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Published in:
IEEE Transactions on Power Electronics

Link to article, DOI:
10.1109/TPEL.2018.2842029

Publication date:
2018

Document Version
Peer reviewed version

Link back to DTU Orbit

Citation (APA):
High Frequency LLC Resonant Converter with Magnetic Shunt Integrated Planar Transformer

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Abstract—Achieving high efficiency and high power density is emerging as a goal in many power electronics applications. LLC resonant converter has been proved as an excellent candidate to achieve this goal. To achieve smaller size of passive components, the resonant inductor in the LLC converter is usually integrated into the transformer by utilizing its leakage inductance. However, the leakage inductance of transformer is usually insufficient and thus the LLC converter has to be operated in a limited frequency range (this limits the input voltage range accordingly), otherwise the power efficiency will drop dramatically. Therefore, a larger resonant inductance in the LLC converter is expected to operate in a wider input voltage range. This paper proposes a new method to create a larger resonant inductance by using a magnetic shunt integrated into planar windings. The accurate leakage inductance modelling, calculation and optimal design guideline for LLC planar transformer, including optimal magnetic shunt selection and winding layout, are presented. A 280-380V input and output 48V-100W half bridge LLC resonant converter with 1MHz resonant frequency is built to verify the design methodology. A comparison is made between two converters with the same parameters, one using magnetic shunt integrated transformer and the others using traditional planar transformer and external inductor. Experimental result shows the proposed converter with magnetic shunt is capable to achieve comparable high efficiency and regulation capability with the other under a wide input voltage, which verify the optimal design methodology. Above all, this magnetics integration methodology reduces the whole converter’s volume and thus increases the power density.

Index Terms—magnetic shunt, LLC resonant converter, planar transformer, high frequency, wide input voltage.

I. INTRODUCTION

There has been a long desire for converter miniaturization to increase its power density and efficiency. Resonant converters are capable to achieve soft switching and operate at high frequency [1]-[3]. LLC resonant converter has been proved as an excellent candidate to satisfy these requirements by increasing its switching frequency and integrating its magnetic components. It has become a popular converter and been used in many applications [4]-[11].

It is well understood that there is a strong trade-off between the input-voltage and/or output-voltage range and the power conversion efficiency. This trade-off is more severe in resonant converters than in pulse-width-modulated (PWM) converters. LLC resonant converters regulate the voltage by changing the switching frequency. They suffer from progressively increased losses when the input voltage is widened so that the typical LLC converters are not able to maintain high efficiency across the entire range when input voltage is wide. Some publications focus on decreasing magnetizing inductance to obtain high voltage gain [12]-[14]. The literature [12] proposes a transformer structure with auxiliary small inductance connected in parallel with the magnetizing inductance and controlled by a switch. The switch turns on to decrease the equivalent magnetizing inductance so as to obtain the high voltage gain. In literature [13], a capacitor is connected in series with the magnetizing inductance. When the switching frequency drops, the equivalent magnetizing inductance decreases so that the peak voltage gain increases. The other approach with auxiliary LC resonant circuit is proposed in [14] and is utilized to compensate the voltage gain. Although these methods [12]-[14] show high performance by optimal design methodology, they all require additional passive devices or power stage and none of them tests their methodology in high frequency-MHz operation.

An alternative to obtain high voltage gain with narrowed operating frequency range can be achieved by designing a large resonant inductor to create a low inductance ratio of magnetizing inductance to resonant inductance. If discrete magnetic components are used for LLC resonant converter, the large inductor may require extra winding loss and core loss, as well as large size of the component [15]-[20]. Therefore the resonant inductor is usually integrated into the transformer by utilizing its leakage inductance to increase its power density and efficiency especially when LLC converter operates at high frequency [25]-[28]. Interleaved structure is often implemented to reduce AC resistance [21]-[24], but causing a small leakage inductance and a large inductance ratio and leading to reduced regulation capability. When the primary and secondary windings are not interleaved [25]-[26], the leakage inductance can be utilized to replace the resonant inductance. However, the leakage inductance is still insufficient and hard to deal with wide input voltage.

This paper is to create a large leakage inductance of planar transformer by inserting a magnetic shunt into the transformer. The relation between the leakage inductance and the characteristic of magnetic shunt is investigated in [30].

