Vdc
	Rated output power	250 W
Power semiconductor	Switches on the HV side	GS66502B
	Switches on the LV side	EPC2016C
Transformer	Core's material, type, size and turns ratio	Hitachi Metals, ML91S, 2×EE-21.8-11.4-15.8, 8:2
Dead Time	HV side & LV side	25 ns

(a) Test setup

(b) Detailed view of the prototype
Fig. 13. Pictures of the hardware.

the input & output voltage, the input & output power, and the efficiency without any probes connected.

The verification of analysis on Switching-Pattern #1 and #2 are given in Subsection-A and -B, respectively. No experiments are performed to investigate Switching-Pattern #3 and #4 because hard switching always comes along with these two patterns. The efficiency measurements are given in Subsection-C to show that splitting the external inductances is able to improve the efficiency. Finally, all the experimental results are summarized in Subsection-D.

A. Experiments on the DAB converter with SPSM

First, place an external inductor of $13.2\ \mu\text{H}$ on the high voltage side (HVs); directly connect the low voltage side (LVs) FB to the transformer and no external inductance in between. The LVs ac current is not measured in such setup to minimize the external inductance on the LVs. This external inductance configuration will ease ZVS realization of the HVs because the external inductance of $13.2\ \mu\text{H}$ is larger than the required minimum inductance of $2\ \mu\text{H}$; however, deteriorate the ZVS on the LVs. As in the measured experimental waveforms shown in Fig.12, GaN transistors GS66502B have soft switching, however, EPC2016C is partial hard switching due to the long rising time of its v_{ds} . The waveforms of the rising and the falling of v_{ds} of EPC2016C are not symmetrical. The first wave crest during rising is larger than V_2 , in converse, the first trough during falling is larger than 0. This is caused by the dc offset in i_{Le2} , which leads to different initial currents for the high-side-to-low-side and low-side-to-high-side commutations. In spite of the asymmetry, the measured time duration for v_{ds} of S_6 to rise from 0 to V_2 after the first wave trough is 20 ns and match the calculations. So does the measured time duration for v_{ds} of S_6 to fall from V_2 to 0 after the first wave crest. However, the rising time of v_{ds} of GS66502B is 21 ns, which is different from the calculated 5ns. This mismatching in GS66502B is caused by the input capacitance of the probe and the parasitic capacitance of the PCB traces, which is around 25 pF. These two together contribute an additional capacitance of around 31 pF in parallel to the output capacitance of GS66502B, which is 17 pF. Without consideration of the parasitic capacitances of the transformer, the total capacitance of 48 pF will lead to a rising time of around 15 ns according to the equations in [7]. Because the output capacitance of EPC2016C is 310 pF, which is much larger than the input capacitance of the probe and the parasitic capacitance of the PCB traces, the calculated rising time of EPC2016C is thereby close to the measured one. Therefore, the validity of the derived model in Section II is verified.

Second, the external inductance of 787 nH is placed on the LVs and directly connect the FB on the HVs to the transformer. This configuration will ease ZVS realization on the LVs because the external inductance of 787 nH is larger than the required minimum inductance of 100 nH, and on the other hand, it will deteriorate the ZVS on the HVs. As the waveforms shown in Fig.13, EPC2016C has soft switching; however, GS66502B is partial hard switching. Comparing the measured rising time to the calculated one, v_{ds} rising time of EPC2016C is 5 ns, which matches the calculation. For GS66502B, partial hard switching occurs when the external inductor is located on the LVs. Under the same dead time of 25 ns, GS66502B achieves ZVS when the external inductor is located on the HVs. Therefore, the influences on the ZVS due to the transformer parasitic capacitances, which have been analyzed in Section IV-A, are verified.

