Optimal Design Methodology for LLC Resonant Converter

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Abstract: Although LLC resonant converter can achieve wide operation range with high efficiency, lack of design methodology makes it difficult to be implemented. In this paper, based on the theoretical analysis on the operation principles during normal condition and holdup time, the relationship between converter efficiency and operation range with different circuit parameters has be revealed. An optimal design methodology has been developed based on the revealed relationship. A 1MHz, 1kW LLC converter is designed to verify the proposed method.

I. INTRODUCTION

With the development of power conversion technology, power density becomes the major challenge for front-end AC/DC converters [1] [2] [3]. Although increasing switching frequency can dramatically reduce the passive component size, its effectiveness is limited by the converter efficiency and thermal management design. Meanwhile, to meet the holdup time requirement, bulky capacitors have to be used to provide the energy during holdup time, which is only affected by DC/DC stage operation input voltage range [1]. The relationship between holdup time capacitor requirement and minimum DC/DC stage input voltage for different front-end converter power levels is shown in Figure 1. Apparently, wide operation range DC/DC stage can reduce the holdup time capacitor requirement and improve the system power density. However, when the minimum voltage is less than 200V, very limited effects can be observed.

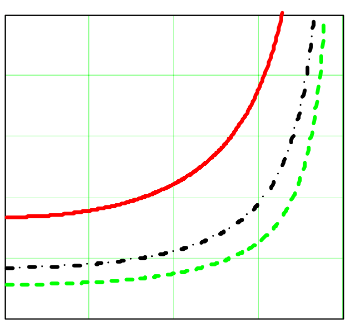

Figure 1. Holdup time capacitor requirement for DC/DC stage with different minimum input voltage.

To reduce the holdup time capacitor requirement, different research efforts have been implemented, by using extra holdup time extension circuit or by developing better topologies [4][5][6]. Among different solutions, LLC resonant converter becomes the most attractive topology due to its high efficiency and wide operation range.

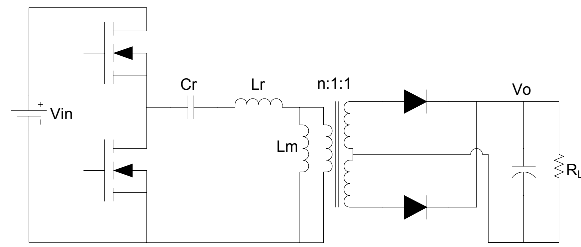

Figure 2. LLC Resonant Converter.

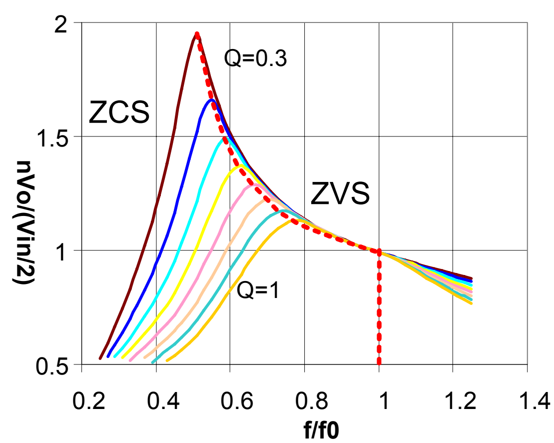

Figure 3. Gain Characteristic of LLC Converter.

The LLC resonant converter topology is shown in Figure 2. By utilizing the transformer magnetizing inductance, LLC converter modifies the gain characteristic of series resonant converter (SRC). Its voltage gain characteristics for different loads are shown in Figure 3, due to the half bridge structure, the output voltage is normalized with half of the input voltage. Comparing with SRC, the converter can achieve both Buck mode and Boost mode. When the switching frequency is higher than resonant frequency, voltage gain of LLC converter is always less than one, and it operates as an SRC converter and zero voltage switching (ZVS) can be achieved. When the switching frequency is lower than resonant frequency, for different load conditions, both ZVS and zero current switching (ZCS) could be achieved. At the boundary of ZVS and ZCS regions, as shown in the dashed line in Figure 3, converter voltage gain reaches it maximum value.

