Design of High Power Density LLC Resonant Converter with Extra Wide Input Range

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Abstract—A high power density LLC resonant converter design with extra wide input range is presented. The design consideration for the resonant parameters is discussed. Effect of magnetic component design on the converter efficiency is studied by loss analysis and experiment. The prototype with switching frequency higher than 500 kHz is built to verify the design.

I. INTRODUCTION

The portable device industry has been quickly expanding in the last decade. The charger of low power, a “partner” of the portable devices such as cell phones, palm, notebook, etc., has attracted more attention. Nowadays, the chargers are mainly pulse width modulation (PWM) converters, such as flyback and forward converters. In need of higher power density, the converters used as chargers are always required to operate at higher switching frequency with higher efficiency. With higher switching frequency, the size of passive components can be reduced. However, it is not easy for the traditional PWM converters to realize ZVS (which is critical for converters to achieve high efficiency at high frequency).

Due to the feature of high efficiency at very high switching frequency, resonant converters have redrawn more and more attention. The LLC resonant converter, one resonant converter, has been successfully employed in the medium power applications in distributed power systems [2, 3, 4]. In this paper, the LLC resonant converter (see Fig. 1) is proposed for the low power applications. The well-designed LLC resonant converter can realize ZVS for the main switches from no load to full load. Besides, ZCS can also be realized for the rectifier diodes of the LLC converter so that the voltage stress on them can be minimized. By applying frequency control, the LLC converter can operate well with extra wide input voltage range.

In this paper, the DC characteristic of the LLC converter is introduced and the design method of the LLC converter with extra wide input voltage range is presented in Section II. The key point to obtain high efficiency of the LLC converter in the lower power applications is explored in Section III. In Section IV an appropriate magnetic integration structure for the low-power LLC resonant converter is presented. A prototype is built to verify the theoretical analysis.

II. DESIGN CONSIDERATION OF LLC CONVERTER

The topology of the LLC resonant converter is shown in Fig. 1. The principle and the operation modes of it have been discussed in many papers [1-5]. The fundamental component simplification (FES) method is usually used to simplify the theoretical analysis. The fundamental component equivalent circuit of the LLC converter is shown in Fig. 2. The input voltage $E_{in}$ is a series of square wave pulses generated by the half-bridge inverter. The fundamental element RMS value of $E_{in}$ can be expressed by (1).

$$E_{in} = \frac{\sqrt{2}}{\pi} V_{in} \quad (1)$$

The output voltage $E_o$ is the primary-side voltage of the transformer. The fundamental element RMS value of $E_o$ can be expressed by (2).

$$E_o = n \cdot \frac{2\sqrt{2}}{\pi} V_{in} \quad (2)$$

The equivalent circuit (see Fig. 2) contains three passive components in the resonant tank: the series resonant inductor $Ls$, the series resonant capacitor $Cs$ and the parallel inductor $Lp$. The symbol $R_{ac}$ denotes the load resistance reflected to the transformer primary side ($R_{ac} = n^2 \cdot \frac{8}{\pi^2} R_L$ where $n$ is the transformer turns ratio and $R_L$ is the load resistance).

The DC voltage gain $G_{dc}$ of the LLC converter can be derived as

Fig. 1 LLC resonant converter as charger

Fig. 2 AC equivalent circuit of the LLC resonant converter shown in Fig. 1
\[
G_{dc} = \frac{V_o}{V_{in}} = \frac{1}{2n} \left\{ \frac{1}{1 + \frac{1}{k} \left( 1 - \frac{f_s^2}{f_m^2} \right) + j \left( \frac{f_s - f_m}{f_m} \right) Q} \right\}
\]

(3)

Where inductor ratio \( k = \frac{L_L}{L_s} \), quality factor \( Q = \sqrt{\frac{L_s}{C_s R_m}} \),
series resonant frequency \( f_s = \frac{1}{2\pi\sqrt{L_s C_s}} \).

