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An Improved Partially Interleaved Transformer Structure for High-voltage High-frequency Multiple-output Applications

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Abstract— This paper proposes an improved partially interleaved structure for high-voltage (several kV) high-frequency (several hundred kHz) multiple-output applications. The proposed transformer structure is compared with other typical structures with the leakage inductance, AC capacitance, AC resistance and the ratio of AC/DC resistance taken into consideration. The proposed structure features lower leakage inductance, smaller AC capacitance, lower AC resistance and lower ratio of AC-DC resistance, which is suitable for high-frequency high-efficiency applications. A planar transformer with the proposed structure was built and tested in an LCLC resonant converter, where the input voltage is 40 V, output is 4800 V, switching frequency is 500 kHz, the output power is 288 W and the efficiency is 96.8%, which validates the analysis.

Index Terms— high-voltage, high-frequency, improved partially interleaved structure, multiple-output, planar transformer

I. INTRODUCTION

High-voltage power supplies with multiple outputs are widely used in Travelling-Wave Tube Amplifiers [1] (TWTA s), lasers [2], Magnetic Resonance Imaging (MRI) [3], where several high-voltage outputs are required by the system. Nowadays, the power supplies are moving towards high power density and high efficiency, which is mainly restricted by the magnetic components and semiconductors in high-voltage multiple-output application. The emergence of novel power switches, such as GaN and SiC devices [4, 5, 6] helps to increase the power density and efficiency from the semiconductor aspect. However, the magnetic design, especially the high-voltage transformer design is still a big challenge. The reason can be given as follows. In order to achieve high efficiency, resonant topologies are applied to reduce the switching loss. The structure of a high-voltage planar transformer has a crucial role on the performance of a resonant converter since its parasitic impedances are utilized to form a resonant tank [7, 8, 9]. As for the leakage inductance, the general way to reduce the leakage is using the interleaving structure. However, the parasitic capacitance will increase as the interleaving will lead to more overlapped area. From the aspect of the parasitic capacitance, the common way is to shift the winding to reduce the overlapped area. However, the shifting will result in the increase of the leakage. As a result, the leakage inductance and parasitic capacitance are interrelated and need careful consideration. Taking the LCLC resonant converter as an example (Fig. 1a), the leakage inductance (Lr), magnetizing inductance (Lm) and parasitic capacitance (Cp) of the transformer are utilized in the resonant tank. The 3D structure of a high-voltage planar transformer is shown in Fig. 1b. As for the switching frequency of the LCLC resonant converter, it is lower than 100 kHz, which is mainly limited by the parasitic parameters of the high-voltage transformer [1].

The planar transformer has been investigated and used comprehensively since it was proposed. Previous research investigated different winding configurations [10, 11], the analysis of impedances [12] and the design of the planar transformers [13, 14]. However, these researches mainly focused on low-voltage high-current applications, which is due to the difficulty in designing the adequate electrical insulation. However, with the development of the Polyimide PCB, the capability for providing electrical insulation has improved greatly, with the electric field intensity increasing from 3 kV/mm to 80 kV/mm or higher [15], which contributes to the application of the planar transformer in the high-voltage field. In high-voltage applications, apart from the AC/DC resistance and Lr, the electrical insulation and Cp need further investigation [14]. However, Cp is always ignored in the previous researches, such as LLC resonant converters. This is due to the larger impedance of Cp compared with that of Lm. However, in high-voltage applications, in order to boost the output voltage, there are few turns in the primary side with a large number of turns in the secondary side. As a result, a large Cp exists, the impedance of which is comparable with that of Lm. Therefore, Cp cannot be ignored.

In resonant converters, Cp is utilized as the resonant component, which will affect the efficiency and switching
frequency of the converter. Take the LCLC resonant converter as an example [16], the energy to charge or discharge \( C_p \) is mainly provided by \( L_m \). With higher \( C_p \), the energy required for the charging or discharging will increase, which will increase the conduction loss. In this way, \( C_p \) needs careful design.

This paper focuses on the applications of the planar transform in high-voltage field and manages to do the trade-off design among all the parasitics. An improved partially interleaved transformer structure for high-voltage high-frequency multiple-output applications is proposed. The proposed structure and other typical structures are shown in Section II. In Section III, IV and V, the comparison between the proposed structure and other typical structures is made, with \( L_r \), AC resistance (\( R_{ac} \)), the ratio of AC/DC resistance and AC winding is often connected to a voltage doubler. Each output of

II. WINDING CONFIGURATIONS FOR HIGH-VOLTAGE PLANAR TRANSFORMERS

In high-voltage multiple-output DC-DC converters, as illustrated in Fig. 1, there are always several secondary windings and to further boost the voltage, each secondary winding needs at least two layers, with one layer for the incoming line and the other for the outgoing line. The configurations of secondary windings considered in this work are illustrated in Fig. 2. The parasitic capacitance between two adjacent turns (\( C_{inters} \)) and between two layers (\( C_{intra} \)) are labeled.

