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Design and optimization of a wide output voltage range LED driver based on LLC resonant topology

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***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 本质上；固有的 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.

传统的LLC谐振变换器已经被用来作为电压源而且被认为不适合于恒流输出应用。

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 absolve current ripple and provide DC current to the LED array.

输出电容C用来解决电流纹波和给LED阵列提供直流电流。

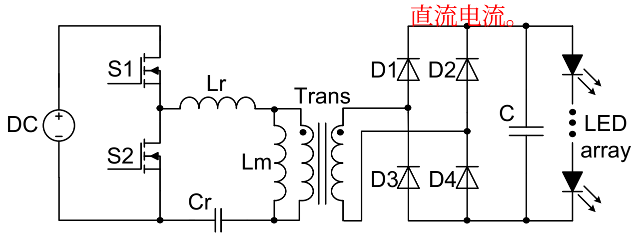

Fig. 1: half-bridge LLC resonant converter as LED driver

FHA近似原理假定只有初级电流的基波用来传送功率给负载

The working principle and operation modes of LLC resonant converter have been discussed in many previous papers [6-9]. The most popular analytical method is first harmonic approximation (FHA) technique, which assumes that only fundamental-harmonic of the primary current transfers power to the load. To apply the FHA method, the AC equivalent circuit has to be deduced first. Figure 2 shows the AC equivalent circuit of LLC, in which Rac is the equivalent load resistance reflected to the transformer primary side ($Rac = n^2 \cdot \frac{8}{\pi^2} \cdot Ro$, where n is the transformer turns ratio and Ro is the load resistance.)

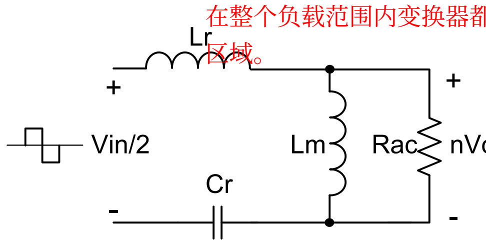

Fig. 2: AC equivalent circuit of LLC

Using the equivalent circuit in figure 2, the DC voltage gain G(dc) can be deduced as follows.

$$G(dc) = \frac{1}{\sqrt{(1+\frac{1}{m}-\frac{1}{m \cdot fn^2})^2 + Q^2 \cdot (fn-\frac{1}{fn})^2}} \qquad (1)$$

Where $m = \frac{Lm}{Lr}$, $fn = \frac{fs}{fr}$, $Q = \frac{\sqrt{Lr/Cr}}{Rac}$. According to (1), the DC characteristic of LLC with different Q can be illustrated in figure 3, in which there are two resonant frequencies. Series resonant frequency fr is determined by Lr and Cr, while parallel resonant frequency fm is determined by Cr and the inductance of Lm and Lr in series, which can be shown as follows.

$$fr = \frac{1}{2\pi\sqrt{LrCr}} \qquad (2)$$

$$fm = \frac{1}{2\pi\sqrt{(Lm+Lr)Cr}} \qquad (3)$$

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 fs equals to series resonant frequency fr, 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. 在区域2，变换器工作在断续模式

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 ZCS 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.

如果我们选择开关频率低于串联谐振频率Fr,那就很难保证在整个负载范围内变换器都工作于ZVS区域。

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 fr, 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 fr. Thus, full load efficiency cannot be optimized in region 2.

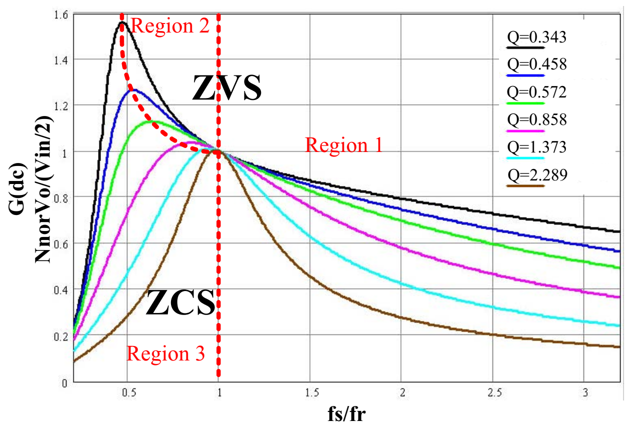

Fig. 3: DC characteristic of half-bridge LLC

在区域1，虽然能取得ZVS，单丝次级整流二极管的反向恢复存在，因此ZCS丢失。但是对于低输出电流应用场合，它对效率的影响是很小的。

In region 1, switching frequency fs is higher than fr. The impedance of the resonant tank is inductive, and the primary switches can achieve ZVS turn on so that switching loss can be reduced. However, reverse recovery of secondary rectifier diodes exists, thus ZCS 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 Vin/2. Thus, DC gain can be expressed as follow.

$$\mathrm{G(dc)} = \frac{nVo}{Vin/2} \tag{4}$$

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.

$$Nnor = \frac{Vin/2}{Vo} \tag{5}$$

Where Nnor 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 fr, which means an extremely wide switching frequency range. In order to narrow fs 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<Nnor.