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According to the circuit operation analysis [4], at the resonant frequency, because the impedance of resonant tank, constructed by L_r and C_r, is zero, input and output voltages are virtually connected together. Thus, converter voltage gain is equal to one for all the load conditions. When the input AC line exists, DC/DC stage input voltage is generated by PFC stage and it is regulated at 400V. At this condition, by choosing a suitable transformer turns ratio, converter could always operate at resonant frequency. Therefore, the conduction loss and switching loss can be minimized. During holdup time, energy transferred to the load comes from bulky holdup time capacitor. While DC/DC input voltage keeps decreasing, converter reduces its switching frequency to operate in Boost mode and regulate output voltage. Due to the complexity of resonant tank, design of the LLC resonant converter needs to consider three key elements, resonant frequency, characteristic factor, and inductor ratio,

$$f_0 = \frac{1}{2\pi\sqrt{L_r C_r}} \tag{1}$$

$$Q = \frac{\sqrt{L_r / C_r}}{n^2 R} \tag{2}$$

$$L_n = \frac{L_m}{L_r} \tag{3}$$

Here f_0 is the resonant frequency, which defines the switching frequency of LLC resonant converter. The characteristic factor Q is the ratio between the characteristic impedance and the load. L_n is defined as the ratio between the magnetizing inductance and the resonant inductance.

Although different literatures [9]-[11] have discussed operation principles and benefits of the topology, there is no design guideline developed. Moreover, instead of simply choosing Q value in the conventional SRC or PRC design, LLC requires defining two coupled elements L_n and Q. Apparently, try and error method could result in a good design. However, it is time consuming and not cost effective. As a result, the topology is difficult to be adopted by the industries. In this paper, based on the analysis of LLC resonant converter at different operation conditions, including the normal operation and during holdup time, an optimal design methodology has been developed. Based on the developed method, designed LLC converter can achieve maximum efficiency with desired operation range, which is verified by a 1MHz LLC resonant converter.

II. Circuit Operation Analysis Of LLC Resonant Converter

A. Normal Operation Analysis

At normal operation condition, LLC converter input voltage is regulated by PFC stage. From the gain characteristic curves shown in Figure 3, converter gain can keep constant at resonant frequency. Therefore, by designing a suitable transformer turns-ratio to make the converter voltage gain equal to one, at normal operation condition, LLC resonant converter can always operate at its resonant frequency for different load conditions. For most of the time, front-end converter is operating under this operation mode. Therefore, the efficiency at resonant frequency is the key aspect for the LLC converter performance.

According to the operation of LLC resonant converter, at resonant frequency, resonant tank current is a pure sinusoidal waveform as shown in Figure 4. The dashed line is the magnetizing inductor current. The equivalent circuit of the operation is shown in Figure 5. At first half line cycle, resonant tank current i_r resonates up. At the same time, output voltage is applied to magnetizing inductor. Therefore, the magnetizing inductor current increases linearly. At the end of this half switching cycle, primary switch turns off with peak magnetizing inductor current, and the other switch turns on under ZVS condition with same current. During the other half line cycle, the resonant tank current keeps resonant and output voltage is applied to the magnetizing inductor with reverse polarity. Therefore, the magnetizing inductor current decreases linearly. Thus, a square wave voltage is applied to the magnetizing inductor, and the magnetizing inductor current is a triangle shape as shown in the dashed line in Figure 4. Moreover, at the end of each half switching cycle, magnetizing inductor current reaches its maximum value and the resonant tank current gets the same value at the same time.

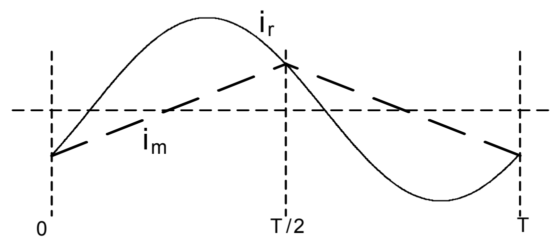

Figure 4. Resonant Tank Current Waveform at Resonant Frequency.

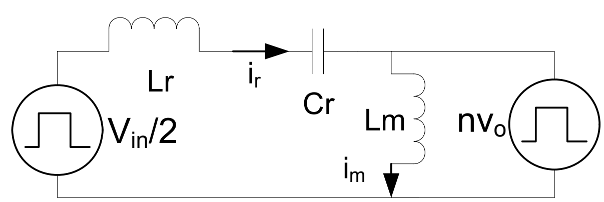

Figure 5. Equivalent circuit at resonant frequency.

The magnetizing inductor peak current can be determined by

$$I_{pk} = \frac{nV_o}{L_m \frac{T}{4}} \tag{4}$$

Here n is the transformer turns ratio between the primary side and secondary side, V_O is the output voltage, T is the switching cycle and L_m is magnetizing inductance.

Since the resonant tank current at resonant frequency is a sinusoidal wave, it can be represented by the equation

$$I_r = \sqrt{2} I_{rms} \sin(2\pi f_0 + \phi) \tag{5}$$

Here I_{rms} is the resonant tank RMS current, and f_0 is the resonant frequency and ϕ is the initial angle of the resonant tank current, which represent the phase difference between the resonant tank current and magnetizing inductor current. According to the current waveforms, at the end of each half switching cycle, magnetizing inductor current is equal to the resonant tank current, which means

$$\sqrt{2}I_{rms}\sin(\phi)=\frac{nV_o}{L_mT/4} \tag{6}$$

On the other hand, the difference between resonant tank current and magnetizing inductor current is the current transferred to the load, thus

$$\int_0^{T/2}(i_r-i_m)dt=\frac{V_O}{nR_L}\frac{T}{2} \tag{7}$$

Here R_L is the load resistance, n is the transformer turns ratio.

By summarizing these equations, the resonant tank RMS current can be solved as

$$I_{rms}=\frac{1}{8}\frac{V_O}{nR_L}\sqrt{\frac{2n^4R_L{}^2T^2}{L_m{}^2}+8\pi^2} \tag{8}$$

Here V_O is the output voltage, n is transformer turns-ratio, R_L is load resistance, T is switching cycle at resonant frequency, and L_m is the magnetizing inductance.

Since the resonant tank current continuously flows trough the primary side switches, its RMS value determines the primary side conduction loss. Comparing with the load current reflected to primary side, resonant tank RMS current is only related to the magnetizing inductance, the load resistance and the switching cycle. While the switching cycle and load resistance are predetermined values for certain converter specifications, resonant tank RMS current is only determined by the magnetizing inductance.

Besides the primary side conduction loss, secondary side rectifier conduction loss is also a major concern. For diode rectifier, its conduction loss major comes from diode forward voltage drop and is proportional to the average output current. However, if considering synchronous rectification, it is also desirable to minimize the secondary side RMS current. Since we already get the formulas for both the resonant tank current and magnetizing inductor current, secondary side current can be easily calculated. Based on previous analysis, secondary RMS can be expressed as

$$I_{RMS\_S}=\frac{1}{4}\frac{V_O}{nR_L}\sqrt{\frac{5\pi^2-48}{12\pi^2}\frac{n^4R_LT^2}{L_m{}^2}+1} \tag{9}$$

From this equation, same as the primary side RMS current, secondary side RMS current is also entirely determined by the magnetizing inductance.

Based on the analysis o the LLC resonant converter operating at resonant frequency, the converter conduction loss is mainly affected by the magnetizing inductance, instead of resonant inductor or the resonant capacitor. At the same time, the primary side switches can achieve ZVS for all the load conditions, the switching loss is mainly coming from the turn off loss, which is also depends on the magnetizing inductance. Therefore, to design a high efficiency LLC resonant converter, it is essential to find a suitable magnetizing inductor.