Partial work is from the papers published in IECON 2017 [30] and APEC 2018 [31]. Authors are all from Technical University of Denmark, Kgs. Lyngby, 2800 Denmark.
Separating the primary and secondary windings away from the magnetic shunt further improve the leakage inductance while the winding loss due to the fringing effect is also minimized. The proposed integrated transformer structure and design methodology maintain the optimal magnetizing inductance while increasing leakage inductance to obtain a narrowed switching frequency band to deal with wide input voltage range. Thus high power efficiency is ensured with an optimal magnetizing inductance and high regulation capability is achieved with a small inductance ratio.

This paper is organized as follows. Section II analyzes the resonant tank design for the LLC resonant converter with high regulation capability. Section III elaborates the calculation and analysis of the leakage and magnetizing inductance for the proposed transformer structure. Section IV gives detailed design considerations for the proposed transformer structure and high performance LLC converter. In section V, a converter prototype with 1MHz 100W is built to demonstrate the proposed transformer structure, and a comparison with the prototype with 1MHz 100W is built to demonstrate the proposed transformer structure. Section VI concludes this paper.

![Topology of half bridge LLC resonant converter][31]

**Fig.1.** Topology of half bridge LLC resonant converter[31]

**II. LLC RESONANT TANK DESIGN**

![LLC resonant converter waveforms at resonant frequency][1]

**Fig.2.** LLC resonant converter waveforms at resonant frequency

Fig.1 shows the half bridge LLC converter topology. Waveforms at the resonance are illustrated in Fig.2. When Q1 turns on, the magnetizing inductance \( L_{m} \) is charged and the magnetizing current \( I_{Lm} \) linearly grows until Q1 turns off at \( t_{1} \). The drain-to-source voltage of Q2 \( V_{ds,Q2} \) drops to zero from \( t_{1} \) to \( t_{2} \) to achieve ZVS turn-on, where the output capacitance of Q2 is fully discharged by the magnetizing current.

It is known that LLC resonant converter can achieve maximum efficiency at resonant frequency, because there is no circulating energy loss. However, to achieve high voltage gain, the switching frequency of LLC resonant converter usually swings below the resonant frequency. The converter with higher regulation capability can operate at narrowed switching frequency band, which gives lower circulating energy loss and high efficiency compared to the converter with a lower regulation capability. The regulation capability is mainly affected by the inductance ratio of the magnetizing inductance \( L_{m} \) to the resonant inductance \( L_{r} \). This relation is illustrated in Fig.3. The following description in this section gives a method to select the reasonable inductance ratio based on a specified voltage gain.

![Voltage gain vs. normalized frequency][2]

**Fig.3.** \( L_{n} \) impacts on LLC converter voltage gain

Resonant frequency \( f_{r} \) formed by resonant inductor \( L_{r} \) and resonant capacitor \( C_{r} \), quality factor \( Q \) and inductance ratio \( L_{n} \) are important parameters to be considered:

\[
f_{r} = \frac{1}{2\pi \sqrt{L_{r}C_{r}}} \tag{1}
\]

\[
Q = \frac{\sqrt{L_{r}C_{r}}}{R_{r}}, \quad \text{where} \quad R_{r} = \frac{8\pi^{2}R}{\pi^{2}} \tag{2}
\]

\[
L_{n} = \frac{L_{m}}{L_{r}} \tag{3}
\]

Assume optimal magnetizing inductance is obtained. The real quality factor varies with \( L_{n} \) is expressed in (4) by combining (1), (2) and (3).

\[
Q_{\text{real}} = \frac{L_{m}}{L_{r}} \times \frac{2\pi f_{r}}{R_{r}} \tag{4}
\]

The maximum quality factor \( Q_{\text{max}} \) that allows the required maximum voltage gain at the boundary between capacitive and inductive mode is indicated in [27]

\[
Q_{\text{max}} = \frac{1}{L_{n}M_{\text{max}}^{2}} \times \sqrt{\frac{M_{\text{max}}^{2}}{M_{\text{max}}^{2} - 1}} \tag{5}
\]
For a given specification, the maximum voltage gain $M_{\text{max}}$ is specified. Thus the maximum quality factor is only varied with the inductance ratio $L_n$. This is also seen in minimum operating frequency that allows the converter to achieve the maximum voltage gain, which is expressed by

$$f_{n \_ \text{min}} = \frac{1}{\sqrt{1 + L_n \left(1 - 1/M_{\text{max}}^2\right)}}$$  \hspace{1cm} (6)