The single-layer structure is shown in Fig. 2(a), where all the turns in a secondary winding are in a single layer, while the other layer is only used for the outgoing line. As a result, \( C_{intra} \) is low. However, it is obvious that the layer for the outgoing line is not fully utilized, which results in higher DC resistance.

The two-layer structure is shown in Fig. 2(b). In this structure, half of the turns in a secondary winding are on one layer while other turns are on the other layer. The two layers are fully utilized and the conductors have the largest width compared with the single-layer structure and the non-interleaved structure, which means that the lowest DC resistance can be achieved. However, because of the large overlapped area, the parasitic capacitance may be high.

The non-overlapped structure is shown in Fig. 2(c). In this structure, half of the turns are on one layer while other turns are on the other layer, which is similar to the two-layer structure. Compared with the two-layer structure, the overlapped area of the turns is minimized to reduce the intra-winding parasitic capacitance.

Combing the primary windings with the secondary windings, a range of winding configurations of the planar transformer is derived, as shown in Fig. 3. There is one primary winding with \( m \) turns and \( n \) secondary windings. Fig. 3(a), (c) and (e) show the non-interleaved structures, while Fig. 3(b), (d) and (f) show the partially interleaved structures. In order to ensure the effectiveness of the electrical insulation, a thick insulation is used between the high voltage sides of the primary and secondary windings in the partially interleaved structures. The thickness of this insulation is determined by the insulation material and the voltage between the primary and secondary windings. In addition, the distribution of the parasitic capacitance is also shown in Fig. 3, which will be discussed in Section VIII.

The derivation of the equations, such as the leakage inductance and AC capacitance, are based on the general structure shown in Fig. 3 and can be easily adopted to other cases easily. In addition, an example is also given to validate the analysis. In order to make a fair comparison, in this example, all the six winding structures use the same magnetic core, which is FEE 38/16/25 and the magnetic material is N87 from
TDK. In addition, each structure has one primary winding and six secondary windings, with two turns in the primary winding and twenty turns in each secondary winding. The parameters of the windings are given in Table I.

In the following parts, the parasitic parameters of the winding configurations ($L_r$, the ratio of AC/DC resistance, $R_{ac}$ and $C_p$) will be compared.

### III. Leakage Inductance

In a resonant converter, $L_r$ is always used as a resonant component in the resonant tank and will affect the resonant condition. As a result, it is important to predict $L_r$. In Part A, the $L_r$ of each winding configuration is calculated based on the Magneto-motive Force (MMF). In Part B, the $L_r$ is simulated in Ansys Maxwell. Finally, the calculated results are compared with the simulated results to validate the correctness of the calculations.

#### A. Calculations of leakage inductance

The calculation of $L_r$ is based on the MMF distribution. The magnetizing current is ignored and the ampere-turn of the primary windings is equal to that of the secondary windings. Assuming $i_p$ is applied to the primary winding, the MMF distribution is shown in Fig. 4. It can be seen from Fig. 4 that compared with the corresponding non-interleaved structure, the highest magnetic field intensity is reduced with partially interleaved structures, which means the even distribution of the magnetic field.

Assuming the magnetic field intensity is constant in the horizontal direction, based on the distributions of MMF, $H$ can be calculated by

$$H = \frac{1}{N} \sum_{i=1}^{N} i_H$$

where $w_w$ is the width of the core widow. $\sum_{i=1}^{N} i_H$ is the total current that contributes to MMF. Calculations are carried out by assuming the direction out of the paper is positive while the direction into the paper is negative.

