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物理設計と実装に関する考慮事項

著者

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引用

Active and Passive Electronic Components, 2016, 027406; 2016

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Research Article
Considerations of Physical Design and Implementation for 5 MHz-100 W LLC Resonant DC-DC Converters

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Received 3 June 2016; Accepted 29 August 2016

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Recently, high power-density, high power-efficiency, and wide regulation range isolated DC-DC converters have been required. This paper presents considerations of physical design and implementation for wide regulation range MHz-level LLC resonant DC-DC converters. The circuit parameters are designed with 3–5 MHz-level switching frequency. Also, the physical parameters and the size of the planar transformer are optimized by using derived equations and finite element method (FEM) with Maxwell 3D. Some experiments are done with prototype LLC resonant DC-DC converter using gallium nitride high electron mobility transistors (GaN-HEMTs); the input voltage is 42–53 V, the reference output voltage is 12 V, the load current is 8 A, the maximum switching frequency is about 5 MHz, the total volume of the circuit is 4.1 cm³, and the power density of the prototype converter is 24.4 W/cc.

1. Introduction

Recently, high power-density and high power-efficiency isolated DC-DC converters have been required in information and communication technology (ICT) facilities. Therefore, high switching frequency and low power loss technique have been studied. In particular, to achieve these requirements, the isolated DC-DC converters with GaN-HEMTs have been studied by some research institute in recent years [1–7].

Furthermore, the regulated MHz-level operated isolated DC-DC converters have been studied with wide input voltage: 60–120 V input voltage, 20 V output voltage, 2.6 MHz switching frequency, and 30 W output power, series resonant DC-DC converter [8]; 36–72 V input voltage, 12 V output voltage, 5 MHz switching frequency, and 30 W output power, flyback DC-DC converter [9]; 110–250 V input voltage, 15 V output voltage, 2 MHz switching frequency, and 45 W output power, quasi-square-wave zero-voltage switching three-level half-bridge DC-DC converter [10]; 60–120 V input voltage, 15 V output voltage, 3–4.5 MHz switching frequency, and 40 W output power, zero-voltage switching (ZVS) half-bridge DC-DC converter [11]. These isolated DC-DC converters achieve high power efficiency. However, higher power efficiency and higher output power are required in MHz-level frequency isolated DC-DC converter.

LLC resonant DC-DC converters have been known to achieve high power efficiency with high switching frequency. In case of this topology, the regulated converters of 1 MHz-level switching frequency have been studied so far [12, 13]. Also, the unregulated LLC resonant DC-DC converters operated at 5 MHz have been considered so far [14]. In order to achieve the miniaturization of the converter, LLC resonant DC-DC converters with wide input voltage and high power efficiency are needed at higher switching frequency.

In this paper, the designs of high power-density and wide regulated LLC resonant DC-DC converter have been considered. The circuit parameters have been calculated for wider regulation at MHz-level switching frequency. The target input
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Transformer

\[
\begin{align*}
+ & - \\
\end{align*}
\]

\[L_{\text{add}}\]

\[\alpha: 1: 1\]

\[\alpha: 1\]

\[\ell_{12} = \ell_{21}\]

\[L_{11}\]

\[L_{22}\]

\[\ell_{m}\]

\[\ell_{\text{lk1}}\]

\[\text{Figure 1: LLC resonant DC-DC converter circuit topology.}\]

Table 1: The target specification of the prototype DC-DC converter.

<table>
<thead>
<tr>
<th>Specification</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Input voltage range: (V_i)</td>
<td>42–53 V (48 V)</td>
</tr>
<tr>
<td>Reference output voltage: (V_o)</td>
<td>12 V</td>
</tr>
<tr>
<td>Output power: (P_o)</td>
<td>100 W</td>
</tr>
<tr>
<td>Power efficiency: (\eta)</td>
<td>More than 90%</td>
</tr>
<tr>
<td>Volume</td>
<td>3.3 cm³</td>
</tr>
<tr>
<td>Power density</td>
<td>30 W/cc</td>
</tr>
</tbody>
</table>

Voltage is 42–53 V, output voltage is 12 V, output power is 100 W, and power efficiency is more than 90%. This design procedure includes planar transformer design with FEM of Maxwell 3D. To evaluate the performance of the prototype board, the experiment has been conducted with open loop.