Finally, the external inductors of $6.9\ \mu\text{H}$ and 336 nH are added and placed on both the HVs and LVs, respectively. Based on the analysis, this configuration will ease ZVS realization of both the FBs since the external inductance on both sides of the transformer are larger than the required minimum ones. The

corresponding experimental waveforms are given in Fig. 16. As shown in these waveforms, both GS66502B and EPC2016C are able to achieve ZVS. Therefore, the proposed solution to mitigate the effects of the transformer parasitic capacitances on ZVS, i.e. assigning the external inductance to ensure the

Fig. 14. Measurements of the DAB converter with SPSM and HVs-located external inductance.

minimum needed inductance on both sides of the transformer, is verified on the DAB converter with SPSM.

B. Experiments on the DAB converter with TPSM

The following experiments are implemented on the same DAB prototype with TPSM to investigate the impacts of ac inductor placement on Switching-Pattern #2. The test procedure

Fig. 15. Measurements of the DAB converter with SPSM and LVs-located external inductance.

is the same as that presented in the Sub-section A. First, place an external inductance of 787 nH on the LVs and directly connect the FB on the HVs to the transformer. The measured waveforms are given in Fig. 17. As shown, this configuration eases ZVS realization on the LVs, and deteriorates the ZVS on the HVs. Both of the FBs achieve ZVS when the external inductances are

Fig. 16. Measurements the DAB converter with SPSM and split external inductance.

placed on both the HVs and LVs by adding the external inductors of $6.9 \mu\text{H}$ and 336 nH , respectively. The measured waveforms are given in Fig. 18. As shown, both GS66502B and EPC2016C achieve ZVS. Therefore, the proposed solution to mitigate the effects of the transformer's parasitic capacitances on ZVS, i.e. assigning the external inductance to ensure the

Fig. 17. Measurements of the DAB converter with TPSM and LVs-located external inductance.

minimum needed inductance on both sides of the transformer, is valid and verified on the DAB converter with TPSM.

C. Power and efficiency measurements

The input power, the output power and the efficiency of the prototype under the test conditions stated above are listed in Table VI, where ‘SPSM-H’ means SPSM with external inductor

Fig. 18. Measurements the DAB converter with TPSM and split external inductance.

TABLE VI. POWER AND EFFICIENCY MEASUREMENTS WITHOUT ANY PROBES ATTACHED

	SPSM-H	SPSM-L	SPSM-HL	TPSM-L	TPSM-HL
V_1 (V)	200	201	201	200	201
V_2 (V)	51.0	51.7	49.0	52.8	49.7
ϕ_{P-S}	$-6\pi/25$	$-6\pi/25$	$-6\pi/25$	$-6\pi/25$	$-6\pi/25$
ϕ_P				$2\pi/25$	$2\pi/25$
ϕ_S				$2\pi/25$	$2\pi/25$
Input Power (W)	245	265	266	263	264
Output Power (W)	229	250	250	247	251
Efficiency (%)	93.6	94.6	94.0	94.2	95.0

TABLE VII. SUMMARY OF THERMAL BEHAVIOR OF THE PROTOTYPE UNDER ‘SPSM-L’ AND ‘SPSM-HL’

		SPSM-L	SPSM-HL
Switches on the HVs	Temperature rise (°C)	28.8	21.8
	Thermal picture
Switches on the LVs	Temperature rise (°C)	32.3	34.1
	Thermal picture
Transformer	Temperature rise (°C)	15.3	16.6
	Thermal picture
Inductor on LVs	Temperature rise (°C)	9.9	9.9
	Thermal picture
Inductor on HVs	Temperature rise (°C)		12.1
	Thermal picture

on the HVs, naming of other columns in the header row follows this naming rule, which is ‘modulation strategy’–‘location of the external inductor’. From Table VI, we can see the possibility of increasing the prototype’s efficiency by splitting the external inductance on both sides of the transformer is verified on the prototype modulating by TPSM. Under the same input/output power, the same phase shift angle, and the same input/output voltage, the efficiency is increased by 0.8% by splitting the external inductance.