Fig. 3 shows the relationship between the DC voltage gain \( G_{dc} \) and \( f_s/f_m \) obtained from Mathcad simulation. In Fig. 3 (b), the operation region can be divided into three subregions due to the two resonant frequencies \( f_s \) and \( f_m \) (series parallel resonant frequency \( f_m = \frac{1}{2\pi\sqrt{(L_s + L_p) C_s}} < f_s \)). When the switching frequency \( f \) is higher than \( f_m \) but lower than \( f_s \), zero voltage switching (ZVS) turn-on and low current turn off can be achieved for the LLC resonant converter. As a result, the switching loss can be minimized. Meanwhile the voltage stress on the secondary rectifier is limited to two times of the output voltage, low voltage-rating diodes can be used and the secondary conduction loss can be reduced [2,3]. For these reasons, the determination of the transformer turns ratio should satisfy (4).

\[
G_{dc} \leq \frac{V_o}{V_{in(Max-dc)}} (@ f = f_s)
\]

(4)

By substituting (3) into inequality (4), we can get

\[
n \geq \frac{V_{in(Max-dc)}}{2V_o}
\]

(5)

On the other hand, too large transformer turns ratio may increase the magnetic core loss of both the transformer and the paralleled inductor \( L_p \). It is also true for the copper loss of the parallel inductor \( L_p \). (The total loss of the LLC converter will be analyzed in detail in the next section). Therefore, the minimum value of the transformer turns ratio needs to determined to meet the input voltage variation range.

From Fig. 3(a), the smaller the inductor ratio \( k \) is, the more sensitive the DC voltage gain \( G_{dc} \) to the change of the switching frequency \( f \). Any small change of the switching frequency can bring a large change of the DC voltage gain, which means the LLC converter with frequency control can operate well over the extra wide input voltage range. However, the small inductor ratio \( k \) means the large current through the parallel inductor \( L_p \) and the large MOSFET turn off current, which may increase the switching loss and the loss of the parallel inductor \( L_p \). Thus the value of \( k \) cannot be designed too small.

The quality factor \( Q \) is another important parameter of the LLC resonant converter. From Fig. 3(b), the higher \( Q \) is, the lower the peak DC voltage gain. When the load becomes heavier, \( Q \) will become higher. The determination of \( Q \) must meet the requirement of the DC voltage gain for the full load in case of minimum input voltage. After the determination of \( n, k, Q \), and \( f_s \), the values of \( L_s, C_s \), and \( L_p \) can be calculated.

Fig. 3 DC Characteristic of the LLC converter

In this paper, a charger with the LLC resonant converter topology is used as an example. The requirement of it is as
follows:
- Input voltage: 85-265 Vac (the DC voltage after the rectifier bridge is 70-370 Vdc)
- Output voltage: 6 V
- Output current: 200-800 mA
- Switch Frequency: above 500 kHz

Applying the method mentioned above, the final design results are: \( n = 31, Q_{\text{max}} = 0.16, k = 2 \). Then the circuit parameters are calculated as follows: \( L_r = 160 \, \mu\text{H}, C_r = 220 \, \mu\text{F}, L_p = 320 \, \mu\text{H} \). When the circuit parameters are selected, the DC voltage gain \( G_{\text{dc}} \) as the function of the switching frequency \( f \) can be shown as Fig. 3(c). Four different color curves correspond to the different load conditions (100% load, 75% load, 50% load, and 25% load). From Fig. 3(c), we can see the DC voltage gain requirement is met.

### III. LOSS BREAKDOWN

In the low power applications, the efficiency is very sensitive. The LLC resonant converter can successfully realize ZVS for the main switches and ZCS for the rectification diodes. Except these, other losses affect the efficiency greatly. We take the LLC converter designed above as an example to analyze the loss. The total loss of the converter mainly includes MOSFET conduction loss, MOSFET turn-off loss, \( L_r \) core loss, \( L_s \) copper loss, \( L_p \) core loss, \( L_o \) copper loss, transformer core loss, transformer copper loss and diode conduction loss.