As a result, the magnetic energy, $E_m$, can be derived by the integration of the magnetic energy density over the core window volume

$$E_m = \frac{1}{2} \int_{V} B_0 H^2 dV$$

According to the relation of the transformer leakage inductance, $L_r$, and the magnetic energy

$$E_m = \frac{1}{2} i_p^2 L_r$$

Combining (2) and (3), the $L_r$ referred to the primary side can be calculated by

### TABLE I  PARAMETERS OF HIGH VOLTAGE PLANAR TRANSFORMERS

<table>
<thead>
<tr>
<th>Symbol</th>
<th>Quantity</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>$w_p$</td>
<td>Width of the primary winding</td>
<td>7.27 mm</td>
</tr>
<tr>
<td>$d_p$</td>
<td>Thickness of the primary winding</td>
<td>0.2 mm</td>
</tr>
<tr>
<td>$d_t$</td>
<td>Thickness of the secondary winding</td>
<td>70 µm</td>
</tr>
<tr>
<td>$d_n$</td>
<td>Thickness of the normal insulation</td>
<td>0.3 mm</td>
</tr>
<tr>
<td>$d_s$</td>
<td>Thickness of the thicker insulation</td>
<td>1.6 mm</td>
</tr>
<tr>
<td>$d_{s1}$</td>
<td>The safe distance between the windings and the core</td>
<td>1.7 mm</td>
</tr>
<tr>
<td>$d_{s2}$</td>
<td>The width of the turns in single-layer structure</td>
<td>0.27 mm</td>
</tr>
<tr>
<td>$d_{s3}$</td>
<td>The width of the turns in two-layer structure</td>
<td>0.64 mm</td>
</tr>
<tr>
<td>$d_{s4}$</td>
<td>The width of the turns in non-overlapped structure</td>
<td>0.36 mm</td>
</tr>
<tr>
<td>$P_T$</td>
<td>Nominal power of the transformer</td>
<td>288 W</td>
</tr>
</tbody>
</table>
where $i_p$ is the current excitation applied to the primary winding. It should be pointed out that (4) is only valid considering a linear operation point, far from magnetic saturation.

The winding configurations with the parameters in Table I are used as an example to compare the leakage inductance. The calculation results of $L_r$ are listed in Table II. It may be concluded that with the partially interleaved structures (W2, W4, and W6), compared with the corresponding non-interleaved structures (W1, W3, and W5), the $L_r$ is reduced by half.

### TABLE II

<table>
<thead>
<tr>
<th>Winding configurations</th>
<th>Value (nH)</th>
</tr>
</thead>
<tbody>
<tr>
<td>W1</td>
<td>125.62</td>
</tr>
<tr>
<td>W2</td>
<td>54.54</td>
</tr>
<tr>
<td>W3, W5</td>
<td>137.78</td>
</tr>
<tr>
<td>W4, W6</td>
<td>60.37</td>
</tr>
</tbody>
</table>

It should be pointed out that the analytical calculations are based on the assumption that the current is distributed evenly through the conductors. However, in the range of several hundred kHz, the relative errors are acceptable [17].

### B. Simulation of leakage inductance

In order to validate the accuracy of the calculation results, Finite Element Analysis (FEA) simulations were carried out in Ansys Maxwell with the same current levels applied to the primary and secondary windings as in calculations. The distributions of the magnetic field intensity, $H$, in W1, W2, and W4 at 500 kHz are shown in Fig. 5.

It can be seen from Fig. 5 that the compared with the non-interleaved structure (Fig. 5a), in the partially interleaved structures (Fig. 5b), the highest magnetic field intensity is reduced, which is accordance with the analytical analysis.

Based on the distributions of the magnetic field intensity, the $L_r$ of each winding configuration can be calculated. The $L_r$ in each winding configuration versus frequency is shown in Fig. 6.

It may be concluded that compared with non-interleaved structures (W1, W3, and W5), the $L_r$ of the partially interleaved structures (W2, W4, and W6) are greatly reduced. Among the partially interleaved structures, the single-layer structure (W2) and two-layer structure (W4) share almost the same $L_r$, which is slightly lower than that of W6. W4 and W6 have the same MMF distribution and therefore the analytic formulas predict that they have the same $L_r$. However, because the turns in the two layers of W6 are shifted from one another, coupling between them is lower than in W4 where the turns on all winding layers are in the same radial location, and this results in higher $L_r$ for W6.

### C. Comparison of the calculation results and the simulation results

The comparison of the simulation and the calculated results is shown in Fig. 7.

It can be seen that the calculated results show good agreement with the simulation results, which validates the effectiveness of the calculations. In order to operate at high switching frequency, a lower $L_r$ is required.

In this way, W2 and W4 are suitable for high switching frequency, high power density applications.