In Section 2, the consideration of LLC resonant DC-DC converter circuit parameter is described. In Section 3, the physical design considerations of planar transformer are revealed. In Section 4, the experimental results are demonstrated.

2. The Consideration of Circuit Parameter Design

The circuit topology is based on a LLC resonant DC-DC converter, as shown in Figure 1. The target specification is shown in Table 1. The primary side is a half-bridge topology. \(Q_1\) and \(Q_2\) are driven in 50% duty ratio, alternatively. \(C_{r1}\) and \(C_{r2}\) are the resonant capacitances which have the same capacitance. Also, they make averaged voltage of \(C_{r2}\) to one half of the input. The magnetic transformer is composed of leakage inductance \(L_{\text{lk1}}\) and magnetizing inductance \(L_m\). \(L_{\text{add}}\) is primary-side additional inductance. The secondary side is the center-tap topology with diodes \(D_1\) and \(D_2\). To achieve target input voltage range, transformer turn ratio is set to be 3.

LLC resonant DC-DC converters have two resonant frequencies. One is the series resonant frequency \(f_0\) which is series resonance between resonant capacitance \(C_r\) and inductance \(L_r\). The other is the resonant frequency \(f_1\) which is resonance between \(C_r\), \(L_r\), and magnetizing inductance \(L_m\). LLC resonant DC-DC converter can operate in the region of above resonance \(f_s > f_0\) or below resonance \(f_1 < f_s < f_0\). In case of above resonance, ZVS of primary-side switches can be achieved. In this region, the converter behaves as buck-mode operation. In case of below resonance, ZVS of primary-side switches and zero-current switching (ZCS) of secondary-side diodes can be achieved. In this region, the converter behaves as boost-mode operation. LLC resonant DC-DC converter used in this paper is operated in below resonance at rated current.

In order to achieve the target specification, the design of LLC resonant DC-DC converter is considered. At first, the equivalent circuit of the magnetic transformer can be drawn, as shown in Figure 2 [15]. Also, the following equations can be obtained:

\[
\begin{align*}
L_{\text{lk1}} &= (1 - k^2) L_{11}, \\
L_m &= k^2 L_{11}, \\
\alpha &= kn = k \sqrt{\frac{L_{11}}{L_{22}}},
\end{align*}
\]

where \(L_{\text{lk1}}\) is the leakage inductance, \(L_m\) is the magnetizing inductance, \(k\) is the coupling coefficient, \(L_{11}\) is the primary-side self-inductance, \(L_{22}\) is the secondary-side self-inductance, and \(n\) is the turn ratio.
In LLC resonant DC-DC converter, for achieving ZVS operation of primary-side switches, the magnetizing inductance has to be designed small enough. The magnetizing inductance can result in the following equation [16, 17]:

\[ L_m \leq \frac{T_s t_{\text{dead}}}{16C_{\text{oss}}}, \tag{2} \]

where \( T_s \) is the switching period, \( t_{\text{dead}} \) is the dead time of the primary-side switches, and \( C_{\text{oss}} \) is the output capacitance of the primary-side switches. From this equation, when the parameters are set as \( T_s = 250\,\text{ns} \), \( t_{\text{dead}} = 10\,\text{ns} \), and \( C_{\text{oss}} = 654\,\text{pF} \) at \( V_i = 42\,\text{V} \), the value of \( L_m \leq 238.9\,\text{nH} \) can be derived. Therefore, the magnetizing inductance can be set at 220nH.