When the input voltage is 300 Vdc and the output voltage is 6 Vdc (@ \( I_o = 0.8 \, \text{A}, f = 700 \, \text{kHz} \)), the measurement data of the currents and voltages in the circuit, together with the information from the magnetic design were used to calculate the loss in the different parts of the circuit. The result is presented in the Fig. 4 and Table 1. The estimated efficiency is 80.5% (which is in a good agreement with the efficiency of 79% measured in the experiment. The detailed experimental data will be shown in the next section). The conduction loss of the rectification diode takes a major part in the total loss. However, the diode conduction loss cannot be reduced further if the synchronous rectification is not used. The core loss of the three magnetic components ranks the second in the total loss. As far as the converter with high frequency and low power is concerned, the magnetic material and the flux density enormously affect the magnetic core loss. TP5A made in TDG and 3F35 made in Ferroxcube both operate well from 500 kHz to 1 MHz. Fig. 5 shows the influence of the flux density change \( \Delta B \) in the transformer on the conversion efficiency which is measured in the experiment. When \( \Delta B \) increases from 58 mT to 138 mT, the efficiency decreases by 10%. As for the LLC circuit, the flux density of the transformer can be described as:

\[
\Delta B = \frac{n_p \cdot V_o}{2n_p \cdot f \cdot A_{\text{sec}}} 
\]

Where \( n \) is the turns ratio of the transformer, \( V_o \) is the output voltage, \( n_p \) is the primary winding turns, \( f \) is the switching frequency, and \( A_{\text{sec}} \) is the section area of the magnetic core used.

With the certain circuit parameters and structure of the magnetic core, increasing the winding turns is the only way to decrease the flux density change. However, the number of the winding turns is restricted by the window areas. Increasing winding turns will increase the copper loss. We should optimize the total efficiency by compromising between magnetic core loss and copper loss. The similar situation is also in another two magnetic components. Thus the selection of the flux density change is a key point to design high frequency and low power LLC converter.

![Fig. 4 Loss breakdown of the LLC converter (exclude the MOSFET Driving Loss)](image)

**Table 1 Loss breakdown (Units: mW)**

(exclude the MOSFET Driving Loss)

<table>
<thead>
<tr>
<th>Loss Type</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>MOSFET Conduction Loss</td>
<td>69</td>
</tr>
<tr>
<td>MOSFET Turn-off Loss</td>
<td>3.75</td>
</tr>
<tr>
<td>Ls Core Loss</td>
<td>135</td>
</tr>
<tr>
<td>Ls Copper Loss</td>
<td>11.6</td>
</tr>
<tr>
<td>Lp Core Loss</td>
<td>270</td>
</tr>
<tr>
<td>Lp Copper Loss</td>
<td>15.6</td>
</tr>
<tr>
<td>Transformer Core Loss</td>
<td>135</td>
</tr>
<tr>
<td>Transformer Copper Loss</td>
<td>40</td>
</tr>
<tr>
<td>Diode Conduction Loss</td>
<td>480</td>
</tr>
<tr>
<td>Total Losses</td>
<td>1180</td>
</tr>
<tr>
<td>Output Power</td>
<td>4800</td>
</tr>
<tr>
<td>Efficiency</td>
<td>80.5%</td>
</tr>
</tbody>
</table>

![Efficiency VS Input Voltage](image)

**Fig. 5 Influence of the flux density in the transformer (TDG TP5A) on conversion efficiency**

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IV. MAGNETIC INTEGRATION

In recent years, integrated magnetics have been investigated for many applications. The purpose of the magnetic integration is to reduce the number of magnetic components, and therefore to reduce the total size of the magnetic components. In the LLC resonant converters, all the magnetic components can be integrated into one core. They are shown in Fig. 6. From Fig. 6, we could see that the leakage inductance of the transformer can be used as $L_s$ and the magnetizing inductance of the transformer can be used as $L_p$. With the proper design, the total magnetic loss of the three components can be reduced.