### IV. AC RESISTANCE

The $R_{ac}$ is a significant parameter to determine the copper loss of a transformer, especially in high frequency applications,
where the proximity effect and eddy skin effect will lead to the huge increase of the $R_{ac}$. In this section, the total AC/DC resistance referred to the primary side is analyzed. At first, the ratio of AC/DC resistance referred to the primary side is simulated in FEA software to show the high-frequency characteristics. After that, in order to assess the copper loss, the $R_{ac}$ of each winding configuration referred to the primary side is simulated.

A. The ratio of AC/DC resistance

The ratio of AC/DC resistance can be used to reflect the high-frequency characteristics. The simulation results of the ratio of AC/DC resistance are shown in Fig. 8.

It can be seen from Fig. 8 that the AC/DC resistance of the partially interleaved structures (W2, W4 and W6) is lower than that of the non-interleaved structures (W1, W3 and W5). It can be explained as follows. In the non-interleaved structure, higher leakage energy in the window leads to higher proximity effect, which means the conductors in the non-interleaved structures suffer more severe proximity effect.

As for the AC/DC resistance of W2 and W4, it can be seen from Fig. 5 that the distance between two adjacent layers in W2 is doubled compared with that of W4. In this way, the proximity effect is reduced further.

B. AC resistance

Although it is proved that W2 has the lowest ratio of AC and DC resistance, it should be noted that the DC resistance of W4 is lower than that of W2, which means that in the low-frequency range, the AC resistance of W4 is still lower than that of W2. Therefore, it is significant to investigate the absolute value of the $R_{ac}$.

The $R_{ac}$ is simulated in FEA software and the $R_{ac}$ versus frequency is shown in Fig. 9.

It can be seen from Fig. 9 that partially interleaved structures (W2, W4 and W6) have better high-frequency characteristics than the corresponding non-interleaved structures (W1, W3 and W5). In addition, compared with the non-interleaved structure, the corresponding partially interleaved structure has 50% reduction in the $R_{ac}$. However, the critical frequency may vary from case to case and need further investigation according to the specific application.

V. AC capacitance

In the LCC and LCLC resonant converters, the AC capacitance ($C_p$) of a transformer is used as a resonant component in the resonant tank. In the LCLC resonant converters, higher $C_p$ means more circulating energy, which will lead to higher conduction loss. Therefore, it is significant to investigate the winding configurations to find one with lower $C_p$.

The distributions of the $C_p$ in each structure are shown in Fig. 3. $C_{ps1}$ and $C_{ps2}$ are the parasitic capacitances between the primary winding and the secondary windings. $C_{ss}$ is the parasitic capacitance between two secondary windings. $C_{intra}$ is the parasitic capacitance between two layers in a secondary winding. $C_{inter}$ is the parasitic capacitance between two adjacent turns in the same layer.

In this section, taking the voltage doublers into consideration, the $C_p$ of each winding configuration is calculated and the calculated results are validated by the simulation results.

A. Calculation of AC capacitance

The calculation of the $C_p$ referred to the primary side is based on the following method. Assuming $V_p$ is applied to the primary winding, the voltage per turn is $V_p/n_p$. Then the voltage of each turn both in the primary and secondary windings is calculated. The $C_p$ referred to the primary side can be calculated by [14]

$$C_p = \frac{2E_{tot}}{V_p^2}$$  \hspace{1cm} (5)

Where $E_{tot}$ is the total electric energy stored in the window of the magnetic core.
In this case, the electric energy contains four parts: the electric energy stored between the primary winding ($E_{pp}$), the electric energy stored in the secondary windings ($E_{ss}$), the electric energy stored between the primary windings and the secondary windings ($E_{ps}$), and the electric energy stored between two secondary windings ($E_{int}$).

The electric energy in the primary winding is

$$E_{pp} = \frac{1}{2} \sum_{j=1}^{n_p} C_{pp,j} V_{pp,j}^2$$  \hspace{1cm} (6)

Where $C_{pp,j}$ is the parasitic capacitance between two turns in the primary winding, $V_{pp,j}$ is the voltage of the capacitance.

The electric energy stored between the primary windings and the secondary windings can be calculated by

$$E_{ps} = \frac{1}{2} \sum_{j=1}^{n_p} \sum_{k=j+1}^{n_p} C_{ps,jk} V_{ps,jk}^2$$  \hspace{1cm} (7)

Where $C_{ps,jk}$ is the parasitic capacitance between two turns in the primary winding, $V_{ps,jk}$ is the voltage of the capacitance.

The energy stored in a secondary winding contains two parts. The first part is the electric energy stored between two adjacent turns in the same layer, which can be calculated by

$$E_{int\_1} = \frac{1}{2} \sum_{j=1}^{N_i-1} C_{int\_1,j} V_{int\_1,j}^2$$  \hspace{1cm} (8)

Where $N_i$ is the number of the turns in a secondary winding. $C_{int\_1,j}$ is the capacitance between two adjacent turns in the same layer, which can be calculated by the equation for the plane-parallel capacitor. $V_{int\_1,j}$ is the voltage between two adjacent turns, which is $V_p/n_p$.

The second part is the electric energy stored between two layers in a secondary winding, which can be calculated by

$$E_{int\_2} = \frac{1}{2} \sum_{j=1}^{n_{int\_2}} C_{int\_2,j} V_{int\_2,j}^2$$  \hspace{1cm} (9)

Where $C_{int\_2,j}$ is the capacitance between two layers in a secondary winding, which can be calculated by the equation for the plane-parallel capacitor.

The electric energy stored between two adjacent secondary windings can be calculated by

$$E_{ss} = \frac{1}{2} \sum_{j=1}^{n_{ss}-1} C_{ss,j} V_{ss,j}^2$$  \hspace{1cm} (10)

Where $C_{ss,j}$ is the capacitance between two adjacent secondary windings.

As a result, the total electric energy of a system is the sum of the electric energy stored in all the capacitors in the system, which can be calculated by

$$E_{tot} = \sum_{j=1}^{n_p} E_{pp,j} + \sum_{j=1}^{n_p} E_{ps,j} + \sum_{j=1}^{n_p} E_{int\_1,j} + \sum_{j=1}^{n_p} E_{int\_2,j} + \sum_{j=1}^{n_p} E_{ss,j}$$  \hspace{1cm} (11)

Where $i$ is one in the non-interleaved structure and two in the partially interleaved structure.

Substituting (10) into (5), the $C_p$ referred to the primary side can be calculated by

$$C_p = \frac{2}{V_p} \left( \sum_{j=1}^{n_p} E_{pp,j} + \sum_{j=1}^{n_p} E_{ps,j} + \sum_{j=1}^{n_p} E_{int\_1,j} + \sum_{j=1}^{n_p} E_{int\_2,j} + \sum_{j=1}^{n_p} E_{ss,j} \right)$$  \hspace{1cm} (12)

However, in the high-voltage multiple-output applications, as shown in Fig. 1, since the output of each secondary winding is connected to a voltage doubler and in series with the other secondary winding, there is an increasing DC component across the winding capacitance, which is caused by the capacitors ($C_1, C_2, ..., C_n$) in the voltage doubler. Assuming 1V is applied to the primary winding, the voltage of each turn with the DC component in the secondary windings is listed in Table III.

Using the equations (6), (7), (8), (9), (10) and (11), and the AC voltage excitations in Table IV, the $C_p$ referred to the primary winding in each winding configuration is calculated and calculated results are shown in Table V.

The following simulation results will validate the calculated results.

### B. Simulation results of $C_p$

In order to validate the calculated results, the $C_p$ is simulated in the electrostatic solver of Ansys Maxwell with the same voltage boundary conditions applied as in calculations. The distributions of the electric field intensity in $W_1$, $W_2$, and $W_3$ are shown in Fig. 10. For ease of comparison, all the color maps in Fig. 10 are produced to the same scale. The $C_p$ is calculated by the electric field energy. The simulated results are shown in Fig. 11. In order to facilitate comparison, the calculated results are included in Fig. 11 as well.
It can be seen from Fig. 11 that the calculation results are in accordance with the simulation results, which validates the accuracy of the calculations. The \( C_p \) of the partially interleaved structure is slightly higher than that of the corresponding non-interleaved structure because the partially interleaved structure causes extra capacitance between the primary winding and the secondary winding.

Among the three partially interleaved structures (W2, W4 and W6), W4 has the highest \( C_p \), which can be explained by the following two reasons. Firstly, in the two-layer structure, half of the turns are on one layer while the other half are on another neighboring layer. As a result, there is high intra-winding capacitance \( C_{\text{intra}} \), which is shown in Fig. 2. In addition, the overlapped winding area in W4 is the largest. Compared with W4, the \( C_p \) of W6 is much lower, mainly because the overlapped area is reduced. However, the large intra-winding capacitance \( C_{\text{intra}} \) still exists, which mainly contributes to the AC capacitance of W6. W2 has the lowest capacitance among all the partially interleaved structures. All the turns are in one layer while the other layer is used for the outgoing line only. In this way, \( C_{\text{intra}} \) is reduced greatly. As a result, the \( C_p \) of the single-layer structure is reduced.

### VI. COMPARISON OF THE TRANSFORMER DESIGN

#### A. Effect of the transformer design on resonant converter operation

The analysis about the parasitics is summarized in Table VII. In the resonant converters, the parasitic parameters of the transformer will be utilized in the resonant tank and have crucial effect on the performance of the converter.

**Effect of \( L_r \)**. With higher leakage inductance \( (L_r) \), the switching frequency will be reduced, which will lead to the increase in the volume of the passive components, such as the inductors, transformers and the filter capacitors. For high-voltage applications, it is desirable that the transformer has low \( L_r \). As a result, it can be seen from Table VII that the partially interleaved structures (W2, W4 and W6) are suitable for high-frequency applications.

**Effect of \( R_{\text{ac}} \)**. The copper loss is main part of the power loss in a transformer. In order to reduce the copper loss, a winding configuration with low AC resistance \( (R_{\text{ac}}) \) is better than others. It can be seen from Table VII that taking the high-frequency application (higher than 500 kHz) into consideration, W2, W4 and W6 have relative lower \( R_{\text{ac}} \).

**Effect of the \( C_p \)**. As for the parasitic capacitance \( (C_p) \) of a transformer, a larger \( C_p \) will lead to high current to charge and discharge the capacitance and the conduction loss will increase. For low power loss considerations, a winding structure with low \( C_p \) is desirable. As a result, it can be seen from Table VII that W1 and W2, featured with low \( C_p \), are desirable.

As a result, it can be concluded that from different points of view, the partially interleaved structures are suitable for high-frequency applications. The AC capacitance of the proposed transformer is shown in Table V.

### TABLE V

<table>
<thead>
<tr>
<th>Winding</th>
<th>( C_p ) (nF)</th>
</tr>
</thead>
<tbody>
<tr>
<td>W1</td>
<td>6.72</td>
</tr>
<tr>
<td>W2</td>
<td>9.0</td>
</tr>
<tr>
<td>W3</td>
<td>44.83</td>
</tr>
<tr>
<td>W4</td>
<td>45.2</td>
</tr>
<tr>
<td>W5</td>
<td>25.4</td>
</tr>
<tr>
<td>W6</td>
<td>27.3</td>
</tr>
</tbody>
</table>

### Table VI

<table>
<thead>
<tr>
<th>Voltage Excitations</th>
<th>Voltage (kV)</th>
</tr>
</thead>
<tbody>
<tr>
<td>P and S1</td>
<td>1.2</td>
</tr>
<tr>
<td>S1 and S2</td>
<td>1.2</td>
</tr>
<tr>
<td>S2 and S3</td>
<td>1.2</td>
</tr>
<tr>
<td>S3 and S4</td>
<td>1.2</td>
</tr>
<tr>
<td>S4 and S5</td>
<td>1.2</td>
</tr>
<tr>
<td>S5 and S6</td>
<td>7.5</td>
</tr>
</tbody>
</table>

### Table VII

<table>
<thead>
<tr>
<th>Summary of the Parasitics</th>
<th>( L_r ) (nH)</th>
<th>( R_{\text{ac}} ) (mΩ)</th>
<th>( C_p ) (nF)</th>
</tr>
</thead>
<tbody>
<tr>
<td>W1</td>
<td>95.5</td>
<td>18.38</td>
<td>8.72</td>
</tr>
<tr>
<td>W2</td>
<td>46.4</td>
<td>9.6</td>
<td>11.1</td>
</tr>
<tr>
<td>W3</td>
<td>103.6</td>
<td>20</td>
<td>44.83</td>
</tr>
<tr>
<td>W4</td>
<td>44.9</td>
<td>9.14</td>
<td>45.2</td>
</tr>
<tr>
<td>W5</td>
<td>105.0</td>
<td>18.54</td>
<td>25.4</td>
</tr>
<tr>
<td>W6</td>
<td>50.16</td>
<td>9.43</td>
<td>27.3</td>
</tr>
</tbody>
</table>

Fig. 11 AC capacitance
view, the optimal design result will be different, which means the design of the planar transformer needs tradeoff. In this case, taking all the parasitics into consideration, W2, which is featured with low $L_r$, low $C_p$ and low $R_{ac}$ is suitable for high-voltage high-frequency multiple-output applications.

B. Analysis of the partial discharge

Partial discharge is a big issue in high-voltage transformers. In this paper, as the partially interleaved structure is applied, it is necessary to analyze the effect of the partially interleaving on the partial discharge.