Based on (4), (5) and (6), the maximum quality factor $Q_{\text{max}}$, the real quality factor $Q_{\text{real}}$, and the minimum operating frequency $f_{n \_ \text{min}}$ as a function of the inductance ratio are plotted in Fig. 4. The real quality factor must be always below the maximum quality factor. The selected inductance ratio should be always larger than $L_{n \_ \text{min}}$, in order to allow the $Q_{\text{real}}$ to be smaller than the $Q_{\text{max}}$. Nevertheless, a larger $L_n$ allows a wider switching frequency range for a given voltage gain, and thus the regulation capability of the converter becomes weaker. Taken together, the inductance ratio $L_n$ should be designed carefully to achieve ZVS in a full load range and obtain a narrowed switching frequency range. A small inductance ratio is expected to achieve high regulation capability. This implies a large resonant inductance $L_r$ is needed in the assumption of unchanged magnetizing inductance $L_m$. The large resonant inductance may require extra windings and magnetic cores. This may cause additional power loss and increase the converter’s volume. This paper integrates the large resonant inductance into the transformer by using magnetic shunt. The detailed analysis is shown as follows.

### III. PROPOSED INTEGRATED TRANSFORMER STRUCTURE

The proposed transformer structure is shown in Fig. 5. Magnetic shunt is inserted in between the two magnetic cores. Both primary and secondary windings are located away from the magnetic shunt with controllable distances $x_p$ and $x_s$, respectively. The air gaps $l_{ap}$ and $l_{as}$ are inserted in between the magnetic cores and the magnetic shunt to get a desired magnetizing inductance, while the gaps have negligible effect on leakage inductance. The magnetizing inductance and leakage inductance in this configuration are modelled and analyzed in the following description.

**A. Leakage energy stored in window area $E_{\text{air}}$**

The magnetomotive force (MMF) in each layer of the primary windings is $F_p = k_p F_p$, where $k_p$ is the number of turns on each layer of primary windings. $N_p$ is the number of layers for primary winding. The MMF within the window area $E_{\text{air}}$ can be assumed to be $N_p F_p$. The magnetic field intensity within the air area $E_{\text{air}}$ is

$$H_{\text{air}} = \frac{N_p F_p}{b_w}$$  \hspace{1cm} (8)

where $b_w$ is the window width. Then the energy stored in the air is

$$E_{\text{air}} = \frac{1}{2} \mu_0 l_w \cdot b_w \int H^2 dx$$  \hspace{1cm} (9)

where $l_w$ is the winding mean turn length (MTL). Applying (8) into (9), the energy stored in the air is

$$E_{\text{air}} = 2 \cdot \frac{1}{2} \mu_0 l_w \int N_p^2 k_p^2 I_p^2 (x_p + x_s)$$  \hspace{1cm} (10)

where $x_p$ and $x_s$ are the distances from the primary and secondary windings to the shunt, respectively.

![Fig.5. transformer with air gaps and both primary and secondary windings away from the shunt](image-url)

The leakage energy contains the energy stored within the primary and secondary windings, the window air between windings, and the magnetic shunt. The leakage inductance is obtained by

$$E = \frac{1}{2} V \cdot B \times H \cdot dV = \frac{1}{2} L_k \cdot I_p^2$$  \hspace{1cm} (7)

where $V$ is the total effective volume; $B$ is the flux density; $H$ is the magnetic field, $L_k$ is the leakage inductance and $I_p$ is the primary current. The leakage energy stored in different elements can be calculated individually with the following sections.
B. Leakage energy stored in primary and secondary windings

Similarly, the leakage energy stored in the primary and secondary windings including the energy stored in the dielectric layers can be calculated from the MMF distribution [29]-[33], which are

\[ E_{pri} = \frac{1}{6} \mu_0 \frac{l_m}{b_w} k_p^2 \left[ h_{gp} \left( 2N_p^3 - 3N_p^2 + N_p \right) + 2h_p N_p^3 \right] I_p^2 \]  

\[ E_{sec} = \frac{1}{6} \mu_0 \frac{l_m}{b_w} k_p^2 \left[ h_{gs} \left( 2N_s^3 - 3N_s^2 + N_s \right) + 2h_s N_s^3 \right] I_s^2 \]

where \( h_{gp} \) and \( h_{gs} \) are the thickness of the dielectric layer of the primary winding and the secondary winding respectively.