Also, LLC resonant DC-DC converter has been analyzed with fundamental harmonic approximation (FHA) [16, 17]. The definitions for this analysis are as follows:

\[ \omega_s = 2\pi f_s, \tag{3} \]
\[ \omega_0 = 2\pi f_0 = \frac{1}{\sqrt{L_r C_r}}, \tag{4} \]
\[ F = \frac{f_s}{f_0} = \frac{\omega_s}{\omega_0}, \tag{5} \]
\[ Q = \frac{Z_0}{R_{\text{ac}}} = \frac{1}{R_{\text{ac}}} \sqrt{\frac{L_r}{C_r}}, \tag{6} \]
\[ L_{11} = L_{\text{lik1}} + L_m, \tag{7} \]
\[ L_r = L_{\text{add}} + L_{\text{lik1}}, \tag{8} \]
\[ K = \frac{L_m}{L_r}, \tag{9} \]
\[ C_r = C_{r1} + C_{r2}, \tag{10} \]
\[ R_{\text{ac}} = \frac{8}{\pi^2} R_L, \tag{11} \]

where \( f_s \) is the switching frequency, \( f_0 \) is the resonant frequency, \( Q \) is the quality factor of the circuit, \( R_{\text{ac}} \) is the equivalent resistance, \( C_r \) is the resonant capacitance, \( L_r \) is the resonant inductance, \( L_{\text{add}} \) is the additional inductance for optimizing \( L_r \), and \( K \) is the ratio of the inductances. The equivalent circuit of LLC resonant DC-DC converter can be drawn, as shown in Figure 3. From this equivalent circuit, the ratio of the input and output voltage \( M \) can be obtained as follows:

\[ M = \frac{V_{\text{ac}}}{V_s} = \frac{2\alpha V_{\text{ac}}}{V_i} = \frac{1}{\sqrt{(1 + (1/K)(1 - 1/F^2))^2 + Q^2 (F - 1/F)^2}}. \tag{12} \]

When \( k \) is set to 0.94, \( \alpha \) is decided at 2.82. The minimum input voltage is 42 V; thus, the required peak ratio \( M_p \) can be derived about 2 from (12). This result includes 20% margin of \( M \). Figure 4 can result from (12). The limitation of \( Q \) in each \( K \) can be found from this graph. B82801B manufactured by EPCOS is used as not only current measurement but also additional inductance. The short inductance of the current transformer is 7nH. Therefore, from Figure 5, when \( K = 6 \), \( L_r \) is decided about 36.7nH. Also, from Figure 4, \( Q < 0.23 \). Then, \( C_r \) and \( f_0 \) can be decided from (6) and (4), respectively; \( C_r = 12.66\,\text{nF} \) and \( f_0 = 7.4\,\text{MHz} \).

### 3. The Physical Design Considerations of Planar Transformer

In general, Nickel Zinc (NiZn) ferrite core has been used in MHz-level switching frequency operation. In terms of the core of this prototype isolated DC-DC converter, SY22 NiZn ferrite core manufactured by TDK is used [18]. The relative permeability \( \mu_r \) of the core material is 80. This core material can be used from 5 to 15 MHz. The good features are guaranteed in MHz-level frequency. The core shape is EI.
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core. In this paper, the physical parameters of this EI core are optimized for obtaining good performance.

For the prototype LLC resonant DC-DC converter, 8-layer printed circuit board (PCB) is used. The winding arrangement of the prototype planar transformer is shown in Figure 6. In terms of planar transformer, layers 1, 2, 7, and 8 (L1, 2, 7, and 6) are assigned to primary-side windings, while layers 3, 4, 5, and 6 (L3, 4, 5, and 6) are assigned to secondary-side windings. The primary-side turn number is 3 turns, 1 layer has 2 turns, and parallel number is 2. Also, the secondary-side turn number is 1 turn, 1 layer has 1 turn, and parallel number is 2. Each diode is connected with each secondary-side winding, as shown in Figure 1.

Figure 7 shows the definitions for each physical parameter of magnetic transformer which have EI core. From these definitions, the following equation can be obtained:

\[
\begin{align*}
w_2 &= 2w_1 + g, \\
p &= w_2 + 2g, \\
b &= \frac{l_g}{2} - h - p, \\
c &= 2(b + p), \\
V_e &= ab \cdot 2(b + p + h) = A_e l_e,
\end{align*}
\]  

where \(l_g\) is effective magnetic path length, \(A_e\) is effective core cross-sectional area, and \(V_e\) is volume of core material. In addition, based on magnetic circuit of EI core transformer, the following equation can be derived:

\[
A_e = \frac{(C_c L_{11} \left(2\mu \mu_0 N_p^2 + l_e\right))}{(\mu_0 \mu_s N_p^2)},
\]  

where \(C_c\) is correction coefficient, \(\mu_0\) is absolute permeability, \(l_g\) is air gap, \(\mu_s\) is relative permeability of core material, and \(N_p\) is number of primary-side turns.

The restriction of this converter is considered. The physical features of each component are shown in Figure 8. Target depth, width, height, and volume are written in this figure. From these physical features, the limitation of physical parameters in magnetic transformer can be derived as follows: \(0 < a < 15 - 2p\) and \(c < 16.5\).

The relationship between \(B_m\) and temperature can be drawn from past experimental data, as shown in Figure 9. \(B_m\) can be derived as the following equation from experimental results:

\[
B_m = \frac{(r_{fm} r_m)}{(2N_p A_e)}.
\]  

From this data, when the allowable temperature of magnetic core is set to 80°C, \(B_m\) is restricted from 32 to 36 mT.

From Figures 10(a) and 10(b), the length of the winding in the primary-side and secondary-side can be calculated as follows:

\[
\begin{align*}
l_p &= 4a + 3.5b + 15w_1 + 21.5g, \\
l_s &= 2a + b + 8w_2 + 8g.
\end{align*}
\]  

Each layer of resistance is shown here:

\[
\begin{align*}
R_{lay,i} &= r_{lay,i} l_p, \\
R_{lay,j} &= r_{lay,j} l_s,
\end{align*}
\]  

where \(i = 1, 2, 7, 8\) and \(j = 3, 4, 5, 6\). Also, \(r_{lay,i}\) and \(r_{lay,j}\) are winding resistance per unit length. The winding resistance per unit length \(r_{lay}\) is obtained from simplified Maxwell 3D winding model, as shown in Figure 11. The solver of the simulation is eddy current. The thickness of L1 and L8 is 0.073 mm, that of L2 and L7 is 0.055 mm, and that of L3–6 is
Figure 7: The definitions for each physical parameter of magnetic transformer.

Figure 8: The physical features of each component.

Figure 9: $B_m$ versus temperature derived by past experimental data (core: SY22, $N_p = 2$ turns, $V_e = 1159.62$ mm$^3$, $A_e = 30.8$ mm$^2$, and $I_g = 0.5$ mm).
0.090 mm. These thicknesses are the real copper foil thickness of PCB board. Also, each material is copper where bulk conductivity is $46 \times 10^6$ S/m. The frequency of the simulation is 5 MHz. The current direction of this simulation is shown in Figure 11(b). The results of the simulation are shown in Figure 12. These results indicate the relationship between the winding width $w$ and the winding resistance per unit length $r_{lay}$. From the approximate of the plot, the relationship equation can be obtained as follows:

$$r_{lay1,8} = 0.7047w^{-0.605},$$

$$r_{lay2,7} = 0.9195w^{-0.667},$$

$$r_{lay3,4,5,6} = 0.8040w^{-0.635}. \tag{18}$$
As mentioned above, layers 1, 2, 7, and 8 are assigned to primary-side windings, while layers 3, 4, 5, and 6 are assigned to secondary-side windings. The primary-side turn number is 3 turns, 1 layer has 2 turns, and parallel number is 2. Also, the secondary-side turn number is 1 turn, 1 layer has 1 turn, and parallel number is 2. Thus, each winding resistance is

\[ R_{\text{pri}} = \left( \frac{R_{\text{lay}1} + R_{\text{lay}2}}{R_{\text{lay}1} + R_{\text{lay}2}} \right) \left( \frac{R_{\text{lay}7} + R_{\text{lay}8}}{R_{\text{lay}7} + R_{\text{lay}8}} \right), \]

\[ R_{\text{sec}} = \frac{R_{\text{lay}3}R_{\text{lay}4}}{R_{\text{lay}3} + R_{\text{lay}4}} = \frac{R_{\text{lay}5}R_{\text{lay}6}}{R_{\text{lay}5} + R_{\text{lay}6}}. \]  