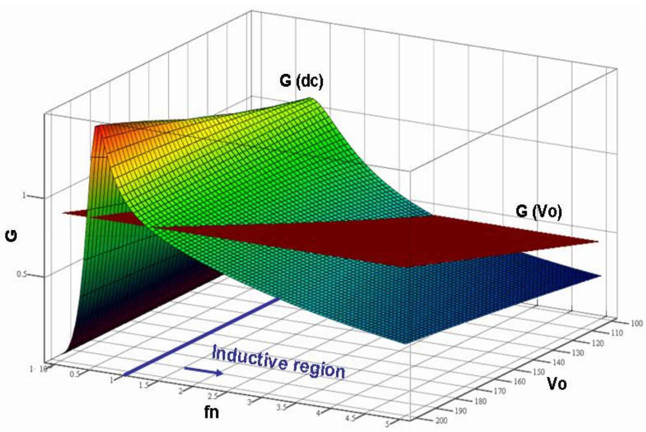

Fig. 4: constant current curve

Figure 4 shows the constant current curve, in which the colored surface G(dc) is the normalized DC gain as a function of Vo and fn, and the brown surface G(Vo) is the DC gain corresponding to different output voltages with the same current. Thus, the intersection curve of the two surfaces in the inductive region is the constant current curve. In order to get its projection curve on (Vo,fn) plane, Vo must be solved using (1) and (4) as shown in (6), where Ro is expressed by Vo/Io. After simplification, Vo can be solved as a function of fn as shown in (7).

$$\frac{nVo}{Vin/2} = \frac{1}{\sqrt{(1+\frac{1}{m}-\frac{1}{m\cdot fn^2})^2 + \left(n^2\cdot\frac{8}{\pi^2}\cdot\frac{Vo}{Io}\right)^2(\frac{1}{fn}-fn)^2}} \tag{6}$$

$$Vo = \sqrt{\frac{1-\left(\frac{n}{Vin/2}\right)^2\cdot(\frac{1}{fn}-fn)^2\cdot\frac{Lr/Cr\cdot Io^2}{\left(n^2\cdot 8/\pi^2\right)^2}}{\left(\frac{n}{Vin/2}\right)^2\cdot(1+\frac{1}{m}-\frac{1}{m\cdot fn^2})^2}} \tag{7}$$

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 fs range can be defined. The curve Io4=0.7A is the projection on (Vo, fn) plane of the intersection curve in figure 4. Switching frequency varies from 1.22fr to 2.11fr when Vo varies from 200V to 100V. As discussed above, n is chosen to be less than Nnor. In figure 5, n=0.88Nnor.

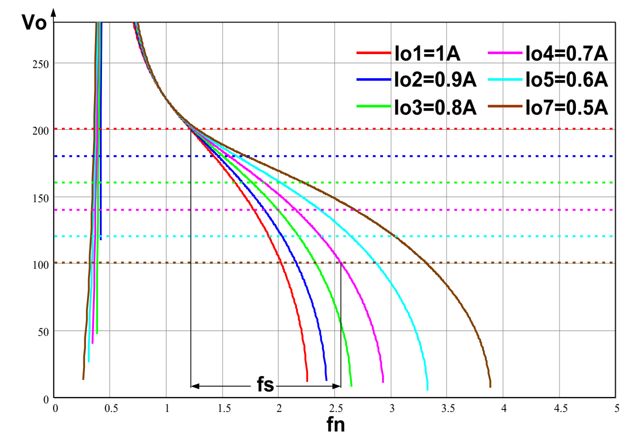

Fig. 5: constant current curves on (Vo, fn) plane

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 fr, resonant tank parameters Lr and Cr, transformer turns ratio n and inductor ratio of the magnetizing inductor of the transformer Lm to the resonant conductor Lr are chosen as design parameters.