B. Holdup Time Operation Analysis

During the holdup time, input AC line doesn't exist and PFC stage no longer provides energy to DC/DC stage. All the energy transferred to the load during holdup time is purely coming from holdup time capacitor. Therefore, the input voltage of DC/DC stage will keep decreasing during holdup time. To maintain regulated output voltage, switching frequency of the LLC resonant needs to be reduced so that the converter gain can be boosted up. Different from PWM converters, LLC converter could achieve highest efficiency at high input voltage. During holdup time, the converter operates far away from its resonant point and has less efficiency. However, the holdup time only requires 20mS, and low efficiency could be tolerated and would not cause excess thermal stress.

Because the transformer turns ratio is a fixed value, the required gain is determined by the relationship between the input and output voltage, which can be represented by

$$g=\frac{nV_O}{V_{in}/2} \tag{10}$$

Here, g is the required voltage gain for LLC converter, V_O is the output voltage and V_{in} is the input voltage. From this equation, lower the input voltage, higher the voltage gain is required.

As shown in Figure 1, the holdup time capacitor requirement is largely affected by the operation range of the DC/DC stage. Wider operation range can dramatically reduce the holdup time capacitor requirement and improve the whole converter power density. Therefore, wide operation range of DC/DC stage is desired.

The operation range of LLC converter is decided by the peak voltage gain that can be achieved. At normal operation mode, input voltage is 400V and LLC has a voltage gain equal to one. If the converter can achieve a maximum gain of 2, it will be able to regulate output voltage with 400/2=200V input. Obviously, the higher the peak gain, the wider the operation range of LLC resonant converter.

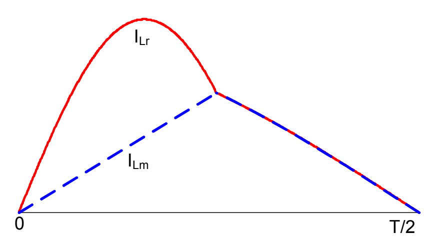

Figure 6. Resonant tank current at peak gain point

From gain characteristic curves in Figure 3, the peak gain happens when the circuit is running at the boundary of zero current switching (ZCS) and zero voltage switching (ZVS) modes. The resonant tank current at this condition is shown in Figure 6. In each half switching cycle, the magnetizing inductor is firstly charged by the output voltage. After that, it participates in the resonance (the resonant tank is constructed by L_r, C_r and L_m) and transfers it stored energy to the resonant capacitor. At the end of each half switching cycle, its current is reset to zero. Therefore, entire energy stored in the magnetizing inductor can be transferred to the load, and the converter gain reaches its peak value. Although the peak gain can be calculated based on the current waveform, it is difficult to solve the equations and get the analytical solution. Therefore, to simplify the analysis, peak gains at different L_n and Q combinations are simulated based on the simulation tool Simplis, which can automatically reach the circuit steady state within short simulation time. The peak gains for different L_n and Q values are summarized in the contour curves in Figure 7. In this set of curves, each line shows the combinations of different L_n and Q values that can achieve same peak voltage gain. For instance, if we want to design a converter with a peak gain of 1.3, any combination of L_n and Q along the line 1.3 can be chosen as a valid design.

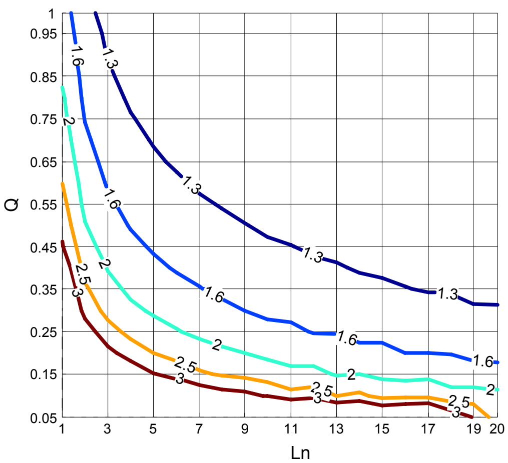

Figure 7. Relationship between converter peak gain and L_n, Q

Apparently, peak gain is affected by both L_n and Q values. By reducing L or Q value, higher peak gain can be achieved.