![Fig 6. Magnetic components of the LLC resonant converter](image)

A simple structure is adopted to increase the leakage of the transformer, as shown in Fig. 7. The core is air gapped to get the desired magnetizing inductance $L_p$. In order to get a enough leakage inductance, we reduced the couplings of the primary winding $n_p$ and the second windings $n_{s1}$ and $n_{s2}$. The simulation software Ansoft is used to determine the thickness of the air gap. The result is when the gap is about 0.4mm, the leakage inductance is 119.7 $\mu$H (@ 500 kHz) and the magnetizing inductance is 240.3 $\mu$H (@ 500 kHz). Fig. 8 shows the influence of the magnetic integration on the conversion efficiency which is measured in the experiment. By integrating magnetic components, the efficiency of the LLC converter increases nearly 4% on the average under the conditions of both the full load and 25% load.

V. EXPERIMENTAL RESULTS

A prototype with three magnetic components integrated is built to study the performance of the LLC resonant converter in the low power applications, as shown in Fig.9.

![Fig 9. Photo of the LLC converter](image)

The circuit parameters of the LLC converter (see Fig. 1) are:

- $C_{in}$: 2.2 uF*2,
- $L_s$: 120 uH EE13 (TDG TP5A),
- $C_s$: 220 pF (MLCC),
- $L_p$: 240 uH (TDG TP5A),
- Turns ratio of the integrated transformer: 155:5:5 (TDG TP5A),
- Co: 100uF/25V (Tan) + 4.7uF/16V (MLCC).
The devices used are:
Bridge rectifier: DB107,
Main switches: SPD03N60C3 (Infineon)*2,
Rectifier diodes: B330A (Diode)*2.

Fig. 10(a) shows the $v_{gs}$ and $v_{ds}$ of the switch $S_1$. Fig. 10(b) shows the voltage $v_{ab}$ of the center point of the bridge and the tank current $i_p$ with 110 Vac (150 Vdc) input voltage and full load.

Fig. 11(a) shows $v_{gs}$ and $v_{ds}$ of the switch $S_1$. Fig. 11(b) shows the voltage $v_{ab}$ of the center point of the bridge and the tank current with 220 Vac (300 Vdc) input voltage and full load. From Figs. 10 and 11, it can be seen that $S_1$ is turned on in ZVS condition.

Fig. 12 gives the switching frequency curves at the different input voltages when the output current is 0.8 A, 0.6 A, 0.4 A, and 0.2 A respectively. The switching frequency increases when the input voltage increases, but varies little with the load change. Fig. 13 gives the efficiency curves at different input voltages when the output current is 0.8 A, 0.6 A, 0.4 A, and 0.2 A respectively. It indicates that the efficiency nearly keeps constant when the input voltage increases. The efficiencies at the output currents of 0.8 A, 0.6 A, 0.4 A are all more than 70% over nearly full range of the input voltage. When the output current decreases to 0.2 A, the efficiency decreases rapidly. The main reason may be that the magnetic core loss of the integrated transformer does not decrease with the output power.
VI. CONCLUSION

The LLC resonant converter is studied for high power density chargers with extra wide input voltage range. The design consideration for the resonant parameters is discussed. Effect of the magnetic component design on the efficiency is studied by loss breakdown and experiment. The prototype with switching frequency higher than 500 kHz is built. ZVS for the main switches and ZCS for the rectification diodes can be realized over the whole input voltage range from 70 to 370Vdc. High efficiency of the low power LLC converter is achieved from 50% load to the full load.

ACKNOWLEDGMENT

We would like to thank support of National Natural Science Foundation of China with Project Number 50377037 and 50237030.

REFERENCES:


