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Influence of Parasitic Parameters on DC–DC Converters and Their Method of Suppression in High Frequency Link 35 kV PV Systems

Rui Li ^{1,*} , Fangyuan Shi ¹, Xu Cai ¹ and Haibo Xu ²

¹ School of Electronic Information and Electrical Engineering, Shanghai Jiao Tong University, Shanghai 200240, China; Fangyuan_shi@sjtu.edu.cn (F.S.); xucai@sjtu.edu.cn (X.C.)

² Dongguan South Semiconductor Technology Co., Ltd., Dongguan 523808, Guangdong, China; xuhb@eastups.com

* Correspondence: liruiqd@sjtu.edu.cn

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Abstract: Photovoltaic (PV) power generation has shown a trend towards large-scale medium- or high-voltage integration in recent years. The development of high-frequency link PV systems is necessary for the further improvement of system efficiency and the reduction of system cost. In the system, high-frequency high-step-up ratio LLC converters are one of the most important parts. However, the parasitic parameters of devices lead to a loss of zero-voltage switching (ZVS) in the LLC converter, greatly reducing the efficiency of the system, especially in such a high-frequency application. In this paper, a high-frequency link 35 kV PV system is presented. To suppress the influences of parasitic parameters in the LLC converter in the 35 kV PV system, the influence of parasitic parameters on ZVS is analyzed and expounded. Then, a suppression method is proposed to promote the realization of ZVS. This method adds a saturable inductor on the secondary side to achieve ZVS. The saturable inductor can effectively prevent the parasitic elements of the secondary side from participating in the resonance of the primary side. The experimental results show that this method achieves a higher efficiency than the traditional method by reducing the magnetic inductance.

Keywords: PV system; parasitic parameters; LLC converter; ZVS; saturable inductor

1. Introduction

Since PV power generation systems can alleviate the energy crisis and reduce environmental pollution, they have found broad application worldwide [1–4]. Traditional large-scale PV medium-voltage (MVAC) grid-connected systems mostly adopt centralized large-capacity inverters and line frequency transformers because of their simple structure and low installation cost [5–7]. The traditional structure of PV systems has achieved high efficiency and low cost. However, there are many problems in traditional systems, including mismatch power losses due to the limitations of single maximum power point tracking (MPPT) [8], magnetic loss and reactive power loss due to the line frequency transformer [9], and a massive consumption of material resources. As a result, the efficiency of traditional systems is difficult to improve any further. Although DC–DC converters can be introduced into the system, as described in [10], and additional power converters will allow each photovoltaic array to operate at its maximum power point (MPP), the remaining problems still cannot be solved, due to the use of the centralized inverter and line frequency transformer. With the development of wide band gap devices, high frequency isolation is possible, and PV systems with higher efficiency and power density are expected. To this end, many improved architectures have been proposed. In [11,12], a modular multilevel converter (MMC) was used to replace the line frequency

transformer and central inverter in MVAC grid-connected systems. The MMC was able to easily achieve a medium voltage level, but it could not provide galvanic isolation of the photovoltaic array, and was not found to be suitable for PV systems. In [13], an isolated quasi-z-source DC–DC converter was employed in constructing a medium-voltage level DC bus and providing galvanic isolation for MMC. Nevertheless, controlling the whole system became more complex. Then, in [14], a cascaded H-bridge (CHB) multilevel converter was used to replace the MMC, and an isolated DC–DC converter was used to replace the line frequency transformer. In this architecture, the isolated DC–DC converter and the CHB module could easily be integrated into the PV system to improve the power-density of the system. Since the isolated DC–DC converter needs to undertake MPPT, its voltage gain may vary over a wide range, and its efficiency is difficult to improve. Furthermore, large-scale photovoltaic medium-voltage DC (MVDC) or high-voltage DC (HVDC) grid-connected systems are another promising direction in system structures [15,16]. In [17], a single-stage isolated DC–DC converter was used in an MVDC system, and the line frequency transformer was removed. This single-stage MVDC system was able to reduce system cost, power losses, and system bulk. However, its overall efficiency would be decreased by partial shading [18]. Hence, a two-stage DC–DC converter was used in an HVDC system in order to achieve higher frequency [19]. This paper is based on the high frequency link 35 kV PV system. The introduction of a high-frequency high-voltage isolated DC–DC converter is able to replace the line frequency transformer and save non-renewable resources. Therefore, the high-frequency high-voltage isolated DC–DC converter is one of the core technologies in PV systems.

Isolated high-frequency DC–DC converters can provide a fixed step-up ratio and galvanic isolation [19], and they constitute an important technology in PV power generation systems. A dual active bridge (DAB) was used early in PV applications because of its modular symmetric topological structure and fast dynamic response characteristics [20]. However, its ZVS range is limited by its voltage conversion ratio and load conditions. Therefore, it is difficult to achieve soft-switching throughout the full load range, and it is difficult to improve the efficiency any further [21]. In [22], a current source series resonant converter (SRC) was used as a DC–DC transformer (DCX). However, similar to DAB converters, it is difficult to implement ZVS throughout the full load range using an SRC. Recently, the LLC and CLLC topologies have become attractive due to their desirable characteristics, which include high efficiency, full load range ZVS, and zero-current-switching (ZCS) commutation [23–25]. Since bidirectional power flow is not a prerequisite in PV systems, CLLC converters are not suitable for improving power density. Therefore, LLC converters are the best choice.

However, there are still some problems presented by high-frequency LLC converters. On the one hand, unlike traditional low-voltage and low-frequency transformers, the insulation pressure of transformers in the converters makes the design of transformers and the selection of core materials more complex and more difficult [26]. On the other hand, the parasitic parameters of high-frequency transformer are complex, and these may have negative influences on the circuit. Furthermore, in recent years, SiC-MOSFETs have been widely used in DC–DC converters to pursue higher efficiency and higher power density [27]. Many studies have shown that SiC-based converters have greater advantages in terms of efficiency and power density compared to Si-based converters, especially in high-voltage applications [28–30]. However, both SiC-MOSFETs and SiC diodes have parasitic parameters. With the increase of the switching frequency of the LLC converter, the negative influences of parasitic parameters on the primary side ZVS of the high-frequency LLC converter gradually emerge [31,32]. In [33], the parasitic capacitor was investigated, and an optimized dead band was provided. In [34], the negative impacts caused by the parasitic capacitors of rectifier diodes were discussed. However, this only indicated negative influences on the DC voltage gain. In [31] and [35], the influence mechanism of parasitic parameters was investigated, and a solution was given in which the dead band and magnetic inductance were adjusted. However, the means for accurately selecting the dead band and magnetic inductance were not mentioned in those papers. The use of wide band gap devices can further improve the efficiency and power density of converters in PV systems. Unfortunately, without ZVS, the large

switching losses and the deterioration of EMI characteristics caused by high switching frequency greatly reduces the performance of high-frequency LLC converters [36].

Therefore, this paper presents an architecture for the high frequency link 35 kV PV system, as shown in Figure 1. The distributed three-level boost converters with photovoltaic strings execute the MPPT algorithm. Isolated high-frequency DC–DC converters are used as DCXs between the CBH modules and a common DC bus. This paper studies the influence of parasitic parameters and their method of suppression in high-frequency LLC converters in the 35 kV PV system. This paper describes how the parasitic parameters of SiC devices affect the ZVS of high-frequency LLC converters. Then, a method for suppressing the influence of parasitic parameters is proposed. By adding a saturable inductor on the secondary side, the parasitic components of the secondary side are effectively blocked from participating in the resonance of the primary side, and primary-side ZVS can be ensured. Finally, a multiple-pulse test (MPT) is carried out and a 2000 W small-scale experimental prototype is built to verify the theoretical analysis and suppression method. The experimental results show the proposed suppression methods are effective and the performance of the high-frequency LLC converter can be significantly improved.

Figure 1. The architecture of the high frequency link 35 kV PV system.

Compared with the existing study, this work has some advantages:

- The influence of the parasitic parameters of SiC devices in high-frequency LLC converters is analyzed step by step, which is more accurate.
- The mechanism of influence of parasitic parameters on the resonance process is discussed systematically. The paper reveals how different parameters can affect the resonance process with respect to the parasitic parameters.
- The proposed method is completely different from previous studies, and achieves better performance. Since the proposed method is able to achieve ZVS without a reduction in magnetic inductance and extension of the dead band, the efficiency can be further improved.

This paper is organized as follows. In Section 2, the high-frequency isolated cascaded MVAC grid-connected system is introduced. A downscaled system and its control method are provided. In Section 3, the analysis of influences of parasitic parameters are given, and the conventional suppression method by reducing magnetic inductance is analyzed. In Section 4, the proposed suppression method and its theoretical analysis are given. In Section 5, the theoretical analysis and proposed method are verified using the experimental prototype. Section 6 concludes the paper.

2. Architecture of High Frequency Link 35 kV PV System

The architecture of the presented high frequency link 35 kV PV system is shown in Figure 1. The system can be divided into three stages, including a distributed MPPT stage (non-isolated DC–DC converter), a DC–DC step-up stage (isolated DC–DC converter), and an inverter stage (cascaded inverter). The three-level boost topology is applied in the distributed MPPT stage. The three-level boost is responsible for MPPT, and is directly connected to the PV array. The output terminal of each converter is connected to the DC bus. The isolated DC–DC step-up stage uses a high-frequency LLC converter that works as DCX to provide galvanic isolation and a high step-up ratio. The inverter stage has a cascaded inverter topology. The CHB is used in this stage, which is directly connected to the MVAC grid.

In the presented system, a high-frequency isolated DC–DC converter is used to replace the line frequency transformer. The efficiency and power density can be increased. Additionally, copper consumption can be greatly reduced. The cascaded inverter replaces the centralized inverter. Thus, the switches can operate with lower frequency, lower switching losses and higher conversion efficiency.

The presented system is oriented towards 35 kV large-scale PV power generation. A 6 MW/35 kV PV system is designed based on the presented structure. The parameters of the system are shown in Table 1. To pursue higher efficiency and higher power density, SiC devices are used in the designed system because of their lower conduction loss and better performance under high temperature conditions, even though the switching frequency of CHB is very low. Then, according to the voltage level of the grid and devices, the number of CHB is determined to be 40. For high-frequency LLC converters, SiC devices are also used to achieve higher switching frequency and higher efficiency. In addition, in order to reduce the losses associated with single diodes, several diodes are connected in parallel on the secondary side of the LLC converter. For three-level boost, the devices can be selected to be the same as the LLC converter to reduce the number of different types of device. Besides, since there are a lot of symbols in the paper, all the symbols has been listed in the Appendix C as well.

Table 1. Parameters of designed 6 MW/35 kV system.

Symbol	Description	Parameters
V_{bus}	DC bus voltage	800 V
V_{grid}	Grid voltage	35 kV
P	Power	6 MW
N_{CHB}	Number of CHB in one phase	40
L_g	Inductance of grid side	8 mH

The control scheme of the system is shown in Figure 2. Since an isolated DC–DC converter operates as the DCX, its voltage gain remains unchanged. As long as the output voltage remains stable, the DC bus voltage will be stabilized. Therefore, for controlling the cascaded inverter, the DC side voltage and the AC side power factor are the main control objectives, as shown in Figure 2a.

In Figure 2a, $v_{dc(ki)}$ ($k = a, b, c$ and $i = 1, 2, \dots, n$) is the DC side voltage of unit(i) in phase k . The power reference of each CHB module p_{ki} is obtained by comparing $v_{dc(ki)}$ with the common reference v_{dc}^* . In the presented system, v_{dc}^* remains constant, despite the power variation required for controlling the DC bus voltage. Hence, it is possible to achieve a standard multilevel waveform with an equal and constant voltage level. The PI controller is used to control the power in the dq. When load imbalance or asymmetric faults occur, the grid voltage will be unbalanced. In this case, in addition to decoupling the power control, the cascade inverter also needs to control the balance of the current. In Figure 2a, v_d, v_{d-}, v_q, v_{q-} are the d-axis and q-axis components of the positive and negative sequence of the grid voltage respectively, and i_d, i_{d-}, i_q, i_{q-} are the d-axis and q-axis components of the positive and negative sequence of the grid current. These parameters can be obtained by using the phase-locked loop (PLL) based on the decouple double synchronous reference frames (DDSRF). Then, the three-phase current can be balanced by suppressing negative sequence current by feedforward

negative sequence voltage of grid. For the MPPT stage, non-isolated DC–DC carries out MPPT shown in Figure 2b to improve the utilization of solar energy.

Figure 2. Control block diagram of high-frequency isolated cascaded photovoltaic MVAC grid connected system.

To verify the architecture and control scheme of the system, a downscaled system is built. The waveforms and the control effect of the 40 kW/380 V downscaled system are shown in Figure 3. Figure 3a shows the process of connection to the grid. It can be seen that the cascaded inverter can be connected to the grid quickly. Figure 3b shows the control effect of the control scheme discussed above. It is obvious that the control scheme of the system is able to suppress the negative sequence current and ensure the output current balance.

Figure 3. Waveforms and the control effect of the downsized system.

Although the downscaled system is able to operate well, there are still some problems that need to be solved in the experiment. The influence of parasitic parameters under high step-up ratio conditions is more serious, leading to the loss of ZVS. To achieve higher efficiency in high-frequency situations, it is necessary to study the influence of parasitic parameters and their suppression methods.

3. Influences of Parasitic Parameters on the Resonance Process

3.1. Operation Principle of the Ideal High-Frequency High Step-Up Ratio LLC Converter

The topology, fundamental equivalent circuit, and theoretical waveforms of the ideal LLC converter are shown in Figure 4. In Figure 4a, $M_1 \sim M_4$, $D_1 \sim D_4$ and $C_{oss1} \sim C_{oss8}$ represent the channels, body diodes and parasitic capacitor of the devices, respectively. C_r is the resonant capacitor, L_r is the resonant inductor, L_m is the magnetic inductor, and $D_5 \sim D_8$ are the secondary-side rectifier diodes. The turn ratio of transformer is $1:n$.

Figure 4. Topology, fundamental equivalent circuit and theoretical waveforms of the ideal LLC converter.

When the parasitic capacitance of the secondary side is very small and can be ignored, the operation of the LLC converter can be analyzed as being under ideal conditions. Figure 4c shows the theoretical waveforms of the ideal LLC converter, in which $V_{dri,M1}$ and $V_{dri,M2}$ are the drive signals of M_1 and M_2 , $V_{ds,M1}$ is the drain-source voltage of M_1 , and i_r and i_m represent the resonant current and magnetic current, respectively. At t_1 , as the current drops to zero, the secondary diodes turn off naturally and achieve ZCS because of lower di/dt . Since the magnetic inductance of the ideal LLC converter can be considered to be large enough, i_m remains unchanged and i_r remains the same as i_m . At t_2 , M_1 and M_4 are turned off and i_r start to charge or discharge parasitic capacitors of devices. Because i_r is always maintained at the peak value of i_m , the commutation of primary side can be completed in a very short time. Thus, the achievement of ZVS can be ensured at t_4 .

When the dead band and parasitic capacitance of secondary side are ignored, according to Figure 4b, the DC voltage gain G can be deduced from the fundamental equivalent circuit and can be expressed as:

$$G = \frac{V_{out}}{V_{in}} = \frac{1}{\sqrt{\left(1 + \frac{1}{k} - \frac{1}{k\Omega^2}\right)^2 + Q^2\left(\Omega - \frac{1}{\Omega}\right)^2}}, \quad (1)$$

where V_{out} is the output voltage, V_{in} is the input voltage, k is the ratio of L_m to L_r , Ω is the normalized frequency, the ratio of switching frequency f_s to resonant frequency f_r and Q is the quality factor of the fundamental equivalent circuit, in which R_{eq} is the equivalent resistance of the secondary side of the transformer. The DC voltage gain of LLC has been shown in Figure 5. According to Figure 5, when $f_s > f_r$, there is no ZCS in the secondary side because of the high di/dt ; however, when $f_s < f_r$, the LLC converter will operate in the capacitive zone and there is no ZVS. Therefore, the operation zone shown in Figure 5 is the best choice and $f_s < f_r$. Since the high-frequency ratio LLC converter operates as DCX, the larger k is preferred to resist the frequency disturbance because of its flatter DC voltage gain curve

according to Figure 5b. It is obvious that the gain of the LLC converter changes by less than 10% with the changing of the load, even under the condition that $f_s = 0.9f_r$.

Figure 5. DC voltage gain of the ideal high-frequency LLC converter. (a) Normalized frequency vs. quality factor; (b) Normalized frequency vs. inductance ratio.

3.2. Analysis of the Operation Principle of the LLC Converter Considering Parasitic Parameters

In the practical application of the high-frequency LLC converter, the parasitic capacitor of the secondary diodes cannot be ignored. The topology of the LLC converter in consideration of the parasitic parameters is shown in Figure 6a. $C_{oss5} \sim C_{oss8}$ represent the junction capacitors of the rectifier diodes. The reference direction of i_r , i_m and i_s are the same as the directions shown in Figure 6a.

Figure 6. Topology and theoretical waveforms of the LLC converter considering parasitic parameters.

In general, the LLC converter is designed in consideration of ideal conditions. Nevertheless, with the increase in the magnetic inductance of the LLC converter, the influence of the magnetic current becomes weaker, and the equivalent parasitic parameters of the secondary devices becomes much greater compared with LLC converters in traditional step-down applications. The parasitic parameters have a negative influence on the primary-side ZVS. The theoretical waveforms of the LLC converter considering parasitic parameters are shown in Figure 6c. $V_{dri,M1}$ and $V_{dri,M2}$ are the drive signals of M_1 and M_2 , $V_{ds,M1}$ is the drain-source voltage of M_1 , V_{D5} represents the voltage of the secondary rectifier

diode D_5 , and i_r and i_m represent the resonant current and magnetic current, respectively. In the first half cycle, the LLC converter can be divided into four states, as shown in Figure 7.

Figure 7. Current flow state of each stage of LLC converter considering parasitic parameters.

In Stage3, the parasitic capacitance of both sides participates in the resonance process, as shown in Figure 7c. The equivalent circuit of Stage3 between t_2 and t_3 is shown in Figure 6b. $C_{oss,p}$ and $C_{oss,s}$ represent the parasitic capacitance of the primary and secondary devices, respectively. According to the equivalent circuit, the secondary-side parasitic capacitance was changed by n^2 , and the influence on the resonance process was changed as well. Additionally, with the extensive use of SiC devices, the switching frequency of the LLC converters is continuously improved, and the loss increase caused by the loss of ZVS is more serious.

Since the waveform of the LLC converter is completely symmetrical in every half cycle, only the former half cycle is analyzed.

Stage 1 ($t_0 \sim t_1$): As shown in Figure 7a, both M_1/M_4 and D_5/D_8 are conductive. The voltage of L_m is clamped by the output voltage, and the i_m increases linearly. Because the resonant elements L_r and C_r form a band-pass filter, only the current near the resonant frequency is able to pass through the resonant network. Therefore, i_r varies as a resonant frequency sinusoidal curve.

Stage 2 ($t_1 \sim t_2$): As shown in Figure 7b, M_1 and M_4 are still conductive, while D_5 and D_8 are naturally turned off, because the current drops to zero with a low di/dt . Thus, ZCS is achieved on the secondary side. Then, the secondary-side parasitic capacitors $C_{oss5} \sim C_{oss8}$ are introduced into the resonant network. A higher frequency resonance occurs in both V_{D5} and i_r . In this stage, i_r fluctuates near i_m .

Stage 3 ($t_2 \sim t_3$): As shown in Figure 7c, M_1 and M_4 are turned off and i_r starts to charge or discharge the parasitic capacitors $C_{oss1} \sim C_{oss4}$. In this stage, both $C_{oss1} \sim C_{oss4}$ and $C_{oss5} \sim C_{oss8}$ are involved in the resonance process. This stage is very important, because the resonance process has a great influence on the primary-side ZVS. In this stage, according to the Kirchhoff voltage laws (KVL), i_r should meet:

$$\begin{cases} u_{cq} + u_{cr} + L_r \frac{di_r}{dt} + L_m \frac{di_m}{dt} = 0 \\ L_m \frac{di_m}{dt} = u_{cd} \end{cases} \quad (2)$$

Furthermore,

$$\begin{cases} i_r = C_q \frac{du_{cq}}{dt} = C_r \frac{du_{cr}}{dt} \\ i_m = i_r - C_d \frac{du_{cd}}{dt} \end{cases} \quad (3)$$

where C_q , C_r and C_d are capacitances of $C_{oss1} \sim C_{oss4}$, the resonant capacitor and $C_{oss5} \sim C_{oss8}$, respectively. U_{cq} , u_{cr} and u_{cd} are the voltage of corresponding capacitor, and i_r and i_m represent the current of the resonant inductor L_r and magnetic inductor L_m . The initial values are the final values of Stage 2.

Stage 4 ($t_3 \sim t_4$): As shown in Figure 7d, primary-side commutation is completed and D_2 and D_3 are turned on naturally. At the same time, D_6 and D_7 are turned on as well. The voltage of L_m is clamped by the output voltage, and the i_m increases linearly. i_r varies as a resonant frequency sinusoidal curve. At t_4 , M_2 and M_3 are turned on, and the ZVS is achieved.

Obviously, with the addition of secondary-side parasitic capacitors, the resonance process in Stage 3 is changed dramatically, according to Equations (2) and (3). This change may mean that the ZVS cannot be achieved. Usually, the achievement of ZVS needs to meet:

$$\begin{cases} i_r(t) \geq 0 \\ \int_0^{T_{dead}} i_r(t) dt \geq 2C_q V_{in} \end{cases} \quad (4)$$

where T_{dead} is the dead band of the converter. In Equation (4), the first condition ensures that the voltage of the primary side devices can achieve ZVS, and the second condition ensures that i_r can provide sufficient charge to achieve ZVS in the dead band. Therefore, $i_r(t)$ is very important for the realization of primary-side ZVS. According to Equations (2) and (3), the $i_r(t)$ in Stage 3 can be plotted by MATLAB. Then, the influence of secondary parasitic capacitance on ZVS can be obtained according to Figure 8. Figure 8 presents the change of $V_{ds,M1}$ during the transient with different C_d and L_m or L_r . When $C_d = 0$, the equivalent circuit becomes the circuit of the ideal LLC converter and $V_{ds,M1}$ change linearly.

Figure 8. Voltage of M_1 in Stage 3 with different L_m and L_r .

As shown in Figure 8, with the increase of C_d , i_r is shared by the parasitic capacitors of the secondary side, which increases the time required to fully charge the primary-side capacitors. To ensure that ZVS can be achieved, the dead band must be longer than the time needed to charge C_{oss1} and C_{oss4} . Furthermore, since the switch turns off after i_r is equal to i_m , $i_r(t_2) = i_m(t_2)$ is assumed.

According to Figure 8a, the L_m influences the resonance process and ZVS. It is obvious that the smaller L_m is, the larger initial $i_r(t_2)$ is and the shorter the time needed to achieve ZVS under the same parasitic parameter conditions. However, smaller L_m will lead to larger conduction loss and switching loss. Therefore, a careful tradeoff has to be made.

Additionally, according to Figure 8b, L_r also influences ZVS. The change of L_r leads to the change of resonant frequency and impedance in the resonance process. Then, the time to achieve ZVS is changed. It is obvious that the larger the L_r is, the more serious the influence of parasitic capacitance on the ZVS will be.

The above analysis shows that when parasitic capacitance can be extracted, to achieve ZVS, L_m , L_r and T_{dead} should be carefully designed in the high-frequency LLC converter.

In the traditional design method for LLC converters, to achieve ZVS, L_m is usually limited by:

$$L_m \leq \frac{nV_o T_{dead}}{8V_{in} C_q f_s} \quad (5)$$

However, Equation (5) ignores the secondary-side parasitic capacitors and resonance process in the dead band. According to the above analysis, there are two ways of ensuring the realization of ZVS via careful parameter design. The first is that i_r is always larger than zero. This means that ZVS can be achieved only by adjusting the dead band (of course, the dead band must be within acceptable limits). The second is that the charge or discharge of primary-side parasitic capacitors can be completed before i_r drops to zero. To achieve ZVS, generally the L_m is selected as a small value compared to the theoretical value calculated by Equation (5). However, L_m is an important parameter in high-frequency LLC converters, because it not only affects the realization of soft switching, but also the power loss of the converter. In the LLC converter, the resonant current i_r , magnetic current i_m , primary-side current of transformer i_p and secondary-side current of transformer i_s can be derived as:

$$\begin{cases} i_r(t) = I_r \sin(2\pi f_s t - \varphi) \\ i_m(t) = \begin{cases} \frac{nV_{out}t}{L_m} - I_m & 0 \leq t \leq \frac{1}{2f_s} \\ -\frac{nV_{out}(t - \frac{1}{2f_s})}{L_m} + I_m & \frac{1}{2f_s} < t \leq \frac{1}{f_s} \end{cases} \\ i_p(t) = i_r(t) - i_m(t) \\ i_s(t) = -ni_p(t) \end{cases} \quad (6)$$

$$\begin{cases} I_m = \frac{nV_{out}}{4L_m f_s} \\ I_r = \sqrt{\left(\frac{\pi P}{2nV_{out}}\right)^2 + I_m^2} \\ \varphi = \arcsin\left(\frac{I_m}{I_r}\right) \end{cases} \quad (7)$$

Thus, the conduction loss and switching loss of LLC can be calculated, and the relationship between the total loss P_{loss} and L_m is shown in Figure 9. There is an obvious demarcation point in Figure 9, namely, $L_{m.min}$. When L_m is smaller than $L_{m.min}$, the P_{loss} increases rapidly with the decrease of L_m , no matter at light load condition or at heavy load condition, which will lead to a lower efficiency in the full load range. Therefore, a new method to suppress the influence of parasitic parameters in the LLC converter without efficiency loss is very important to improve the efficiency of the PV system.

Figure 9. Relationship curves between power loss and magnetic inductance in the high-frequency LLC converter.

4. Suppression Method of the Influence of Parasitic Parameters Based on the Saturable Inductor

In the PV system, LLC converters based on SiC devices have a higher switching frequency, which can further improve the system efficiency and the power density of the system. However, parasitic parameters lead to the loss of ZVS, the increase in power loss, and the deterioration of EMI characteristics at high frequencies. Although the methodological analysis above is able to suppress the influence of secondary parasitic parameters by reducing the L_m , the increase in power loss is still inevitable. To ensure the implementation of ZVS and improve the efficiency of the LLC converter, a structure for an LLC converter with a saturable inductor, as shown in Figure 10a, is proposed. In addition, its theoretical waveforms are shown in Figure 10b.

Figure 10. Topology and theoretical waveforms of LLC converter with saturable inductor.

