

# Chapter 4

## LLC Resonant Converter

### 4.1 Introduction

In previous chapters, the trends and technical challenges for front end DC/DC converter were discussed. High power density, high efficiency and high power are the major driving force for this application. Hold up time requirement poses big penalty to the system performance. Two methods were proposed in chapter 2 to solve this problem and improve the efficiency. Range winding solution could improve the performance at high input voltage significantly, but with extra devices, windings and control circuit. Asymmetrical winding solution provides a simpler solution, but could only apply to asymmetrical half bridge topology. Also it introduced other problems like discontinuous output current and unbalanced stress.

To catch up with and move ahead of the trend, higher switching frequency, higher efficiency and advanced packaging are the paths we are taking now. Within all these issues, a topology capable of higher switching frequency with higher efficiency is the key to achieve the goal.

With the techniques proposed in chapter 2, the performance at normal operation could be improved. But none of these methods dealt with the switching

loss problem of PWM converter. Even with Zero Voltage Switching technique, the turn on loss could be minimized; turn off loss still limits the capability of the converter to operate at higher switching frequency.

Resonant converter, which were been investigated intensively in the 80's [B1]-[B7], can achieve very low switching loss thus enable resonant topologies to operate at high switching frequency. In resonant topologies, Series Resonant Converter (SRC), Parallel Resonant Converter (PRC) and Series Parallel Resonant Converter (SPRC, also called LCC resonant converter) are the three most popular topologies. The analysis and design of these topologies have been studied thoroughly. In next part, these three topologies will be investigated for front-end application.

## **4.2 Three traditional resonant topologies**

In this part, these three topologies will be evaluated for front end DC/DC application. The major goal is to evaluate the performance of the converter with wide input range. For each topology, the switching frequency is designed at around 200kHz.

### **4.2.1 Series resonant converter**

The circuit diagram of a half bridge Series Resonant Converter is shown in Figure 4.1 [B8]-[B13]. The DC characteristic of SRC is shown in Figure 4.2. The resonant inductor  $L_r$  and resonant capacitor  $C_r$  are in series. They form a series

resonant tank. The resonant tank will then in series with the load. From this configuration, the resonant tank and the load act as a voltage divider. By changing the frequency of input voltage  $V_a$ , the impedance of resonant tank will change. This impedance will divide the input voltage with load. Since it is a voltage divider, the DC gain of SRC is always lower than 1. At resonant frequency, the impedance of series resonant tank will be very small; all the input voltage will drop on the load. So for series resonant converter, the maximum gain happens at resonant frequency.

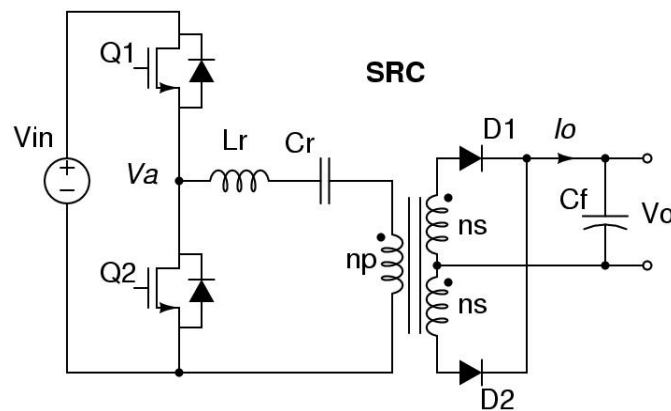


Figure 4.1 Half Bridge Series Resonant Converter

For front end DC/DC application, a SRC is designed to meet the specifications with following parameters:

Transformer turns ratio: 5:2,

Resonant inductance: 37 $\mu$ H,

Resonant capacitance: 17nF.