To determine series resonant frequency fr, firstly we have to define the switching frequency at full load, for fr 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

G（dc）应该小于1，为了防止开关频率在半载时上升过高。

100kHz. As is discussed earlier, G(dc) should be smaller than 1 to prevent the switching frequency at half load from rising too high. Thus, fs should be higher than fr. Therefore, fr should be lower than 100 kHz. 60~80 kHz is an ideal range.

减小谐振电容Cr有助于减小开关频率的范围和提高半载时的效率。但是，小的谐振电容Cr对应大的谐振电感Lr这样会增加更多磁芯损耗和绕组损耗。所以，Cr和Lr的选择是个折中的过程。

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

$$Lr = \frac{1}{4\pi^2 f1^2 Cr} \tag{8}$$

As mentioned above, the real transformer turns ratio n should be smaller than expected turns ration Nnor to narrow switching frequency range. When n deceases, the switching frequency at half load will fall, but the heavy load operation will be father 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 (n1>n2>n3). The curve with n1 has the widest operation range while n3 the narrowest. But the design with n3 operates too far away from the resonance point (fn=1). As a result, n2 is the best selection, of which fs range is 80~150kHz.

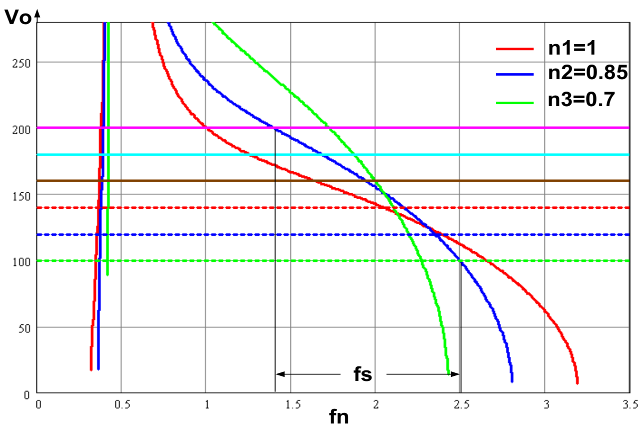

Fig. 6: constant current curves with different n

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

初级和次级电流的有效值均由励磁电感Lm决定。

因此，励磁电感Lm应该最大会，这样可以减小初级和次级的导通损失。

$$I_{rms\_p} = \frac{1}{8}\cdot\frac{Vo}{nRo}\sqrt{\frac{2n^4Ro^2T^2}{Lm^2}+8\pi^2} \tag{9}$$

$$I_{rms\_s} = \frac{1}{4}\cdot\frac{Vo}{nRo}\sqrt{\frac{5\pi^2-48}{12\pi^2}\frac{n^4Ro^2T^2}{Lm^2}+1} \tag{10}$$

Where T is the switching period.

However, with a fixed Lr, a larger Lm means a larger m, which would increase the switching frequency at all load conditions as shown in figure 7. m3>m2>m1, design with m1 has the lowest switching frequency at any output voltage. Moreover, the core loss and winding loss of Lm increases as Lm 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.

而且，Lm的增大会导致磁芯损耗和绕组损耗的增大

Therefore, Lm should be carefully designed with trade-off.

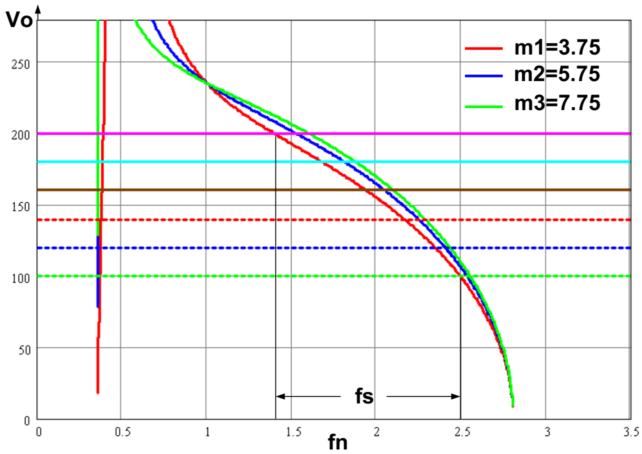

Fig. 7: constant current curves with different m

C. Design Procedure

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

设计第一步就是要知道设计要求：比如输入电压，输出电流，输出电压等等。

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 Nnor 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

Nnor is designed according to maximum output voltage Vomax.

$$Nnor = \frac{Vin/2}{Vo\max} = \frac{400/2}{200} = 1 \tag{11}$$

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

第二步就是设计串联谐振频率Fr，最低开关频率发生在电路满载时。

第五步就是决定K值大小，即励磁电感和谐振电感的比值。K值取3-6合适

below 100kHz. As a result, the series resonant frequency fr is set to about 70kHz. With the defined fr, we hope the highest switching frequency (when the circuit operates at half load, Vo=100V) not to exceed 200kHz. 最高开关频率在半载时

After fr is fixed, step 3 is to design the resonant tank, including resonant inductor Lr and resonant capacitor Cr. The trade-off of the two parameters has already been discussed in the above section. Actually there are only a few fixed values for Cr, such as 47nF, 33nF, 22nF and 15nF. The corresponding value of Lr can be calculated using equation (8). 第三步是设计谐振网络 实际上，谐振电容就那么几个值，如47nF...

Step 4 is to design the real transformer turns ratio n based on the calculated Nnor. Remember that n should be smaller than Nnor. It is clear Nnor=1 in this design. So n could be in the range of 0.8~0.9 according to figure 6.

第四步是根据计算的Nnor设计真实变压器变比,记得n应该小于Nnor

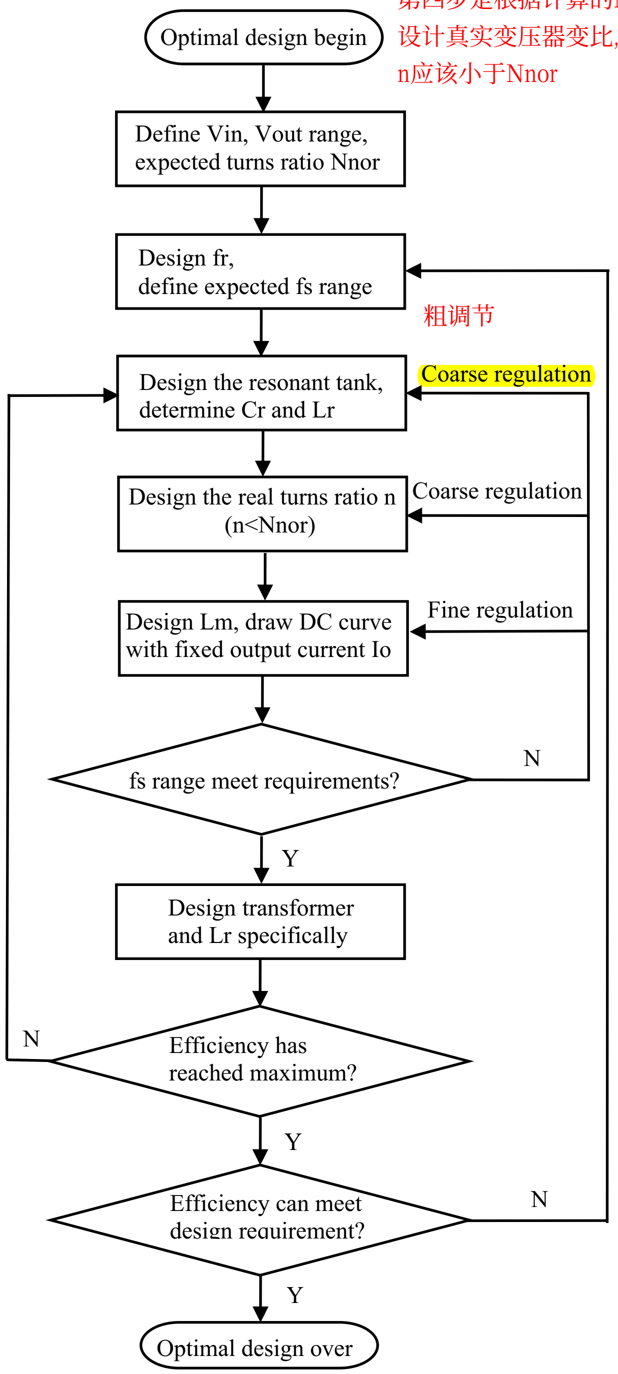

Fig. 8: optimal design procedure

有价值

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.

So far 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 fs 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. 第六步就是设计变压器与谐振电感，包括选择合适的磁芯和计算各自的变比。

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. 确保磁芯和绕组损耗最小化。

With the defined fr, 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 fr can be set to start another design process, until the efficiency can meet the requirement. 最后，我们要检查效率是否能满足要求。如果不能，我们就要重新设计一个串联谐振频率开始设计流程。直到效率达到我们的要求。

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 I
BEST DESIGN PARAMETERS AND FS RANGE

fr/kHz	n	Lm/uH	m	Cr/nF	Lr/uH	fs/kHz
60	0.85	800	3.75	33	213	84-150

The total loss of the converter includes MOSFETs switching loss, MOSFETs conduction loss, core loss and winding loss of resonant inductor Lr 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).

$$P_{off}=\frac{1}{T}\int_0^{t_f}[(I_{off}-\frac{I_{off}}{t_f}\cdot t)\cdot Vin]dt \qquad (12)$$

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.

$$P_{con\_p} = I_{rms\_p}{}^2 \cdot R_{ds} \tag{13}$$

$$P_{con\_s} = 2 \cdot I_{rms\_s}{}^2 \cdot R_{diode} \tag{14}$$

Where R_{ds} and R_{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 Lr, RM8/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
CALCULATED LOSS OF EVERY SINGLE PART

1	switching loss of MOSFETs	0.34W
2	conduction loss of MOSFETs	0.25W
3	Lr core loss	0.23W
4	Lr winding loss	0.37W
5	transformer core loss	0.52W
6	transformer winding loss	0.68W
7	conduction loss of diodes	1.26W

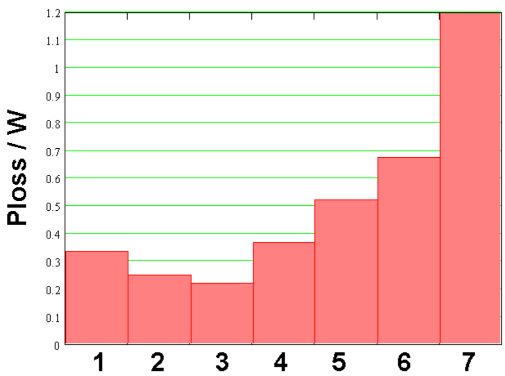

Fig.9: Loss distribution of the whole prototype

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.

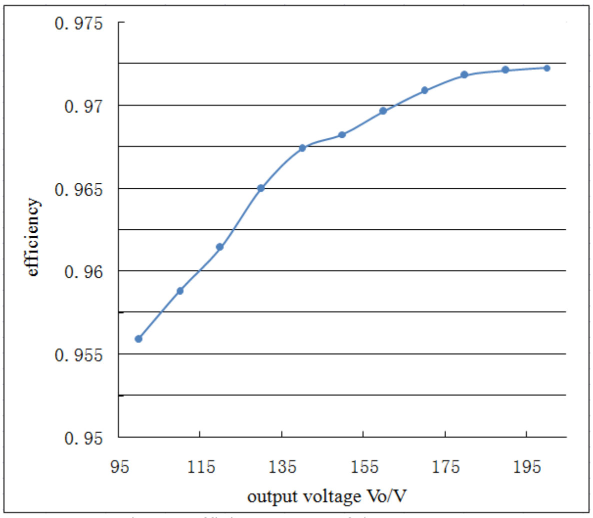

Fig. 10: efficiency curve of the prototype

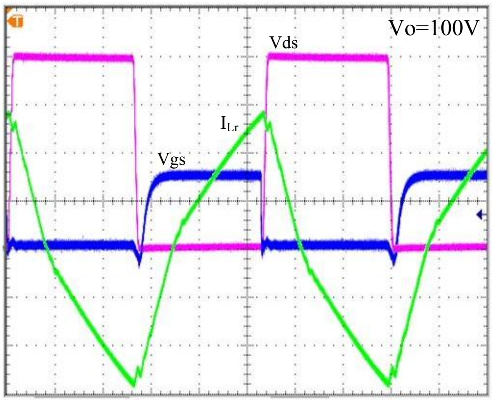

Fig. 11 (a): Vgs, Vds and I_{Lr} waveforms @ half load (10V/div, 100V/div, 500mA/div, 1us/div)

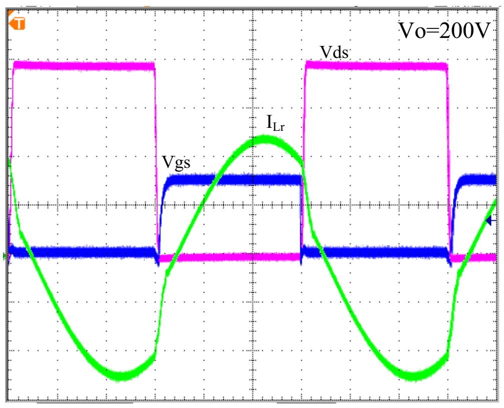

Fig. 11 (b): Vgs, Vds and I_{Lr} waveforms @ full load (10V/div, 100V/div, 500mA/div, 2us/div)

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.

ACKNOWLEDGMENT

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