![3D structure of the proposed transformer](image)

Fig.6. 3D structure of the proposed transformer

![Magnetic reluctance model of the shunt-inserted planar transformer with air gaps](image)

Fig.7. Magnetic reluctance model of the shunt-inserted planar transformer with air gaps [31]

C. Leakage inductance created by magnetic shunt

The reluctance model is utilized to obtain the leakage inductance created by the magnetic shunt. Fig.6 shows the structure of the planar transformer, and the magnetic reluctance model is illustrated in Fig.7. The air gaps are created in each leg in between the magnetic cores and the magnetic shunt. The reluctances of the sectional magnetic cores \( R_{c1}, R_{c2}, R_{cc} \), the reluctances of the sectional magnetic shunts \( R_{a1}, R_{a2}, R_{a}\) and the reluctances of the air gaps \( R_{gp1}, R_{gp2}, R_{gs1}, R_{gs2} \) indicated in Fig.6 are specified by

\[ R_{c1} = \frac{l_1}{\mu_0 \cdot \mu_s \cdot b_d \cdot l_w} \], \[ R_{c2} = \frac{l_2}{\mu_0 \cdot \mu_s \cdot b_d \cdot l_w} \]

\[ R_{cc} = \frac{l_1}{\mu_0 \cdot \mu_s \cdot A_c} \], \[ R_{a1} = \frac{l_{sh}}{2\mu_0 \cdot \mu_s \cdot b_d \cdot l_w} \]

\[ R_{a2} = \frac{l_{sh}}{\mu_0 \cdot \mu_s \cdot A_c} \], \[ R_{a} = \frac{l_{sh}}{2\mu_0 \cdot \mu_s \cdot A_c} \]

\[ R_{gp1} = \frac{l_{gp1}}{\mu_0 \cdot \left( b_d + l_{gp1} \right) \cdot \left( l_{w} + l_{gp1} \right)} \], \[ R_{gp2} = \frac{l_{gp2}}{\mu_0 \cdot \left( b_d + l_{gp2} \right) \cdot \left( l_{w} + l_{gp2} \right)} \]

\[ R_{gs1} = \frac{l_{gs1}}{\mu_0 \cdot \left( b_d + l_{gs1} \right) \cdot \left( l_{w} + l_{gs1} \right)} \], \[ R_{gs2} = \frac{l_{gs2}}{\mu_0 \cdot \left( b_d + l_{gs2} \right) \cdot \left( l_{w} + l_{gs2} \right)} \]

where \( \mu_0 \) is the permeability of the air; \( \mu_s \) and \( \mu_t \) are the relative permeability of the core and the shunt, respectively. \( A_c \) is the effective cross-sectional area of the core. Other quantities can be found in Fig.6. Considering the fringing effect, the literature [36] points out an effective cross section of air gap with dimensions \( a \) by \( b \) would become \( (a+g) \) by \( (b+g) \), \( g \) is the length of the air gap. This consideration is included in the expression of \( R_{gp1}, R_{gs1}, R_{gs2}, R_{gs2} \).

The expression for the leakage inductance calculation based on the equivalent reluctance models shown in Fig.7 is very bulky. But it can be simplified when the air gap length \( l_{gp} \) is identical to the \( l_{as} \).

\[ l_{gp} = l_{as} = a \]  

(14)

It leads to

\[ R_{gp1} = R_{gs1} = R_{g1} \], \[ R_{gp2} = R_{gs2} = R_{g2} \]  

(15)

Define \( R_m \) as

\[ R_m = R_{c1} + R_{c2} + 2R_{cc} + R_{a1} + 2R_{a2} \]  

(16)

The leakage inductance due to the magnetic shunt referred to the primary side is

\[ L_{k_{shunt}} = \frac{4k_p^2 N_p^2}{R_m + R_{c1} + 2R_{cc} + 2R_{a2}} \]  

(17)

Combining (7), (10), (11), (12) and (17), the total leakage inductance referred to the primary side for the above shunt-inserted transformer is

\[ L_{k} = 2\mu_0 \frac{n_p^2}{b_w} \left( x_p + x_s \right) + \frac{4n_p^2}{R_m + R_{c1} + 2R_{cc} + 2R_{a2}} \]

\[ + \frac{1}{3} \mu_0 \frac{l_m}{b_w} k_p^2 \sum_{i=p,s} \left[ 2h_i N_i^3 + h_{sh} \left( 2N_i^3 - 3N_i^2 + N_i \right) \right] \]

(18)

where \( n_p = k_p \cdot N_p \) is total number of turns for primary winding.

IV. DESIGN CONSIDERATIONS

A. Verification of the proposed calculation method

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Transformer</th>
</tr>
</thead>
<tbody>
<tr>
<td>Turns per layer in primary ( k_p )</td>
<td>2</td>
</tr>
<tr>
<td>Primary layers ( N_p )</td>
<td>4</td>
</tr>
<tr>
<td>Primary conductor thickness ( h_p )(mm)</td>
<td>0.07</td>
</tr>
<tr>
<td>Insulation thickness in primary ( h_{sh} )(mm)</td>
<td>0.2</td>
</tr>
<tr>
<td>Turns per layer in secondary ( k_s )</td>
<td>1</td>
</tr>
<tr>
<td>Secondary layers ( N_s )</td>
<td>2</td>
</tr>
<tr>
<td>Secondary conductor thickness ( h_s )(mm)</td>
<td>0.07</td>
</tr>
<tr>
<td>Insulation thickness in secondary ( h_{sh} )(mm)</td>
<td>0.2</td>
</tr>
</tbody>
</table>

2-D Finite Element Analysis (FEA) simulations are carried out to verify the correctness of the equation (18). Details of
planar transformers parameters are given in TABLE I. The air
gap length and the magnetic shunt thickness are kept constant,
which are $2l_{sh}=0.2\text{mm}$ and $t_{sh}=0.2\text{mm}$. Fig.8 shows comparison
between a the lumped calculation and the FEA simulation for
the leakage inductance with different relative permeability of
the shunt: 1) $\mu_{s}=10$; 2) $\mu_{s}=30$; 3) $\mu_{s}=60$; 4) $\mu_{s}=120$; 5) $\mu_{s}=200$.
The comparison between the proposed calculation method and
the FEA simulation for the leakage inductance in the proposed
configuration is shown in Fig.8. The calculation method for the
shunt-inserted planar transformer represents good agreement
with the FEA simulation.

Three experiment tests were performed by using Agilent
4294A at 1MHz to further prove the calculation. The thickness
and the relative permeability of the magnetic shunt, the distance
from the primary winding and the secondary winding to the
shunt, and the air gap length are documented in TABLE II while
other parameters are kept same with TABLE I. Secondary
windings were connected in a short circuit when measuring the
leakage inductance referred to the primary side. The comparison between the calculation and the experiment tests is
shown in Fig.9. Comparing the test 1 to the test 2, it is observed
that the distance in between the primary and secondary
windings to the shunt can create additional leakage inductance.
Comparing the test 2 to the test 3, the leakage inductance is
increased with the increased shunt thickness.

In high frequency (above 1MHz) LLC resonant converter
design, typically the value of resonant inductance is not very
large. Therefore the magnetic shunt with smaller relative
permeability is more preferable. Moreover, both primary and
secondary windings should be arranged away from the air gaps
to minimize the fringing effect so as to achieve a smaller AC
resistance.
Once the magnetic core is chosen, only five parameters can affect magnetizing inductance and leakage inductance: the relative permeability of the magnetic shunt $\mu_s$ and its thickness $t_{sh}$, the air gap length $l_{a}$, the distances $x_p$ and $x_s$. From the expression of (18) and (19), the distances $x_p$ and $x_s$ only influence the leakage inductance $L_{leak}$. Fig.10 shows the influence of the relative permeability $\mu_s$ and air gap length $l_{a}$ on magnetizing and leakage inductances. As illustrated in Fig.10(a), $\mu_s$ can influence the leakage inductance much more than the magnetizing inductance. On the contrary, the magnetizing inductance is mainly affected by the air gap length, which has a small effect on the leakage inductance as shown in Fig.10(b).

![graph](image)

Fig.11. Investigation of AC resistance at 1MHz as a function of the distance from the shunt to windings for different $\mu_s$.

However, both primary and secondary windings must be placed away from the shunt to avoid fringing effect due to the air gap. This means that the distances $x_p$ and $x_s$ to separate windings and the shunt should be as far as possible. This is illustrated in Fig.11. 2D-FEA simulations were conducted to obtain the ratio of AC resistance to DC resistance at 1MHz. The transformer and winding details are the same with those in TABLE I. The relative permeability of the shunt varies from 50 to 200 and the thicker the core is, the smaller the AC resistance. It can be seen that the AC resistance is decreasing as the distances $x_p$ and $x_s$ increase, and a larger $\mu_s$ creates a larger AC resistance. Fortunately, the value of resonant inductance is not very large at high frequency LLC resonant converters. Thus the magnetic shunt with smaller relative permeability is more preferable.

**C. High efficiency LLC resonant converter design considerations**

**TABLE III. Circuit Specifications**

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Values</th>
</tr>
</thead>
<tbody>
<tr>
<td>Input voltage $V_{in}$</td>
<td>280-380V</td>
</tr>
<tr>
<td>Output voltage $V_o$</td>
<td>48V</td>
</tr>
<tr>
<td>Output power $P_o$</td>
<td>100W</td>
</tr>
</tbody>
</table>

The LLC converter specifications are shown in TABLE III. The key part of LLC resonant converter is the resonant tank, which includes the resonant capacitor $C_r$, resonant inductor $L_r$ and the magnetizing inductance $L_m$. In this design, the resonant inductor $L_r$ is integrated into the shunt-inserted planar transformer. The magnetizing inductance is always related to the primary and secondary RMS currents. In other words, it is well connected with the power loss of LLC converter. The optimal magnetizing inductance can be obtained from the analysis of power loss.

(a) Conduction and Switching Losses on Semi-conductors

Primary and secondary RMS currents are given in [34].

$$\begin{align*}
I_{RMS_{pri}} &= \frac{1}{4\sqrt{2}} \frac{V_o}{nR_t} \left( \frac{n^2 R_t^2 T_o^2}{L_m^2} + 4\pi^2 + \frac{16\pi^2 (T_s t_d + t_d^2)}{T_s^2} \right) \\
I_{RMS_{sec}} &= \frac{\sqrt{3}}{24\pi} \frac{V_o}{R_L} \left( \frac{L_m^2 (T_s + 2t_d)}{T_s + 2t_d} + \frac{12\pi^2 T_s}{T_s (T_s + 2t_d)} \right)
\end{align*}$$

(20) (21)

where $V_o$ is the output voltage; $n$ is the number of turns; $R_t$ is the load resistance; $T_o$ is the dead time; $T_s$ is the switching period at resonant frequency. From these two equations, the conduction losses for primary and secondary devices are described as follows:

$$P_{c,pri,RMS} = 2I_{RMS_{pri}}^2 R_{on}$$

(22)

$$P_{c,sec,RMS} = 2I_{RMS_{sec}}^2 R_{on}$$

(23)

where $R_{on}$ are the drain-to-source resistance of primary devices and secondary devices, respectively. Note that the expression (23) is only applicable when synchronous rectification is applied to secondary devices.

Since LLC converter can achieve ZVS turn-on, only turn-off loss $P_{off}$ for the primary devices switching loss is taken into account [34].

$$P_{off} = \frac{V_{in}}{3 \left( \frac{V_o}{4nL_{min} f_s} - C_{pri,off} \frac{V_{in}}{T_{off}} \right) T_{off} f_s}$$

(24)

where $T_{off}$ is the turn-off time; $C_{pri,off}$ is the output capacitance of primary devices.