Comparison of the partially interleaving and non-interleaving is carried out between W1 and its corresponding partial interleaved structure W2. The voltage of each turn under the real working conditions is applied to the corresponding turn, where the input voltage is 40 V, the highest output voltage is 4800 V. The distributions of the electric field intensity are shown in Fig. 12. It should be noted that Fig. 10 shows the AC electric field distribution only while Fig. 12 shows the magnitude of the combined DC and AC electric fields.

It can be seen from Fig. 12 that the highest electric field intensity of W2 ($5.84 \times 10^6 \text{ V/m}$) is slightly higher than that of W1 ($4.84 \times 10^6 \text{ V/m}$), which is caused by the partially interleaving. However, the highest electric field can be reduced by the increase of the insulation.

As a result, it can be concluded that the partially interleaving will increase the partial discharge. However, the increase in partial discharge can be optimized by increasing the thickness of the insulation.

VII. EXPERIMENTAL VALIDATION

A. Planar transformer for high-voltage high-frequency applications

Based on the above discussion about the comparison of the transformer design, a planar transformer with the partially-interleaved single-layer structure for a 40 V input, 4800 V output LCLC resonant converter was built. The planar transformer uses the same parameters as listed in Table I.

The shape of the core for the planar transformer is FEE 38/16/25 and the material is N87. The planar transformer has one primary winding and six secondary windings. The primary winding has two turns. In order to achieve a partially interleaved structure, one turn is placed at the top of the PCB (P1) while the other is at the bottom (P2). The six secondary windings (S1, S2, S3, S4, S5 and S6) are located between the primary winding and each secondary winding has twenty turns with the single-layer structure. As the output voltage increases from S1 at the top to S6 at the bottom, which means the voltage of S1 is the lowest while that of S6 is the highest, in order to satisfy the requirement of the high-voltage insulation, a thicker insulation is added between secondary winding (S6) and the bottom primary winding (P2).

B. Testing of the planar transformer

The planar transformer was tested with a HP 4940 impedance analyzer. The test results are shown in Fig. 13.

The $L_r$ is measured from the primary winding with the secondary windings short circuit. It can be seen from Fig. 13(a) that $L_r$ is $54.9 \text{ nH}$, which shows small relative error compared with the calculation result ($54.54 \text{ nH}$) and the simulation result ($48.04 \text{ nH}$). $L_m$ is $3.9 \mu\text{H}$.

When the secondary side is open circuit, as $L_r$ is much lower than that $L_m$, the resonant frequency between $L_m$ and $C_p$ is measured by the impedance analyzer as shown in Fig. 13(b):

$$f_r = \frac{1}{2\pi \sqrt{L_m C_p}}$$  \hspace{1cm} (13)

Where $f_r$ is the resonant frequency, which can be read from the measured result impedance versus frequency results under open-circuit conditions.

Therefore, $C_p$ of the planar transformer is found to be $C_p = 13.2 \text{ nF}$ (14)
The measured $C_p$ shows acceptable relative error compared with the simulation result (11.1 nF) and the calculation result (9.0 nF).

For high-voltage applications, the effectiveness of the high voltage insulation must be tested. The test is carried out between each of the adjacent windings (P and S1, S1 and S2, S2 and S3, S3 and S4, S4 and S5, S5 and S6, P and S6). The voltage excitations applied to the windings are listed in Table VI. The test of the electrical insulation between P and S6 (highest voltage, 7.5 kV) is shown in Fig. 14.

It can be seen from the current meter that the leakage current is zero, which proves the effectiveness of the electrical insulation.

C. Testing of the planar transformer in an LCLC resonant converter

The LCLC resonant converter with fixed-frequency fixed-duty cycle as a DC transformer is widely used in the second stage of a two-stage converter [16]. There are four resonant elements in the resonant tank: $L_r$, $C_s$, $L_m$ and $C_p$. All the transformer parasitics are incorporated into the resonant tank and only $C_p$ is needed.

The LCLC resonant converter with the planar transformer is shown in Fig. 15. The input voltage is 40 V, the output of each secondary winding is connected to a voltage doubler to produce 800 V. The outputs of the six voltage doublers are connected in series. As a result, each output of a voltage doubler can be used for a separate load and the maximum output voltage is 4800 V.

The measured waveforms of the resonant current $i_r(t)$, the voltage and the driving signal of $S_1$ in the LCLC resonant converter with the planar transformer are shown in Fig. 16, where it can be seen that the switching frequency of the converter is 500 kHz.