In Figure 10a, L_s is the initial inductance of the saturable inductor, and I_c is the current when L_s becomes saturated. The L_s is added to the secondary side. Since the I_c is very small, the L_s is saturated for most of the time in a cycle, and the circuit structure is the same as that of the traditional LLC converter. When the secondary current i_s is near to zero, the L_s gradually withdraws from the saturated state, and is added into the secondary circuit as an inductor. Therefore, L_s blocks the parasitic capacitance of the secondary side to participate in the resonance process and ensures the achievement of ZVS of the primary side. According to Figure 10b, the operation of the LLC converter with the saturable inductor can be divided into five stages.

4.1. Analysis of the Operation Principle of the LLC Converter with the Saturable Inductor

The five stages of the LLC converter with the saturable inductor are shown in Figure 11.

Figure 11. Current flow state of each stage of the LLC converter with the saturable inductor.

Stage 1 ($t_0 \sim t_1$): As shown in Figure 11a, both M_1/M_4 and D_5/D_8 are conductive. The voltage of L_m is clamped by the output voltage, and the i_m increases linearly. Because the resonant elements L_r and C_r form a band-pass filter, only the current near the resonant frequency is able to pass through the resonant network. Therefore, i_r varies as a resonant frequency sinusoidal curve. In this stage, according to KVL:

$$\begin{cases} V_{in} - u_{cr} - L_r C_r \frac{d^2 u_{cr}}{dt^2} - L_m \frac{di_m}{dt} = 0 \\ L_m \frac{di_m}{dt} = V_o \end{cases} \quad (8)$$

Stage 2 ($t_1 \sim t_2$): As shown in Figure 11b, M_1 and M_4 are still conductive, and $|i_s| < I_c$. Thus, the L_s participates in the operation of the LLC converter and D_5 and D_8 remain conductive. Therefore, i_r decreases slowly and linearly. In this stage, according to KVL:

$$\begin{cases} V_{in} - u_{cr} - L_r \frac{di_r}{dt} - L_m \frac{di_m}{dt} = 0 \\ L_m \frac{di_m}{dt} = V_o + L_s \frac{di_s}{dt} \end{cases} \quad (9)$$

According to Kirchhoff current laws (KCL):

$$i_r = i_m + i_s = C_r \frac{du_{cr}}{dt} \quad (10)$$

Stage 3 ($t_2 \sim t_3$): As shown in Figure 11c, M_1 and M_4 are still conductive, while D_5 and D_8 are still conductive because of the presence of L_s . Then, i_r starts to charge or discharge parasitic capacitors $C_{oss1} \sim C_{oss4}$. At this stage, i_r decreases slowly and C_{oss1} and C_{oss4} are fully charged in a short time. Therefore, according to KVL, i_r meet:

$$\begin{cases} u_{cq} + u_{cr} + L_r \frac{di_r}{dt} + L_m \frac{di_m}{dt} = 0 \\ L_m \frac{di_m}{dt} = V_o + L_s \frac{di_s}{dt} \end{cases} \quad (11)$$

According to KCL and voltage-current relationship (VCR):

$$i_r = i_m + i_s = C_r \frac{du_{cr}}{dt} = C_q \frac{du_{cq}}{dt} \quad (12)$$

Stage 4 ($t_3 \sim t_4$): As shown in Figure 11d, D_2 and D_3 naturally become conductive, because C_{oss2} and C_{oss3} were fully discharged in Stage 3. Then, i_s drops to zero, D_5 and D_8 are turned off and ZCS can be achieved because of lower di/dt . After that, $C_{oss5} \sim C_{oss8}$ participate in the resonance process and are charged or discharged. In this stage, according to KVL:

$$\begin{cases} V_{in} - u_{cr} - L_r \frac{di_r}{dt} - L_m \frac{di_m}{dt} = 0 \\ L_m \frac{di_m}{dt} = u_{cd} + L_s \frac{di_s}{dt} \end{cases} \quad (13)$$

According to KCL and VCR:

$$\begin{aligned} i_r &= i_m + i_s = C_r \frac{du_{cr}}{dt} = C_q \frac{du_{cq}}{dt} \\ i_s &= C_d \frac{du_{cd}}{dt} \end{aligned} \quad (14)$$

Stage 5 ($t_4 \sim t_5$): As shown in Figure 11e, both D_2/D_3 and D_6/D_7 are conductive and L_s becomes saturated. The voltage of L_m is clamped by the output voltage, and the i_m decreases linearly and i_r varies as a resonant frequency sinusoidal curve. At t_5 , the M_2 and M_3 are turned on, and primary side ZVS is achieved. In this stage, according to Figure 11e:

$$\begin{cases} V_{in} + u_{cr} + L_r C_r \frac{d^2 u_{cr}}{dt^2} + L_m \frac{di_m}{dt} = 0 \\ L_m \frac{di_m}{dt} + V_o = 0 \end{cases} \quad (15)$$

Besides, some s-domain analysis has been shown in the Appendix B.

4.2. Analysis and Design of the Saturable Inductor

The inductance and energy storage of the ideal saturable inductor are shown in Figure 12.

Figure 12. Relation curves between the ideal saturable inductor and its current.

In Stage 2 and Stage 3, L_s maintains i_s greater than zero, which causes the rectifier diodes to be continuously conductive. At these two stages, secondary-side parasitic capacitance will not participate in the operation of the circuit. In Stage 3, L_s causes the i_r to remain basically unchanged, which leads to a shorter time for the primary side until commutation. Then, the secondary side voltage of the transformer can be approximated as:

$$V_s \approx V_{in} + V_c + L_r \frac{di_r(t)}{dt} \quad (16)$$

where the V_c is the peak value of the resonant capacitor voltage, so the $i_s(t)$ can be expressed as:

$$i_s(t) = I_c - \frac{V_o - V_s}{L_s} t \quad (17)$$

To ensure the completion of primary-side commutation, i_s should always be greater than zero at t_3 , that is:

$$I_c L_s \geq (V_o - V_s)(t_3 - t_1) \quad (18)$$

In Stage 4, the L_s causes the i_s to be less than zero. After the secondary diodes are turned off and ZCS is achieved, the parasitic capacitors participate in the resonance process. At t_4 , i_s decreases to $-I_c$ and D6 and D7 become conductive. Then, the L_s becomes saturated and no longer participates in the operation of the circuit. In Stage 5, the resonant current $i_r(t)$ can be expressed as:

$$i_r(t) \approx I_m - I_c + \frac{V_o + V_c - V_{in}}{L_r} t \quad (19)$$

where I_m is the peak value of i_m . To ensure that the primary-side body diode is conductive in these two stages, the $i_r(t)$ must be greater than zero; therefore,

$$I_c \leq I_m + \frac{V_o + V_c - V_{in}}{L_r} (t_5 - t_4) \quad (20)$$

According to the constraints of (19) and (21), the values of I_c and L_s for the saturable inductor can be designed.

5. Experimental Verification and Analysis

To verify the influence of parasitic parameters on ZVS, as well as to validate the two suppression methods proposed in this paper, a 2000 W down-scale prototype is built.