As shown, the efficiency under ‘SPSM-HL’ is lower than that under ‘SPSM-L’, which means achieving ZVS of all of the switches in the tested DAB converter with SPSM does not necessarily increase the efficiency of the prototype. This observation is worthy of investigation since it indicates that either achieving ZVS does not necessarily brings down the losses of switches or splitting the external inductors on both of the two sides of transformer increases losses of components other than switches.

In order to investigate the loss distribution, thermal measurement is implemented to illustrate the loss changes caused by different placement of inductors. T650sc from FLIR is used to monitor the temperature of individual components of the prototype under ‘SPSM-L’ and ‘SPSM-HL’. The thermal measurements are summarized in Table VII. As illustrated in the rows titled with ‘Switches on the HVs’, the temperature rise of the switches on the HVs under ‘SPSM-HL’ is 7 °C lower or 25% smaller than that under ‘SPSM-L’. It indicates that the losses of the switches on the HVs do decrease when the external inductor is split and placed on two sides of the transformer. The temperature rise of the switches on the LVs increases 1.8 °C in ‘SPSM-HL’ compared to ‘SPSM-L’. Given the same heat dissipation conditions in the experimental setup (the total area of copper used to dissipate heat from GS66502B is the same to that of EPC2016C) and the same junction-to-case thermal resistance of 2 °C/W of GS66502B and EPC2016C, the decreased losses in GS66502B cannot be cancelled out by the increased losses in EPC2016C. Therefore, there are additional losses comes from components other than switches.

This hypothesis is proved by the thermal measurements of the transformer and the inductors, as it shows in the rows titled with ‘Transformer’, ‘Inductor on LVs’ and ‘Inductor on HVs’. The temperature rise of the transformer increases 1.3 °C in ‘SPSM-HL’ compared to ‘SPSM-L’. The loss of the inductor on the LVs remains unchanged in the two setups. The inductor on the HVs has a temperature rise of 12 °C.

Therefore, the efficiency drop of the prototype under ‘SPSM-HL’ compared to ‘SPSM-L’ is mainly caused by the additional losses in the inductor on the high voltage side. This amount of additional loss is large enough to cancel out the amount of reduced losses of the soft-switched transistors on the HVs. The two external inductors need to be carefully designed to keep the inductors’ losses as low as possible.

To conclude, compared to ‘SPSM-L’, all of the switches in the ‘SPSM-HL’ configuration achieve ZVS. However, the additional inductor on the high voltage side might increase the total loss of the converter.

The external inductors used in this paper are designed to achieve ZVS and make the DAB converter symmetrical, as explained in Subsection-D of Section IV. Investigating the design guidelines on how to split the external inductors in order to fulfill other design criteria, e.g. minimize total loss of the converter, could be the topics for the future research.

In summary, it is found that placing inductance on both side of the high frequency transformer can effectively extend the ZVS for all the switching devices on both high voltage and low voltage sides. The converter thereby has less EMI issues. The possibility of boosting the DAB converter's efficiency by solely splitting the external inductance on both of the two sides of the transformer is verified experimentally. Notably, even though the SPSM-HL configuration has slightly lower efficiency than the SPSM-L configuration, GS66502B ZVS operation is ensured and thereby these GaN devices will have lower loss as well as low switching noises. In other words, the losses are moved from GS66502B to the external inductor on the high voltage side, which has a larger area for heat dissipation than GS66502B. Thus, GaN devices can operate more reliably because of less thermal stress.

VI. CONCLUSIONS

This paper addressed the issue of ZVS operation affected by the transformer parasitic capacitance in high frequency DAB converters. A mathematical model, which considers the parasitic capacitances of the transformer, is derived to describe the resonant transition during the dead time. It is found that placing inductance on both side of the high frequency transformer can effectively extend the ZVS for all the switching devices on both high voltage and low voltage sides. A systematic design guideline of the external inductors with the objective of keeping the DAB converter symmetrical is provided. The proposed solution is verified by the measurement results from the constructed 1 MHz GaN-based DAB prototype. In the future research work, inductor design optimization in terms of impedance matching and loss reduction need to be investigated further to improve efficiency under various modulation schemes.

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