III. Design Methodology for LLC Resonant Converter

The LLC resonant converter design goal is to achieve minimum loss with the capability of achieve required maximum gain to ensure wide operation range. According to previous analysis, the relationships between the design parameters L_n and Q with the converter performance, especially the conduction loss and the operation range, is revealed. These relationships can be used to develop an optimal design methodology of LLC resonant converter.

LLC resonant converter operates under normal operation condition at most of the time. When the input Ac line exists, the DC/DC stage input voltage is a regulated 400V. Therefore, LLC converter could always operates at resonant frequency and achieve optimal efficiency. Thus, it is essential to minimize the loss when converter operates at resonant frequency. Based on the operation analysis at resonant frequency, both the primary and secondary conduction loss is purely determined by the magnetizing inductance, as shown in the following equation:

$$I_{rms} = \frac{1}{8}\frac{V_O}{nR_L}\sqrt{\frac{2n^4 R_L{}^2 T^2}{L_m{}^2} + 8\pi^2} \qquad (11)$$

$$I_{RMS\_S} = \frac{1}{4}\frac{V_O}{nR_L}\sqrt{\frac{5\pi^2 - 48}{12\pi^2}\frac{n^4 R_L T^2}{L_m{}^2} + 1} \qquad (12)$$

Therefore, to minimize conduction loss, the magnetizing inductance should be maximized to reduce RMS currents on primary side and secondary side. In turn, the copper loss of the magnetic components can also be reduced.

To achieve high power density, high switching frequency is always desired, because the passive component size reduces dramatically with increasing switching frequency. However, the switching loss increases linearly with the switching frequency. Therefore, it is also important to minimize the LLC converter switching loss.

Based on the operation analysis, LLC converter primary side switches can achieve ZVS turn on for all the load conditions. However, the ZVS condition is ensured by the peak magnetizing current, which can be calculated as

$$I_{pk} = \frac{nV_o}{L_m \frac{T}{4}} \qquad (13)$$

Here, n is transformer turns ratio, V_O is output voltage, L_m is the magnetizing inductance and T is the switching cycle. During the primary side switches commutating period, due to the large magnetizing inductance, the magnetizing inductor current can be assumed to be constant. To ensure ZVS turn on, the peak magnetizing inductor current should be able to discharge MOSFETs junction capacitors within dead time, which can be represented by

$$I_{pk} > \frac{2V_{bus}C_j}{t_{dead}} \qquad (14)$$

Here V_{bus} is input bus voltage, C_j is MOSFET junction capacitance and t_{dead} is the dead time.

Although the large turn off current can ensure soft switching condition, it results in larger turn off loss, because primary side switches turning off is hard switching. Therefore smaller turn off current is desirable to reduce turn off loss.

By summarizing previous analysis, to achieve minimum conduction loss, the magnetizing inductance is required to be as large as possible. To ensure minimum switching loss, the magnetizing inductor needs to be smaller enough to achieve ZVS condition and large enough to have smaller turn off current. Therefore, considering both conduction loss and switching loss, the optimally designed L_m should make the primary side turn off current exactly the same as ZVS requirement. Thus,

$$L_m = \frac{T \cdot t_{dead}}{16C_j} \qquad (15)$$

According to the definition of L_n and Q

$$L_n = \frac{L_m}{L_r} \tag{16}$$

$$Q = \frac{\sqrt{\frac{L_r}{C_r}}}{n^2 R_L} \tag{17}$$

together with the resonant frequency

$$f_0 = \frac{1}{2\pi\sqrt{L_r C_r}} \tag{18}$$

we can calculate magnetizing inductance as

$$L_m = \frac{2\pi f_0 L_n Q}{n^2 R_L} \tag{19}$$

Based on this equation, it can be seen that, as long as the production of L_n and Q keeps constant, L_m is a fixed value. Or in other words, for a designed magnetizing inductance, the relationship between L_n and Q is fixed. Once an L_n value is chosen, the corresponding Q value can be designed.

