Abstract— LLC resonant converter is an excellent candidate for medium power DC/DC converter. However, it is usually used in constant voltage output application. As for the DC/DC stage in high efficiency LED driver, the output voltage may vary widely while output current remains constant. This paper proposes a new design method which is suitable for wide output range application. Design process, analysis and optimization of design parameters are also discussed in detail. A 140W experimental prototype is built to verify the design and proved a high efficiency after optimization. Experimental result shows LLC is a good topology choice for LED driver for it is highly efficient throughout wide output voltage range.

Index Terms—constant current, LED driver, LLC, wide output range

I. INTRODUCTION

Light-emitting diodes (LEDs) are a relatively old technology (1970s) that has advanced from use in numeric displays and indicator lights to a range of new and potential applications, especially general lighting industry. LEDs offer benefits such as high efficiency, long lifetime, durable, small dimensions, producing little heat and no UV, and availability in various colors [1-3]. As a DC source light, an LED needs a constant current flowing through it to maintain a stabilized illumination. Therefore, the design of high efficiency constant current driver is one of the important research fields in LED application.

There are various circuit topologies which can be chosen as an LED driver [2-5]. In some applications, single-stage converters with power factor correction can be used [4]. In some others, passive LED drivers without power switches or control ICs are also utilized [5]. However, in high power applications, two stage LED drivers including both PFC and DC/DC are needed.

In the DC/DC stage of the two-stage LED driver, conventional PWM converters can be used as well as resonant converters. As is known to all, high power density and high efficiency are two major aims of power DC/DC converters we are seeking for, which lead us to develop converters capable of operating at higher switching frequency with high efficiency. As switching frequency increases, the switching losses including the turn-on and turn-off loss of the devices also increase. In switch-mode PWM power converters, the switching losses can be extremely high, which prohibit the operation of the converter at very high frequency. In resonant-mode power converters, however, the switching losses are inherently low, allowing the resonant converter to operate at higher frequency.

Generally, there are three most common resonant converters, series resonant converter (SRC), parallel resonant converter (PRC) and LLC resonant converter. Either SRC or PRC has its drawbacks such as light load output voltage regulation, circulating energy problem and so on [6]. While LLC resonant converter in its half-bridge implementation has drawn more and more attention and become the most popular topology for many applications, since it has many advantages over other topologies, such as high efficiency, high switching frequency, ZVS turn on over entire load range, low level of EMI emissions and so on [6-9]. Operation principle and design method of LLC have been discussed a lot in some previous papers [7-9]. [7] and [9] have proposed some design methods based on frequency domain analysis, while [8] depicts the circuit in a time domain way. All of them have given some perspectives in designing an appropriate LLC resonant converter.

However, conventional LLC resonant converter is usually used as a voltage source and is thought to be inappropriate for constant current output application. When used as an LED driver, the converter has to feed constant current to an LED load. This paper proposes a new design method for constant current LLC, which ensures a high efficiency over wide output voltage range. It can be proved that LLC is an excellent topology selection for LED driver.

II. CIRCUIT OPERATION ANALYSIS

Circuit topology of a half-bridge LLC resonant converter is illustrated in figure 1. S1 and S2 serve as switch transistors with a duty cycle of 0.5 respectively. The series-resonant tank consists of resonant inductor Lr, magnetizing inductance Lm, and resonant capacitor Cr. Diodes D1-D4 form a full-bridge rectifier. Output
capacitance \( C \) is used to absorb current ripple and provide DC current to the LED array.

As shown in figure 3, the DC characteristic of LLC can be divided into three regions according to different modes of operation.

Region 3 is the capacitive region, in which the primary switches operate under ZCS condition, which leads to high switching loss and even abnormal working state. Therefore, the converter should be prevented from entering region 3.

Region 1 and 2 are both ZVS regions. When the converter works at the boundary of region 1 and 2, that is, switching frequency \( f_s \) equals to series resonant frequency \( f_r \), the impedance of resonant tank gets its lowest value and the primary loss reaches minimum. Therefore, the theoretically optimal efficiency can be obtained at the resonance point.

In region 2, the converter works under discontinuous mode. The secondary diodes current decreases to zero naturally without reverse recovery. Both ZVS turn on of primary switches and ZVS turn off of secondary diodes can be achieved. Switching losses of both primary side and secondary side are reduced. Thus, the best design area is in region 2.

However, for the design of wide output range and constant current converter, output voltage varies widely, which leads to a large variation of DC gain. If we choose switching frequency to be less than \( f_r \), it is difficult to ensure the converter operates in ZVS region throughout the entire load range. Furthermore, in a constant current design, the converter operates at the interval between the highest gain point and the resonant point, in which a larger DC gain corresponds to an operation point farther away from the resonant point; which means when the output voltage is high, the switching frequency is much lower than resonant frequency. The switching frequency at light load has to be designed at \( f_r \). Thus, full load efficiency cannot be optimized in region 2.

In region 1, switching frequency \( f_s \) is higher than \( f_r \). The impedance of the resonant tank is inductive, and the primary switches can achieve ZVS condition so that switching loss can be reduced. However, reverse recovery of secondary rectifier diodes exists, thus ZVS turn off condition is lost. But for low output current application, it has little effect on efficiency. As shown in figure 3, the slope of DC curve in region 1 is relatively large. Wide
DC gain variation can meet the design requirement of wide output range. For the above reasons, region 1 is chosen as the operation region for wide output range constant current LLC.

III. DESIGN CONSIDERATION

A. Design Principle of Wide Output Range LLC

For the half-bridge LLC converter, the DC gain is normalized with $\frac{V_{in}}{2}$. Thus, DC gain can be expressed as follow.

$$G_{dc} = \frac{nV_o}{V_{in}/2}$$  \hspace{1cm} (4)

Let $G_{dc}$ equals to 1, we can get the optimal point (the resonant point when switching frequency equals to series resonant frequency) easily from the following equation.

$$N_{nor} = \frac{V_{in}/2}{V_o}$$  \hspace{1cm} (5)

Where $N_{nor}$ is the expected transformer turns ratio to set the LLC working at resonant frequency.

It can be seen from the DC curve shown in figure 3 that, in region 2, the smaller the DC gain is, the larger the slope of DC curve. If the converter is designed to be operating at resonant frequency at full load, the switching frequency at half load (that is the DC gain decrease to 0.5) will increase to almost three times of $f_r$, which means an extremely wide switching frequency range. In order to narrow $f_s$ range, the operating region has to be chosen at higher frequency, which corresponds to a larger slope of DC curve and a DC gain less than 1. Comparing (4) and (5) it is known that, to ensure $G_{dc} < 1$, we must choose $n < N_{nor}$.

$$nV_o = \frac{1}{V_{in}/2} \sqrt{(1 + \frac{1}{m} - \frac{1}{m-f_n})^2 + \left(\frac{n^2}{8} \cdot \frac{V_o}{I_o}\right)^2 \left(\frac{1}{f_n} - fn\right)^2}$$  \hspace{1cm} (6)

$$V_o = \sqrt{\left(\frac{1}{m-f_n}\right)^2 \left(\frac{1}{fn}\right)^2 - \left(\frac{L_r}{Cr \cdot Io^2}\right) \left(\frac{n^2}{8} \cdot \frac{1}{\pi^2}\right)^2}$$  \hspace{1cm} (7)

Curves of different output current can be drawn according to (7) as shown in figure 5, the switching frequency at different output voltage can be easily found, and thus $f_s$ range can be defined. The curve $I_o=0.7A$ is the projection on $(V_o, f_n)$ plane of the intersection curve in figure 4. Switching frequency varies from 1.22$f_r$ to 2.11$f_r$ when $V_o$ varies from 200V to 100V. As discussed above, $n$ is chosen to be less than $N_{nor}$. In figure 5, $n=0.88N_{nor}$.

B. Design Parameters Consideration

In the design process of an LLC resonant converter, several parameters have to be defined. In this proposed design method, series resonant frequency $f_r$, resonant tank parameters $L_r$ and $C_r$, transformer turns ratio $n$ and inductor ratio of the magnetizing inductor of the transformer $L_m$ to the resonant conductor $L_r$ are chosen as design parameters.

To determine series resonant frequency $f_r$, firstly we have to define the switching frequency at full load, for $f_r$ should be lower than switching frequency. Considering the design of magnetic components, normal operation frequency should be set around 100kHz. Higher switching frequency can reduce the size of the magnetic components thus to improve power density with the penalty of increasing switching loss. So, the design in this paper sets switching frequency at full load at about

![Fig. 4: constant current curve](image-url)

![Fig. 5: constant current curves on $(V_o, f_n)$ plane](image-url)
100kHz. As is discussed earlier, $G_{\text{dc}}$ should be smaller than 1 to prevent the switching frequency at half load from rising too high. Thus, $f_s$ should be higher than $f_r$. Therefore, $f_r$ should be lower than 100 kHz. 60~80 kHz is an ideal range.

There are different choices of resonant parameters $L_r$ and $C_r$. For the design of a fixed $f_r$, the smaller $C_r$ is, the larger $L_r$ and $Q$, and the steeper the DC curve is. So, decreasing $C_r$ helps to narrow switching frequency range and increase efficiency at half load. However, a small $C_r$ corresponds to a large $L_r$, which would add more core loss and winding loss. The selection of $C_r$ and $L_r$ is a trade off. Once $C_r$ is determined, $L_r$ can be calculated as follow.

$$L_r = \frac{1}{4\pi^2 f_r^2 C_r} \quad (8)$$

As mentioned above, the real transformer turns ratio $n$ should be smaller than expected turns ration $N_{\text{nor}}$ to narrow switching frequency range. When $n$ deceases, the switching frequency at half load will fall, but the heavy load operation will be farther away from resonant point. That means the converter is operating at a more continuous mode, in which the primary switches are forced to turn off when the resonant current is still large, and the rectifier diodes are turned off at a large current as well, and also with a serious reverse recovery. Therefore, the switching loss of both the MOSFETs and the diodes will increase, especially in large output current application. As a result, $n$ cannot be too small. Figure 6 gives the operation ranges with three different $n$ ($n_1>n_2>n_3$). The curve with $n_1$ has the widest operation range while $n_3$ the narrowest. But the design with $n_3$ operates too far away from the resonance point ($f_n=1$). As a result, $n_2$ is the best selection, of which $f_s$ range is 80~150kHz.

$$V_o = \frac{2n_1^2 R_o T^2}{L_m} + 8\pi^2 \quad (9)$$

$$I_{\text{rms, s}} = \frac{1}{4} \frac{V_o}{n_R} \sqrt{\frac{5\pi^2 - 48 n_1^2 R_o T^2}{12\pi^2 L_m}} + 1 \quad (10)$$

Where $T$ is the switching period.

However, with a fixed $L_r$, a larger $L_m$ means a larger $m$, which would increase the switching frequency at all load conditions as shown in figure 7. $m_3>m_2>m_1$, design with $m_1$ has the lowest switching frequency at any output voltage. Moreover, the core loss and winding loss of $L_m$ increases as $L_m$ increases. As the inductor ratio $m$ increases, the circuit changes from an LLC to a SRC. The above two reasons are why we choose LLC instead of SRC for this design.

Therefore, $L_m$ should be carefully designed with trade-off.

The last parameter to determine is the inductor ratio of the magnetizing inductor of the transformer $L_m$ to the resonant conductor $L_r$. That is to determine $L_m$ since $L_r$ has already been defined. The RMS currents on both primary and secondary sides are purely determined by $L_m$ as shown in (9) and (10) [7]. Therefore, $L_m$ should be maximized to reduce conduction loss on both primary and secondary sides.

$$I_{\text{rms, p}} = \frac{V_o}{8} \frac{2n_1^2 R_o T^2}{n_R L_m} + 8\pi^2$$

$$I_{\text{rms, s}} = \frac{1}{4} \frac{V_o}{n_R} \sqrt{\frac{5\pi^2 - 48 n_1^2 R_o T^2}{12\pi^2 L_m}} + 1$$

C. Design Procedure

Based on the above analysis and discussion, a detailed optimal design procedure is proposed as follows.

There are 6 steps all together in the proposed design procedure. When the optimal design begins, the first step is to know the design specifications, such as the input voltage, output current, and output voltage range and so on. Then, the expected transformer turns ratio $N_{\text{nor}}$ can be calculated using equation (5).

For example, the design specification of a 140W LLC resonant converter is:

- Input: 400VDC;
- Output: 100~200VDC/0.7A

$N_{\text{nor}}$ is designed according to maximum output voltage $V_{o_{\text{max}}}$:

$$N_{\text{nor}} = \frac{V_{o_{\text{max}}}}{V_o} = \frac{200}{100} = 1$$

Secondly, step 2 is to design the series resonant frequency $f_r$. The lowest switching frequency (when the circuit operates at full load, $V_o=200V$) is expected to be

$$f_s = \frac{V_o}{8} \frac{2n_1^2 R_o T^2}{L_m} + 8\pi^2$$

$$f_s = \frac{1}{4} \frac{V_o}{n_R} \sqrt{\frac{5\pi^2 - 48 n_1^2 R_o T^2}{12\pi^2 L_m}} + 1$$
below 100kHz. As a result, the series resonant frequency \( f_r \) is set to about 70kHz. With the defined \( f_r \), we hope the highest switching frequency (when the circuit operates at half load, \( V_o = 100V \)) not to exceed 200kHz.

After \( f_r \) is fixed, step 3 is to design the resonant tank, including resonant inductor \( L_r \) and resonant capacitor \( C_r \). The trade-off of the two parameters has already been discussed in the above section. Actually there are only a few fixed values for \( C_r \), such as 47nF, 33nF, 22nF and 15nF. The corresponding value of \( L_r \) can be calculated using equation (8).

Step 4 is to design the real transformer turns ratio \( n \) based on the calculated \( N_{nor} \). Remember that \( n \) should be smaller than \( N_{nor} \). It is clear \( N_{nor} = 1 \) in this design. So \( n \) could be in the range of 0.8–0.9 according to figure 6.

Step 5 is to design the inductor ratio \( m \). With the help of figure 7, \( m \) can be defined. Generally, \( m = 3–6 \) is a recommended value.

Of all the design parameters have been defined. Therefore, the DC curve with fixed output current can be drawn with the help of equation (7), and the highest and lowest switching frequencies can be easily obtained. If the \( f_s \) range is within the desired range, we can go to step 6 to design the magnetic components specifically; if not, the above parameters have to be redesigned, until the switching frequency can meet the requirement. The “coarse regulation” in figure 8 means the shape of DC curve can be changed a lot by changing the corresponding parameters, while “fine regulation” means the shape can only be changed a little.

Step 6 is to design the transformer and resonant inductor specifically, including choosing the proper magnetic cores and calculate the turns of each winding. Remember that the leakage inductance of the transformer can also be seen as part of resonant inductance. Make sure the core loss and winding loss are minimized.

With the defined \( f_r \), the best design parameters with which the overall efficiency could reach the maximum have to be found. If not, step 3–6 have to be repeated, until the maximum efficiency is reached.

Finally, we have to check whether the efficiency can meet the design requirement. If not, another resonant frequency \( f_r \) can be set to start another design process, until the efficiency can meet the requirement.

The detailed design process diagram is shown in figure 8.

D. Loss Breakdown

A 140W LLC resonant converter with constant output current is taken as an example to analyze the loss based on the proposed design method. The design specifications are listed in the above section. Following the optimal design process in figure 8, the best design parameters are found and listed in table 1.

<table>
<thead>
<tr>
<th>( f_r/\text{kHz} )</th>
<th>( n )</th>
<th>( L_m/\mu \text{H} )</th>
<th>( m )</th>
<th>( C_r/\text{nF} )</th>
<th>( L_r/\mu \text{H} )</th>
<th>( f_s/\text{kHz} )</th>
</tr>
</thead>
<tbody>
<tr>
<td>60</td>
<td>0.85</td>
<td>800</td>
<td>3.75</td>
<td>33</td>
<td>213</td>
<td>84-150</td>
</tr>
</tbody>
</table>

The total loss of the converter includes MOSFETs switching loss, MOSFETs conduction loss, core loss and winding loss of resonant inductor \( L_r \) and transformer, and rectifier diodes conduction loss.

Because the ZVS operation of primary power switches can be ensured throughout the entire load range, MOSFETs turn on loss can be neglected, while the turn off loss can be estimated by the integration of turn-off current and turn-off fall time of MOSFETs as shown in (12).

\[
P_{\text{off}} = \frac{1}{T_f} \int_{t_f} t_{f'} \left[ (I_{\text{off}} - \frac{I_{\text{off}}}{t_f} \cdot t) \cdot V_{in} \right] dt
\] (12)
Where \( I_{off} \) is the turn-off current of the MOSFET and \( t_f \) is the turn-off fall time.

The conduction loss of both MOSFETs and diodes are decided by the RMS current of primary or secondary side and their body resistances as shown in (13) and (14). The RMS currents are expressed in (9) and (10) in the above section.

\[
P_{\text{con, p}} = I_{\text{rms, p}}^2 \cdot R_{ds}
\]

(13)

\[
P_{\text{con, s}} = 2 \cdot I_{\text{rms, s}}^2 \cdot R_{\text{diode}}
\]

(14)

Where \( R_{ds} \) and \( R_{\text{diode}} \) are the body resistances of MOSFETs and diodes respectively.

To estimate the loss of magnetic components, we should know the magnetic cores first. For the core of \( L_r \), RM6/TP4A from TDG is used, while PQ2625/TP4A is used as the transformer core. Referring to the material characteristic sheet from TDG, the core loss of both the resonant inductor and the transformer can be calculated approximately. The winding loss can be estimated by the RMS current of either primary or secondary side and the equivalent resistance of the winding wire, taking skin-effect and eddying-effect into consideration.

The total loss can be calculated based on the above discussion as shown in figure 9 and table 2 at full load and is 3.65W in total. Because of the full-bridge rectifier, conduction loss of the diodes accounts for the largest portion of the total loss.

### Table II

<table>
<thead>
<tr>
<th></th>
<th>Calculated Loss of Every Single Part</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>switching loss of MOSFETs</td>
</tr>
<tr>
<td>2</td>
<td>conduction loss of MOSFETs</td>
</tr>
<tr>
<td>3</td>
<td>( L_r ) core loss</td>
</tr>
<tr>
<td>4</td>
<td>( L_r ) winding loss</td>
</tr>
<tr>
<td>5</td>
<td>transformer core loss</td>
</tr>
<tr>
<td>6</td>
<td>transformer winding loss</td>
</tr>
<tr>
<td>7</td>
<td>conduction loss of diodes</td>
</tr>
</tbody>
</table>

### IV. Experimental Results

Experiments have been carried out with the parameters designed in table 1. The efficiency curve of the converter with different load conditions is shown in Figure 10. The experimental waveforms at half load and full load are shown in figure 11 (a) and (b) respectively. The operations in both figures are in the continuous region. It can be seen from the efficiency curve that the efficiency is more than 97% at full load.
V. CONCLUSION

The design method of wide output range constant current LLC resonant converter has been proposed and different design considerations compared with conventional constant voltage LLC have been pointed out. Also, operation principle, design procedures and parameter analysis are introduced. The developed methodology has been implemented into a 140W LLC resonant converter design. The experimental results also verify the validity of the proposed design method. The efficiency throughout the entire load range is higher than 95%. Therefore, LLC is a good topology for the DC/DC stage of an LED driver.

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