(19)

The copper loss of planar transformer \( P_{\text{copper}} \) can be estimated as follows:

\[ P_{\text{copper}} = R_{\text{pri}}I_1^2 + 2R_{\text{sec}}I_2^2, \]

where \( R_{\text{pri}} \) and \( R_{\text{sec}} \) are winding resistance of primary-side and secondary-side winding, respectively. \( I_1 \) and \( I_2 \) are root mean square (RMS) current of the primary-side and secondary-side winding.

Core loss of planar transformer \( P_{\text{core}} \) and total power loss \( P_{\text{sum}} \) can be derived as follows:

\[ P_{\text{core}} = C_{\text{m}}f_s^a B_m^b \mu_0 V, \]

\[ P_{\text{sum}} = P_{\text{copper}} + P_{\text{core}}. \]

(21)

From these derived equations and FEM with Maxwell 3D, the relationship between primary-side winding width \( w_1 \) and \( P_{\text{sum}} \) can be obtained, as shown in Figure 13. The minimum total power loss can be found in this figure. From this figure, \( w_1 \) and air gap \( l_g \) can be decided. The ratio of \( a \) and \( b \) can be decided from Figure 14. It can be found that \( B_m \) is less than the limitation of \( B_m \) from Figure 15. From these figures and equations, the physical parameters of planar transformer can be obtained as follows: \( a = 5.15 \text{ mm}, b = 3.7 \text{ mm}, c = 16 \text{ mm}, h = 2 \text{ mm}, p = 4.3 \text{ mm}, g = 0.5 \text{ mm}, w_1 = 1.4 \text{ mm}, w_2 = 3.3 \text{ mm}, l_g = 1 \text{ mm}, l_e = 20 \text{ mm}, \) and \( A_e = 19.07 \text{ mm}^2. \)

4. Experimental Results

Some experiments are done with parameters and components, as shown in Table 2. The prototype MHz-level LLC resonant DC-DC converter with GaN-HEMTs is shown in Figure 16. In this prototype circuit, to suppress the power loss at diodes and use the board area effectively, diodes D1 and...
Figure 15: $a/b$ versus $B_m$ (fixed parameters: $L_{11} = 250 \text{nH}, \mu_s = 80, N_p = 3, \mu_0 = 4\pi \times 10^{-7} \text{H/m}, h = 2 \text{mm}, \text{and } g = 0.5 \text{mm}$).

Figure 16: The prototype MHz-level isolated DC-DC converter with GaN-HEMTs.

Table 2: The parameters and components.

<table>
<thead>
<tr>
<th>Name</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>Primary-side GaN-HEMT (EPC2001C)</strong></td>
<td></td>
</tr>
<tr>
<td>ON resistance: $R_{\text{DS(on)}}$</td>
<td>7 m$\Omega$</td>
</tr>
<tr>
<td>Output capacitance: $C_{\text{iss}}$</td>
<td></td>
</tr>
<tr>
<td>Reverse conduction voltage: $V_{\text{DS}}$</td>
<td>430 pF ($V_{\text{DS}} = 50 \text{V}$)</td>
</tr>
<tr>
<td><strong>Secondary-side diode (PDS1040L)</strong></td>
<td></td>
</tr>
<tr>
<td>Forward voltage: $V_F$</td>
<td>0.3 V</td>
</tr>
<tr>
<td>Diode resistor: $R_d$</td>
<td>10 m$\Omega$</td>
</tr>
<tr>
<td>Junction capacitance: $C_j$</td>
<td>1 n$\text{F}$</td>
</tr>
<tr>
<td><strong>Transformer (core: SY22)</strong></td>
<td></td>
</tr>
<tr>
<td>Magnetizing inductance: $L_m$</td>
<td>195 nH</td>
</tr>
<tr>
<td>Leakage inductance: $L_{\text{leak}}$</td>
<td>40.82 nH</td>
</tr>
<tr>
<td>Turns ratio: $n$</td>
<td>3</td>
</tr>
<tr>
<td>Coupling coefficient: $k$</td>
<td>0.9095</td>
</tr>
<tr>
<td><strong>Current transformer (B82801B)</strong></td>
<td></td>
</tr>
<tr>
<td>Secondary-side short inductance: $L_{\text{CT}}$</td>
<td>7 n$\text{H}$</td>
</tr>
<tr>
<td><strong>Resonant capacitor (C1608C0G2A152J)</strong></td>
<td>Capacitance: $C_{11}, C_{12}$</td>
</tr>
</tbody>
</table>