Besides the converter efficiency, the other major aspect of LLC converter is its operation range. Although the converter efficiency is mainly defined by the magnetizing inductance, the L_n and Q value could affect the gain characteristic of LLC resonant converter. As discussed before, wide operation range LLC resonant converter can reduce the holdup time capacitor requirement and improve system power density. To achieve the desired operation range, peak gain is required to be higher than certain level.

A 200 kHz 1kW LLC resonant converter with operation range from 250V to 400V is chosen as an example. In this case, the peak voltage gain is required to be larger than 400/250=1.6. From Figure 7, all the L_n and Q combinations below line 1.6 could meet the gain requirement. However, this results in infinite solutions for the converter design. Furthermore, considering normal operation efficiency, magnetizing inductance is expected to be maximized. Based on the analysis on the magnetizing inductance, we can get the magnetizing inductance should be

$$L_m = \frac{T \cdot t_{dead}}{16 C_{ds}} \tag{20}$$

For 200 kHz switching frequency and 100nS dead time, together with 450pF C_j (IXFH21N50), we can easily calculate the magnetizing inductance should be around 70uH.

Therefore, for required magnetizing inductance of 70uH, a marked line can be added to Figure 7, the peak gain curves, as shown in Figure 8. All the designs along the marked line will have the same magnetizing inductance, which ensures maximum efficiency at normal operation condition. Comparing with peak gain curves, L_n has to be larger than 5.5 to meet the gain requirement. However, along the marked line, there could be infinite solutions for the valid design. To further pickup a suitable value for the L_n and Q combination, we need to further consider the L_n and Q impacts on the circuit operations.

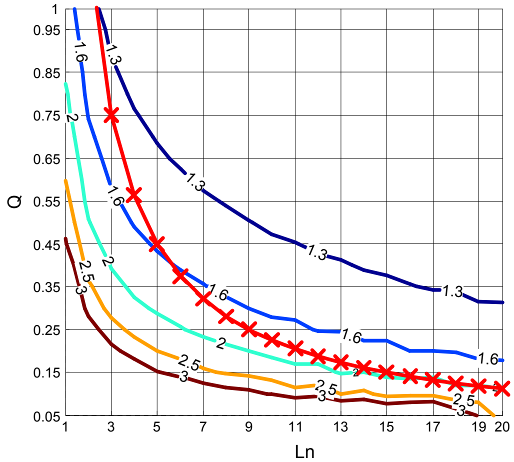

Figure 8. Design example of LLC converter

As shown in Figure 9, the gain characteristics for different L_n values are summarized. It shows that the larger L_n will increase the frequency range of LLC resonant converter. Although at normal operation condition, PFC stage generates a regulated 400V bus, due to the line frequency ripple, LLC converter still needs to do switching frequency modulation to regulated output voltage. Therefore, the switching frequency range should be as small as possible, which requires a minimum L_n value. However, larger L_n makes it easier to use the transformer leakage inductance to realize the resonant inductor. Therefore, trade off between the converter size and efficiency can be carried out and determine the L_n value. Based on the chosen L_n and Q values, L_r and C_r can be calculated accordingly.

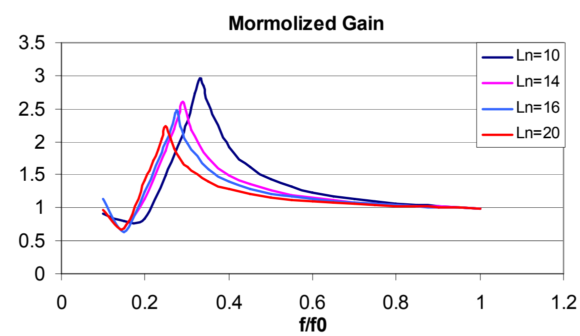

Figure 9. L_n impacts on LLC converter voltage gain

IV. Experimental Implementation

To verify the theoretical analysis, a 1MHz, 1kW LLC resonant converter with wide input voltage range (200V to 400V) and 48V output is designed based on the proposed methodology. To ensure soft switching and minimize turn off loss, the magnetizing inductance is chosen as 14uH. L_n is designed as 17 to utilize the transformer leakage inductance, which results in a 0.8uH resonant inductance. The resonant capacitor can be designed accordingly as 33nF.