Since the system is based on SiC devices, the driving circuit and protection constitute an important technology. In comparison with Si-MOSFET, SiC-MOSFET has some significant differences in terms of its characteristics, due to its different material and structure. SiC-MOSFET has the advantages of low on-state resistance and fast switching speed. However, due to the low gate threshold voltage, it is more likely to suffer from interference and erroneous conduction. Therefore, the driving circuit of SiC-MOSFET cannot be simply replaced by that of Si-MOSFET.

In the LLC converter, SiC devices are mainly used as full-bridge switching devices on the primary side. Usually the switching on and off of SiC devices is affected by their output capacitance under soft-switching conditions, and the crosstalk problem is not serious. However, because the switching speed of SiC devices is very fast and the transient dv/dt is very large, once the ZVS has been lost, the crosstalk problem becomes very serious. Subsequently, the increase in gate voltage caused by crosstalk will be threatening to SiC-MOSFET during this period. Therefore, effective driving and protection circuits are also the focus of the driving design of the SiC devices.

CREE, ROHM and other companies have successively introduced commercial SiC device driving circuits, but few of these circuits involve solutions to driving protection problems. In this paper, C2M0080120D is used as the primary side switch in the experiment. To meet the high reliability requirement of the PV system, short-circuit protection is considered in the design of the driving circuit, as shown in Figure 13. SiC devices are driven by a 24 V isolated power supply, and a BROADCOM ACPL-339J-000E optocoupler driver chip is used to achieve isolation and protection. This chip provides a user-setup desaturation protection circuit scheme, which is able to provide effective protection for the devices and improve system reliability when a short-circuit occurs. In the driving circuit, the gate parallel resistor and capacitor of SiC-MOSFET are able to reduce the influence of crosstalk. In the short-circuit protection circuit, C_{blank} and Z_{DESET} are designed according to the actual needs, whereby C_{blank} determines the allowable blank time of driving protection and Z_{DESET} determines the

corresponding current value of protection. When short-circuit protection is triggered, V_{GMOS} will increase to a high level. The soft turn-off time is determined by the series resistance of Q_3 and the input capacitance C_{iss} of SiC-MOSFET.

Figure 13. SiC-MOSFET driver circuit.

As shown in Figure 14a, the parasitic capacitance of the SiC-MOSFET and SiC diodes is usually nonlinear. Therefore, the time-related equivalent parasitic capacitance $C_{oss(tr)}$ is very important. Generally, the MPT circuit shown in Figure 14b can be used to measure the equivalent parasitic capacitance. Figure 14c shows the hard switching process of the SiC MOSFET. V_{DS} is the drain-source voltage of the device, I_{LOAD} is the load current, I_D is the drain current of the device, and Q_{oss} is the charge caused by C_{oss} . According to the characteristics of the capacitor, the $C_{oss(tr)}$ can be calculated as

$$C_{oss(tr)} = \frac{\int_{t_1}^{t_2} i_D(t) dt}{V_{DS}} \quad (21)$$

where $i_D(t)$ is the expression of drain current between t_1 and t_2 .

Figure 14. Measurements of parasitic capacitance of SiC devices.

The waveforms of the MPT are shown in Figure 15. V_M is the drain-source voltage, i_M is the drain current, i_D is the current of the diode and V_D is the voltage of the diode. The test voltage is 400 V, and the final current is 20 A. According to the results of MPT, the driving circuit was verified. The Q_{oss} was calculated, and the $C_{oss(tr)}$ obtained by (21) was 1.6 nF. Additionally, the switching loss of SiC-MOSFET under hard-switching conditions can be calculated.

Figure 15. Waveforms of MPT of SiC devices.

According to the analysis above, in consideration of parasitic parameters, the parasitic capacitance of primary- and secondary-side devices is very important for the parameter design of high-frequency LLC converters. However, the parasitic capacitance is not a fixed value. It is related to the voltage of the device. Generally, the values of parasitic capacitance are replaced by time-related equivalent capacitance. The equivalent capacitances of C2M0080120D and IDH10G65C5XKSA1, which are used in the experiment, are shown in Table 2. The parameters of the experimental prototype are also listed in Table 2.

Table 2. Experiment parameters of the 2000 W prototype.

Symbol	Description	Parameters		
		Dead Band Extended	Reduce Magnetic Inductance	Add Saturable Inductor
V_{in}	Input voltage	400 V	400 V	400 V
V_o	Output voltage	400 V	400 V	400 V
P_o	Output Power	2000 W	2000 W	2000 W
L_r	Resonant inductance	20 μ H	20 μ H	20 μ H
L_m	Magnetic inductance	300 μ H	80 μ H	300 μ H
C_r	Resonant capacitance	156 nF	156 nF	156 nF
M	Primary side devices	C2M0080120D	C2M0080120D	C2M0080120D
D	Secondary side devices	IDH10G65C5XKSA1 (6 parallel)	IDH10G65C5XKSA1 (6 parallel)	IDH10G65C5XKSA1 (6 parallel)
C_{ossm}	Capacitance of primary side devices	1.6 nF	1.6 nF	1.6 nF
C_{ossd}	Capacitance of secondary side devices	6×40 pF	6×40 pF	6×40 pF
L_s	Inductance of the saturable inductor			300 μ H ($I_c = 1$ A)

Figure 16 shows the waveform of the LLC converter using the traditional magnetic inductance limitation. According to Equation (5), the maximum inductance value of L_m is 500 μ H when the dead band is extended from 500 ns to 700 ns, in accordance with [31]. To ensure that ZVS can be achieved, considering a margin of 1.5 times, the 300 μ H is selected as the experimental inductance in the experiment. In Figure 16, i_r is the resonant current, V_{M1} is the voltage of M_1 , V_{D5} is the voltage of D_5 , and V_{dri} is the driving signal of M_1 . Although the margin is considered in the design, the ZVS still cannot be realized under full-load conditions. Hard switching increases the interference in the driving signals and reduces the efficiency of the converter. In addition, since there is some interference in the driving signals, this experimental method cannot be used in the actual system. In Figure 16, Stage 1~3 correspond to the stages in the theoretical analysis. In Stage 1, i_r is part of a sinusoidal waveform and M_1 has been turned off. At the beginning of Stage 2, i_r begins to oscillate instead of

being maintained near the magnetic current because of the participation of parasitic capacitors of the secondary-side diodes. Since the f_s is only slightly smaller than f_r , this stage is very short. At the beginning of Stage 3, the parasitic parameters of the primary- and secondary-side devices participate in a resonance process together until the dead band ends or the ZVS has been achieved. It is noteworthy that the secondary diodes become conductive before the ZVS is achieved. Then, i_r begins to transmit power to the load, causing the realization of ZVS to become impossible. Therefore, even though the dead band is extended, the ZVS cannot be achieved, as shown in Figure 16.