(b) Transformer losses

Planar transformer is suitable for high frequency converters due to its very low profile and excellent thermal characteristics [37]. Transformer losses consist of core loss and winding loss. The peak flux density is calculated following by the Faraday’s law.

$$B_p = \frac{n V_o (T_s / 2 - T_d)}{2 N_p A_c}$$

(25)

Considering hysteresis loss and eddy current loss, the core loss is calculated by $P_{core} = P_{HC} + P_{EC}$, where $V_o$ is the volume of the core and $P_{HC}$ is calculated by Steinmetz’s equation: $P_{HC} = K_c f^\alpha B_p^\beta$. These three quantities $K_c$, $\alpha$ and $\beta$ can be calculated from relationship curve between unit volume core loss and flux density given in the datasheet of magnetic material. Thus the core loss can be obtained by

$$P_{core} = K_c f^\alpha \left( \frac{n V_o (T_s / 2 - T_d)}{2 N_p A_c} \right)^\beta V_o$$

(26)

Dowell’s model is the common method to calculate AC
winding loss, but its accuracy is affected by the fringing effect due to the air gap. The accurate winding loss can be estimated from 2-D FEA simulations [38]. The copper thickness is selected to be 2oz according to the skin effect at 1MHz.

(c) Magnetic shunt loss

The magnetic shunt loss limits the maximum power level. Losses within the magnetic shunt can be approximately estimated by a complex permeability $\mu''$. The expression for the average rate of energy loss is [39]

$$ P_{\text{shunt-av}} = \frac{1}{2} \mu' \omega h_0^2 $$

where $\omega$ is the angular frequency; $h_0$ is the peak value of the applied magnetic field. The complex permeability $\mu''$ of the magnetic shunt can be found from the datasheet. It can be seen that losses within magnetic shunt is dependent on the operating frequency, applied magnetic field and material complex permeability.

(d) Driving losses

As the converter operating at high frequency, the driving losses may not be neglected. It contains the losses for driving primary switches and secondary synchronous rectifiers,

$$ P_{dr} = 2Q_{gs} V_{gs} f_s + 2Q_{gs_{sr}} V_{gs_{sr}} f_s $$

where $V_{gs}$ and $V_{gs_{sr}}$, $Q_{gs}$ and $Q_{gs_{sr}}$ represent the driving voltages and the gate-source charge for the primary and secondary devices, respectively.

Therefore, the total power loss of the half bridge LLC resonant converter is

$$ P_{\text{total}} = P_{p_{on}} + P_{p_{off}} + P_{sec} + P_{winding} + P_{dr} + P_{\text{shunt-av}} $$

It can be concluded that the magnetizing inductance should be maximized as possible to reduce power loss. Meanwhile, as the magnetizing current should discharge all parasitic capacitances during the dead time, it needs to be smaller enough to allow sufficient magnetizing current to discharge all parasitic capacitances. Thus the magnetizing inductance should be designed to be the largest value that guarantees ZVS. According to [35],

$$ L_m = \frac{t_d \left( T_s / 2 - t_d \right)}{4 \left( 2C_{\text{pri-out}} + 2C_{\text{sec-out}} / n^2 + C_w \right)} $$

where $C_{\text{sec-out}}$ is the secondary device charge equivalent output capacitance, $C_w$ is the transformer winding capacitance referred to the primary side. The equation (30) gives an optimal magnetizing inductance expressed by the dead time. To determine the optimal magnetizing inductance, the total power loss as a function of the dead time, is shown in Fig.12. The optimal $T_d$ can be selected in according with the minimum loss point and then $L_m$ can be determined by the equation (30) with the selected dead time.

V. EXPERIMENTAL IMPLEMENTATION

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Values</th>
</tr>
</thead>
<tbody>
<tr>
<td>Magnetizing inductance $L_m$</td>
<td>26μH</td>
</tr>
<tr>
<td>Resonant inductance $L_r$</td>
<td>3.5μH</td>
</tr>
<tr>
<td>Resonant capacitance $C_r$</td>
<td>7.23nF</td>
</tr>
<tr>
<td>Turns-ratio $n$</td>
<td>8:2:2</td>
</tr>
<tr>
<td>Primary devices</td>
<td>GS66504B</td>
</tr>
<tr>
<td>Secondary devices</td>
<td>BSZ240N12NS3</td>
</tr>
<tr>
<td>Core type and material</td>
<td>E32/6/20-3F46</td>
</tr>
</tbody>
</table>

Fig.12. Total power loss as a function of dead time[31]