It can be seen from Fig. 16(a) that when $S_1$ is turned on, the voltage of $S_1$ is zero and the resonant current is also zero, therefore, $S_1$ operations with Zero-Voltage-Switching (ZVS) and Zero-Current-Switching (ZCS). The switching loss is minimized. The voltage and current of the rectifier diode $D_1$ is shown in Fig. 16(b). ZCS and ZVS are also achieved, which will reduce the switching loss.

D. Loss breakdown of the LCLC resonant converter

In this part, the power loss of the LCLC resonant converter is derived based on circuit simulation, including the driving loss ($P_{s_dr}$), conduction loss of the switches ($P_{s_on}$), copper loss ($P_{T_{Cu}}$), core loss ($P_{T_{Fe}}$) and dielectric loss ($P_{T_{Die}}$) of the planar transformer and the rectifier loss ($P_D$).

As for the power loss of the high voltage planar transformer, in addition to the conventional copper loss and core loss, the dielectric loss is also considered in this case. The dielectric loss is caused by the charging and discharging process of the parasitic capacitance. In this case, the power loss factor $\tan \delta$ of FR4 is 0.01, according to the definition of the power loss factor

$$\tan \delta = 2\pi f_s C_p R_s$$

Where $R_s$ is the equivalent series resistance, $f_s$ is the switching frequency of the capacitance, $C_p$ is the value of the capacitance.

The dielectric loss ($P_{T_{Die}}$) can be calculated by

$$P_{T_{Die}} = \frac{I_{c_{rms}}^2}{R_s}$$

Where $I_{c_{rms}}$ is the RMS current across the parasitic capacitance, which can be derived from the circuit simulation.

The rectifier loss is calculated by the forward voltage ($V_F$) times the average current ($I_{av}$);

$$P_D = V_F I_{av}$$

The total loss ($P_{tot}$) can be calculated by

$$P_{tot} = P_{s_{dr}} + P_{s_{on}} + P_{T_{Cu}} + P_{T_{Fe}} + P_{T_{Die}} + P_D$$

The power loss of each part is derived from circuit
simulation and the loss breakdown is shown in Fig. 18. The simulated $P_{tot}$ is 9.24 W while the measured $P_{tot}$ is 9.44 W. It can be concluded that the simulated $P_{tot}$ is in accordance with the measured result.

The power loss of the switches (driving loss and conduction loss) account for 19%, which is relatively high in the total power loss. However, this can be improved by replacing the Silicon MOSFETs by GaN MOSFETs. The power loss of the transformer accounts for 64%, which is the highest in the total loss. As for the dielectric loss, the Polyimide PCB with lower power loss factor, which will contribute to the reduction of the equivalent series resistance, can be used.

According to Table VII, when other winding configurations are compared with $W_5$, the non-interleaved structures ($W_s$, $W_3$, and $W_5$) will limit the high switching frequency due to higher leakage inductance compared with the partially interleaved structures ($W_2$, $W_4$, and $W_5$). In addition, compared with $W_2$, $W_4$, and $W_5$, they will suffer from higher conduction loss and copper loss due to high AC capacitance.

VIII. CONCLUSIONS

A partially-interleaved single-layer structure for high-voltage high-frequency multiple-output applications is proposed in this paper. The parasitic capacitance is considered and this transformer structure manages to do the trade-off design among all the parasitic parameters. Based on leakage inductance, AC/DC resistance and AC capacitance, this structure is compared with other five typical structures. Simulations and calculations show that both the leakage inductance and the AC/DC resistance are reduced. In addition, this structure shares almost the same AC capacitance with the non-interleaved single-layer structure. Besides, a planar transformer with the partially-interleaved single-layer structure is built and tested with the LCLC resonant topology. Compared with the state of the art in TWTA application, by careful design of a planar transformer, leakage inductance is reduced while maintaining inter-winding capacitance within values that ensure ZVS and ZCS, with the result that the switching frequency is increased from less than 100 kHz to 500 kHz without any decrease in efficiency.

Therefore, based on the above analysis, it may be concluded that the partially-interleaved single-layer structure proposed in this paper is suitable for high-voltage high-frequency multiple output applications.

It should be noted that the analysis in this paper considered a particular application, however, due to the improvement in leakage inductance, AC capacitance and AC resistance, the case of partial interleaving would provide similar benefits in other high voltage applications having high step-up and isolation requirements.

REFERENCES


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