Figure 16. Waveforms of the LLC converter designed using traditional method with an extended dead band.

Figure 17 shows the waveform of the LLC converter, where the magnetic inductance is reduced according to [31,35]. The parasitic capacitance is measured by MPT, and 80 μH is finally selected as the experimental inductance according to a simulation. In Figure 17, it can be seen that ZVS can be achieved under full-load conditions. Additionally, there is no interference in the driving signal. However, the magnetic current clearly increases because of the reduction of the magnetic inductance, which leads to the increase of conduction loss and switching loss. In Figure 17, Stage 1~4 correspond to the stages in the theoretical analysis. Stage 1 and Stage 2 are similar to the stages in Figure 16. At the beginning of Stage 3, the parasitic parameters of the primary- and secondary-side devices also participate in the resonance process together until the dead band ends or ZVS has been achieved. However, since the magnetic inductance is reduced and the magnetic current is increased, the primary- and secondary-side commutation finish almost simultaneously. Then the load will be supported by source instead of by resonant inductor. Therefore, in Stage 4, the primary side ZVS is achieved.

Figure 18 shows the experimental waveform of the LLC converter with the saturable inductor on the secondary side. 300 μH is selected as the magnetic inductance for comparison. It can be seen that ZVS can be realized under full-load conditions. Additionally, there is no interference in the driving signal. Since the magnetic current has hardly increased, the loss is obviously reduced compared with the method of reducing the magnetic inductance. In Figure 18, Stage 1~5 correspond to the stage in theoretical analysis. In Stage 1, i_r is part of sinusoidal waveform and M_1 has been turned off. As the $|i_r|$ decreases, the saturated inductor gradually exits out of the saturated state, and then the Stage 2 begins. Thereafter, the i_r varies linearly. At the beginning of state 3, the primary device begins to commute, which is different from the two experiments mentioned above. Since the saturable inductor keeps the secondary current constantly below 0, the parasitic capacitance of the secondary-side diodes no

longer p in this stage. Then the problem of the secondary side completing the commutation before the primary side can be avoided. At the beginning of Stage 4, the primary side commutation is completed and ZVS-on is realized. When the current i_s exceeds 0, the secondary diodes begin to commute. If the primary switch is not fully turned on and the secondary commutation is finished, the secondary diodes enter Stage 5. It is obvious the resonant current of this experiment is much lower than that shown in Figure 17. This means that higher efficiency can be obtained.

Figure 17. Waveforms of the LLC converter designed by reducing magnetic inductance.

Figure 18. Waveforms of LLC converter with saturable inductor.

Figure 19 shows the efficiency curves and loss breakdowns of three experiments under full load conditions. It is obvious that the proposed method is able to achieve the highest efficiency across the full load range compared with the others. The full load efficiency of the proposed method is able to achieve 98.6%. Although the efficiency of the experimental parameters designed by traditional method with the extended dead band is only slightly lower than the proposed method, it is still not suitable for practical

application due to its terrible driver conditions, as shown in Figure 16. The converter with reduced magnetic inductance is clearly able to achieve ZVS. However, according to the theoretical analysis shown in Appendix A, from the point of view of power loss, the loss of the primary-side devices is significantly increased. Compared with the former, the method based on the saturable inductor is able to reduce the conduction loss by 56.74% and the switching loss by 73%. Then, the full-load efficiency can be increased by 0.7%. Obviously, the suppression method based on the saturable inductor is better.

Figure 19. Efficiency curves and loss break down of the high-frequency LLC converter in three experiments.

6. Conclusions

This paper presents the architecture of a high frequency link 35 kV PV system and studies the influence of the parasitic parameters of the high-frequency high-voltage isolated DC–DC converter. This paper reveals how the parasitic parameters affect the realization of ZVS of the LLC converter in high-frequency situations, and proposes a method for suppressing the influences of parasitic parameters and promoting the realization of ZVS. The method effectively prevents parasitic elements of the secondary side from participating in the resonance of the primary side by adding a saturable inductor on the secondary side to ensure that ZVS can be achieved.

This paper has some advantages compared to the existing research,

- The influence of the parasitic parameters of SiC devices on the high-frequency LLC converter is analyzed step by step, which is more accurate.
- The mechanism of influence of the parasitic parameters on the resonance process is discussed systematically. This paper reveals how different parameters are able to affect the resonance process with respect to parasitic parameters.
- Compared to the extended dead band, the proposed suppression method is able to cause ZVS to be realized and to eliminate the driving interference.
- Compared to the reduced magnetic inductance, the proposed suppression method can significantly decrease the power loss caused by magnetic inductance. The conduction loss can be reduced by 56.74% and the switching loss by 73%. The efficiency can be improved by 0.7%.

Compared with the conventional method, which involves reducing magnetic inductance, the proposed method does not need to reduce the magnetic inductance, and this is beneficial for reducing the magnetic current of the transformer and improving the efficiency of the LLC converter. The experimental verification shows that the theoretical analysis is consistent with the actual situation. The proposed parasitic parameter suppression method is able to significantly improve the efficiency of the LLC converter. Furthermore, the proposed method is able to improve the efficiency and performance of the MVAC PV system.

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Appendix A

On the basis Equations (6) and (7), the resonant current i_r , magnetic current i_m and secondary side current of transformer i_s can be calculated. Since the drain-source voltage V_{ds} and conduction resistance $R_{ds(on)}$ of the device are known, the conduction loss and switching loss of the devices can therefore be calculated.

Usually, the conduction loss P_c , turn-on loss P_{on} and turn-off loss P_{off} can be calculated by (A1)–(A3), as follows:

$$P_c = I_{drms} R_{ds(on)} \quad (A1)$$

$$P_{on} = f_s \int_0^{t_r} (V_{ds} - \frac{V_{ds}}{t_r} t) \frac{i_d}{t_r} dt \quad (A2)$$

$$P_{off} = f_s \int_0^{t_f} (I_D - \frac{i_d}{t_f} t) \frac{V_{ds}}{t_f} dt \quad (A3)$$

where I_{drms} is the RMS value of the drain current, t_r is the rise time of the current, t_f is the fall time of the current, f_s is the switching current and I_D is the conduction drain current.

Furthermore, loss due to magnetic components is also important. During the design process, it is possible to obtain the maximum magnetic flux density B_m . Then, on the basis of the datasheet of the selected core, it is possible to calculate the core loss P_{core} . Since the parameters of the winding are known, the DC equivalent resistance and AC equivalent resistance can be calculated, as well. Finally, the loss due to magnetic components P_m can be obtained.

Therefore, it is possible to accurately calculate the main losses of the converter by theoretical analysis.