Fig.13. Prototype of 1MHz 100W LLC resonant converter

A 1 MHz, 100W half bridge LLC resonant converter is built, as shown in Fig.13. Detailed circuit parameters and the
The proposed planar transformer parameters are listed in TABLE IV and TABLE V, respectively. In this design, the resonant inductor is integrated into the planar transformer by using its leakage inductance created by the magnetic shunt. The material for the magnetic shunt used here is IFL04. The selected inductance ratio is 7.4 to narrow the switching frequency band. The magnetic core used in the transformer is E32/6/20-3F46 and the 8:2:2 turns ratio is used to achieve optimal magnetizing and leakage inductance. GS66504B from GaN system is selected for the primary devices due to its low effective output capacitance and on-state resistance compared with the silicon devices.

Waveforms of the converter operating at resonant frequency (1 MHz) with 380V input and 48V output is shown in Fig. 14(a). To guarantee GaN devices in a safe operation, the output gate signal from gate driver has negative turn-off voltage. It can be seen that the ZVS turn-on is achieved because the drain-to-source voltage drops to zero and then the gate turns on immediately. Thus the reverse conduction through GaN devices is greatly minimized. Fig. 14(b) shows the waveforms of the converter operating at 700 kHz with 280V input and 48V-100W output, and the ZVS operation is also achieved.

Due to the small inductance ratio, switching frequency range is greatly narrowed. Switching frequency swings from 700 kHz to 1MHz to accommodate input voltage changing from 280V to 380V. To prove the LLC converter with integrated magnetic shunt transformer can achieve comparable performance with the one discrete magnetic component, the converter using planar transformer with external inductor is built for the comparison. To conduct a fair comparison, the discrete version has the same parameters and components with the converter with integrated magnetic shunt. The external inductor is made by PCB planar inductor with EI 18/4/10 core shape with 0.18mm air gaps situated in each leg and 3F46 Ferrite material, and 3 turns with one turn each layer. The resonant inductance is the sum of the transformer leakage inductance and external inductance created by the PCB planar inductor. Some waveforms are captured and shown in Fig.15. Fig. 16 shows the photos of the two prototypes and the integrated converter has nearly 20% size reduction compared to the discrete converter.

Efficiency curves of the two converters under different input voltages and load conditions are illustrated in Fig.17. The peak efficiency of the LLC converter with magnetic shunt integrated transformer at resonant frequency is around 95%. When input voltage decreases to 280V, the switching frequency decreases to 700 kHz to achieve the corresponding voltage gain and the efficiency in this case drops around 1.5% at full power. It can also be observed that both cases can achieve almost the same efficiency under the entire load conditions. This verifies the optimal design guideline for the proposed transformer. It can not only regulate a wider voltage range with the low inductance ratio due to the large leakage inductance created by the
magnetic shunt, but also increase power density of the whole converter because of this magnetic integration method. Thus the proposed transformer is well adapted to LLC resonant converter. The loss breakdown of the converter with shunt integrated transformer at 1MHz with 380V input and 48V-100W output is shown in Fig.18. The transformer loss is dominating, which is about 45% of the total loss. The GaN devices and secondary MOSFETs take 38% of the total loss. Loss from the magnetic shunt is around 0.8W and accounts for 17% of total loss.

**VI. Conclusion**

In this paper, a planar transformer structure with a magnetic shunt is proposed to create a large leakage inductance. The leakage inductance calculation methodology is verified by 2-D FEA simulations and experiment tests. Detailed design considerations for the proposed transformer and high performance LLC resonant converter are given. The prototype proves that the transformer structure is well adapted to integrate a larger resonant inductor into the transformer of LLC resonant converter. Moreover, the proposed transformer structure and design methodology maintain an optimal magnetizing inductance while increasing leakage inductance to obtain a narrowed switching frequency band, so that high efficiency is achieved under wide input voltage range. Experiment results show the converter with the proposed integrated transformer obtains the same regulation capability and comparable efficiency with the converter using discrete magnetic components. This verifies the optimal design guideline for the proposed transformer. Moreover, the power density is increased due to this magnetic integration method. Thus the proposed transformer is well adapted to LLC resonant converter. The application of magnetic shunt can be further extended to other topologies since it can integrate the series inductor connected with transformer, for example, phase-shifted full bridge converters (PSFB), and dual active bridge converters (DBA).

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