The experimental waveforms at resonant frequency and during holdup time are shown in Figure 10 and Figure 11, respectively. From the experimental results, at resonant frequency, soft switching can be achieved with minimum turn off current. By changing L_r and C_r combination, the waveform keeps the same, which verifies the analysis that the loss purely relies on the magnetizing inductance. During holdup time, the converter has to reduce switching frequency to boost up voltage gain. As shown in Figure 12, most of the energy stored in the magnetizing inductor is transferred back to the resonant capacitor. The efficiency curve of the 1MHz LLC resonant converter for different load conditions is shown in Figure 12. The efficiency is more than 92.5% efficiency at full load condition.

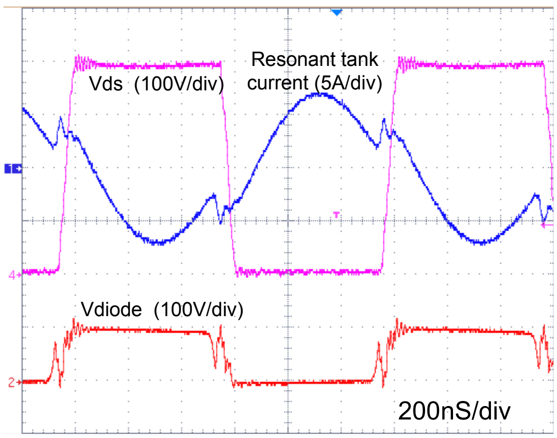

Figure 10. Experimental waveform at resonant frequency

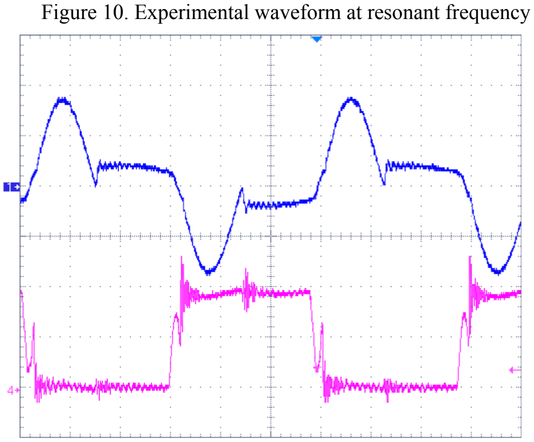

Figure 11. Experimental waveform during holdup time

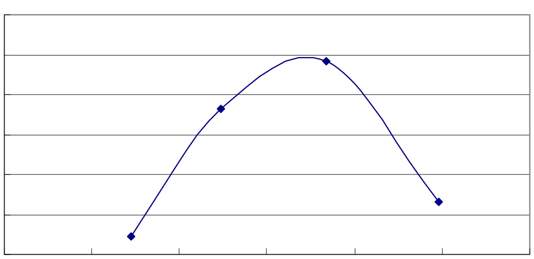

Figure 12. Efficiency of 1MHz LLC resonant Converter

V. SUMMARY

Smaller switching loss and the capability of wide input range operation makes LLC resonant converter attractive for front-end AC/DC converters. However, because of its complexity, lack of design method makes the converter less valuable. In this paper, based on the theoretical analysis on the circuit operation at resonant frequency and maximum gain point, the relationship between the converter loss and operation range has been revealed. Based on the relationship, the converter efficiency can be optimized with a suitable magnetizing inductance design. Moreover, by choose L_n and Q from the peak gain curves, the operation range of the LLC resonant converter can be ensured. The developed methodology has been implemented into a 1MHz 1kW LLC resonant converter design. The experimental results also verify the theoretical analysis.

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