Driver (LM5113)
$D_2$ are 2 parallel connections, respectively. The volume of the prototype converter is 4.1 cm$^3$.

From Figure 17, the ZVS of the primary-side switches at turn-on seems to be achieved during the dead time even if the input voltage is changed. The temperature distributions of the prototype DC-DC converter are exhibited, as shown in Figure 18. Each core temperature can be found lower than 60°C. These results mean low flux density $B_m$. Each calculated $B_m$ is as follows: $B_m = 28.17$ mT under $V_i = 42$ V and $I_o = 8.3$ A condition, $B_m = 27.41$ mT under $V_i = 48$ V and $I_o = 8.3$ A condition, and $B_m = 26.90$ mT under $V_i = 53$ V and $I_o = 8.3$ A condition. Also, at minimum input voltage, the temperature of winding and diode seems to be higher than other conditions.

From these experimental results and datasheets of devices [18, 19], each power loss analysis is conducted, as shown in Figure 19. It can be seen that the target efficiency is achieved at $V_i = 48$ and 53 V. On the other hand, it can be seen that the target efficiency is not achieved at worst condition $V_i = 42$ V. According to this analysis, the major reason of this problem is considered 6W of copper loss.

Also, from these experimental results, it can be found that the power density of the prototype converter is 24.4 W/cc.

5. Conclusions

In this paper, the considerations of physical design and implementation for wide regulation range LLC resonant DC-DC converter with GaN-HEMTs are revealed. Also, the physical parameters and size of the planar transformer are optimized by using derived equations and FEM with Maxwell 3D.

The prototype LLC resonant DC-DC converter using GaN-HEMTs is tested with open loop. The specifications of this converter are that the input voltage is 42–53 V, the reference output voltage is 12 V, the load current is 8 A, the maximum switching frequency is about 5 MHz, and the total volume of the circuit is 4.1 cm$^3$. ZVS operation of the primary-side switches at turn-on can be achieved during the dead time even if the input voltage is changed. The prototype DC-DC converter is obtained more than 90% of power efficiency at $V_i = 48$ and 53 V, $I_o = 8$ A. Also, the power density of the prototype converter is 24.4 W/cc.

Competing Interests

The authors declare that they have no competing interests.
Figure 18: The temperature distributions (A: GaN, B: driver, C: winding, D: core, and E: diode).

(a) $f_s = 3.85 \text{ MHz}$, $V_i = 42 \text{ V}$, $I_o = 8.3 \text{ A}$, and $V_o = 12.647 \text{ V}$

(b) $f_s = 4.31 \text{ MHz}$, $V_i = 48 \text{ V}$, $I_o = 8.3 \text{ A}$, and $V_o = 12.333 \text{ V}$

(c) $f_s = 4.81 \text{ MHz}$, $V_i = 53 \text{ V}$, $I_o = 8.3 \text{ A}$, and $V_o = 12.187 \text{ V}$

Figure 19: The power loss analysis of the experiment at $P_o = 100 \text{ W}$.
References