In Figure A1, the s-domain equivalent circuit of each stage of the LLC converter with a saturable inductor is shown. Then, the accurate expression of each stage in the s-domain can be written and derived as follows:

Appendix B

Figure A1. S-domain equivalent circuit of each stage of the LLC converter with a saturable inductor.

$$i_r(s) = \frac{C_r L_r i_r(t_0)s + C_r[V_{in} - V_o + v_c(t_0)]}{C_r L_r s^2 + 1} \quad (A4)$$

$$i_r(s) = \frac{c_1 s + c_0}{c_3 s^2 + c_2} \quad (A5)$$

$$\begin{cases} c_0 = L_s C_r V_{in} + L_s C_r v_c(t_1) - L_s C_r V_o - V_o L_m C_r \\ c_1 = L_s C_r i_r(t_1) L_r + L_s C_r i_s(t_1) L_s + L_s C_r i_m(t_1) L_m \\ \quad + L_s C_r i_m(t_1) L_s + L_s C_r i_s(t_1) L_m \\ c_2 = L_s \\ c_3 = L_s C_r (L_r + L_m + L_s) \end{cases}$$

$$i_r(s) = \frac{d_1 s + d_0}{d_3 s^2 + d_2} \quad (A6)$$

$$\begin{cases} d_0 = L_s C_r C_q V_{in} + L_s C_r C_q v_c(t_2) - L_s C_r C_q V_o - L_m C_r C_q V_o \\ d_1 = L_s C_r C_q i_r(t_2) L_r + L_s C_r C_q i_s(t_2) L_m + L_s C_r C_q i_m(t_2) L_m + L_s^2 C_r C_q i_m(t_2) + L_s^2 C_r C_q i_s(t_2) \\ d_2 = L_s (C_r + C_q) \\ d_3 = L_s C_r C_q (L_r + L_m + L_s) \end{cases}$$

$$i_r(s) = \frac{e_4 s^4 + e_3 s^3 + e_2 s^2 + e_1 s + e_0}{e_7 s^5 + e_6 s^3 + e_5 s} \quad (A7)$$

$$\begin{cases} e_0 = i_m(t_3) C_r \\ e_1 = C_r C_d [v_c(t_3) - V_o - V_{in}] \\ e_2 = C_r C_d [i_r(t_3) L_r + i_s(t_3) L_s + i_s(t_3) L_m + i_m(t_3) L_s] \\ e_3 = L_s C_r C_d^2 [v_c(t_3) - V_o - V_{in}] - L_m C_r C_d^2 V_o \\ e_4 = L_s C_r C_d^2 [i_r(t_3) L_r + i_s(t_3) L_s + i_s(t_3) L_m + i_m(t_3) L_m + i_m(t_3) L_s] \\ e_5 = C_r + C_d \\ e_6 = L_s C_d^2 + L_r C_r C_d + L_m C_r C_d + 2 L_s C_r C_d \\ e_7 = L_s C_r C_d^2 [i_r(t_3) + L_m + L_s] \end{cases}$$

$$i_r(s) = \frac{C_r L_r i_r(t_4)s + C_r[V_o + v_c(t_4) - V_{in}]}{C_r L_r s^2 + 1} \quad (A8)$$

where C_q and C_d are the capacitances of $C_{oss1} \sim C_{oss4}$ and $C_{oss5} \sim C_{oss8}$, respectively; V_o is the equivalent voltage of V_{out} converted to the primary side of transformer; and $i_r(t_n)$, $v_c(t_n)$ and $i_m(t_n)$ represent the values of i_r , v_c and i_m at $t_n (n = 0, 1, 2, 3)$. L_s is the inductance of the saturable inductor, L_r is the resonant inductance, L_m is the magnetic inductance and C_r is the resonant capacitance.

Appendix C

The symbols used in this paper are shown in Table A1.

Table A1. Symbols used in the paper.

Symbol	Description
V_{bus}	DC bus voltage
V_{grid}	Grid voltage
P	Power
N_{CHB}	Number of CHB in one phase
L_g	Inductance of grid side
$v_{dc(ki)}$ ($k = a, b, c$ and $i = 1, 2, \dots, n$)	DC side voltage of unit(i) in phase k
p_{ki}	Power reference of CHB of unit(i) in phase k
v_{dc}^*	DC side voltage reference
v_d, v_{d-}, v_q, v_{q-}	D-axis and q-axis components of positive and negative sequence of grid voltage
i_d, i_{d-}, i_q, i_{q-}	D-axis and q-axis components of positive and negative sequence of grid current
v_a, v_b, v_c	Grid voltage
i_a, i_b, i_c	Grid current
θ	Phase angle of the grid
v_{abc}^*	Three phase voltage reference

Table A1. Cont.

Symbol	Description
$M_1 \sim M_4$	Channels of MOSFETs
$D_1 \sim D_4$	Body diodes of MOSFETs
$D_5 \sim D_8$	Secondary side rectifier diodes
$C_{oss1} \sim C_{oss8}$	Parasitic capacitor of MOSFETs
L_r	Resonant inductance
L_m	Magnetic inductance
C_r	Resonant capacitance
L_s	Inductance of the saturable inductor
n	Turn ratio of transformer
$V_{dri,M1}, V_{dri,M2}$	Drive signals of M_1 and M_2
$V_{ds,M1}$	Drain-source voltage of M_1
V_{D5}	Voltage of secondary rectifier diode
i_r	Resonant current
i_m	Magnetic current
G	DC voltage gain
V_{out}	Output voltage of LLC
V_{in}	Input voltage of LLC
k	Ratio of L_m to L_r
Ω	Normalized frequency
f_s	Switching frequency
f_r	Resonant frequency
Q	Quality factor
$C_{oss5} \sim C_{oss8}$	Junction capacitors of rectifier diodes
$C_{oss,p}, C_{oss,s}$	Parasitic capacitance of primary and secondary devices
C_q, C_r, C_d	Capacitance of $C_{oss1} \sim C_{oss4}$, resonant capacitor and $C_{oss5} \sim C_{oss8}$
u_{cq}, u_{cr}, u_{cd}	Voltage of C_q, C_r, C_d
T_{dead}	Dead band of the converter
V_o	Equivalent output voltage
i_p	Primary side current of transformer
i_s	Secondary side current of transformer
I_m	The amplitude of magnetic current
I_r	The amplitude of resonant current
φ	Phase of resonant current
P_{loss}	Power loss of LLC
$L_{m,min}$	Demarcation point magnetic inductance
L_s	Initial inductance of the saturable inductor
I_c	Value of current when L_s becomes saturated
V_s	Secondary side voltage of transformer
V_c	Peak value of the resonant capacitor voltage
$C_{oss(tr)}$	Time related equivalent parasitic capacitance
V_{DS}	Drain-source voltage
I_{LOAD}	Load current
I_D	Drain current
Q_{oss}	Charge caused by C_{oss}
P_c	Conduction loss
P_{on}	Turn-on loss
P_{off}	Turn-off loss
I_{drms}	RMS value of the drain current
t_r	Rise time of current
t_f	Fall time of current
B_m	Maximum magnetic flux density
P_{core}	Core loss
P_m	Loss of magnetic components

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