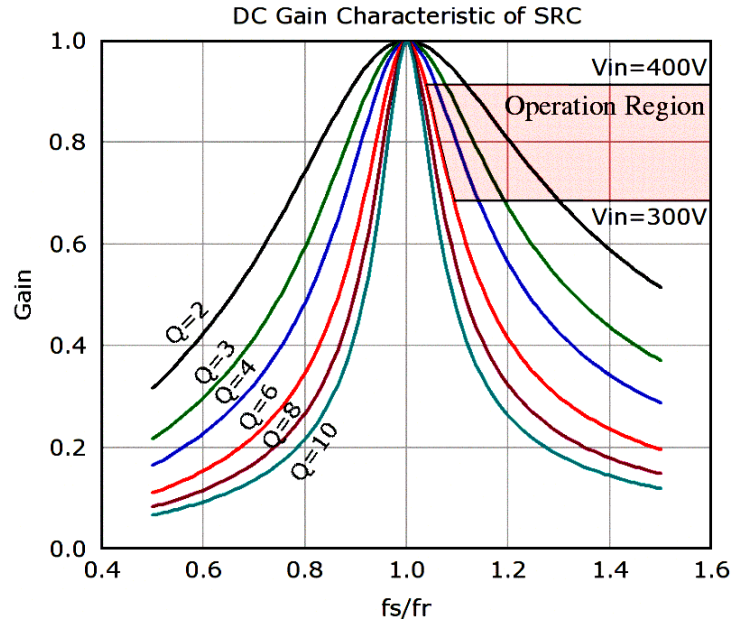


Figure 4.2 DC characteristic and operating region of SRC

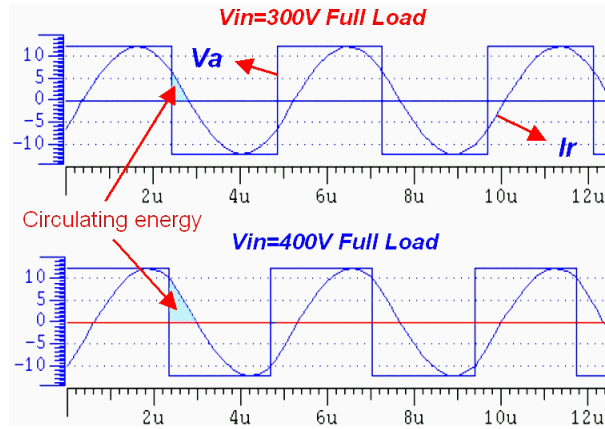


Figure 4.3 Simulation waveforms of SRC

With above parameters, the range of  $Q$  is from 6 (Full load) to 0 (No load). With above design, the operating region of the converter is shown in Figure 4.2 as shaded area. Simulation waveform is shown in Figure 4.3. From the operating region graph and simulation waveforms, several things could be observed:

Operating region is on the right side of resonant frequency  $f_r$ . This is because of zero voltage switching (ZVS) is preferred for this converter. When switching frequency is lower than resonant frequency, the converter will work under zero current switching (ZCS) condition. In fact, the rule is when the DC gain slope is negative; the converter is working under zero voltage switching condition. When the DC gain slope is positive, the converter will work under zero current switching condition. For power MOSFET, zero voltage switching is preferred.

It can be seen from the operating region that at light load, the switching frequency need to increase to very high to keep output voltage regulated. This is a big problem for SRC. To regulate the output voltage at light load, some other control method has to be added.

At 300V input, the converter is working close to resonant frequency. As input voltage increases, the converter is working at higher frequency away from resonant frequency. As frequency increases, the impedance of the resonant tank is increased. This means more and more energy is circulating in the resonant tank instead of transferred to output.

From simulation waveforms, at 300V input, the circulating energy is much smaller than 400V input situation. Here the circulating energy is defined as the energy send back to input source in each switching cycle. The more energy is sending back to the source during each switching cycle, the higher the energy needs to be processed by the semiconductors, the higher the conduction loss.

Also from the MOSFET current we can see that the turn off current is much smaller in 300V input. When input voltage increases to 400V, the turn off current is more than 10A, which is around the same level as PWM converter.

With above analysis, we can see that SRC is not a good candidate for front end DC/DC converter. The major problems are: light load regulation, high circulating energy and turn off current at high input voltage condition.

#### 4.2.2 Parallel resonant converter

The schematic of parallel resonant converter is shown in Figure 4.4 [B14]-[B17]. Its DC characteristic is shown in Figure 4.5. For parallel resonant converter, the resonant tank is still in series. It is called parallel resonant converter because in this case the load is in parallel with the resonant capacitor. More accurately, this converter should be called series resonant converter with parallel load. Since transformer primary side is a capacitor, an inductor is added on the secondary side to match the impedance.

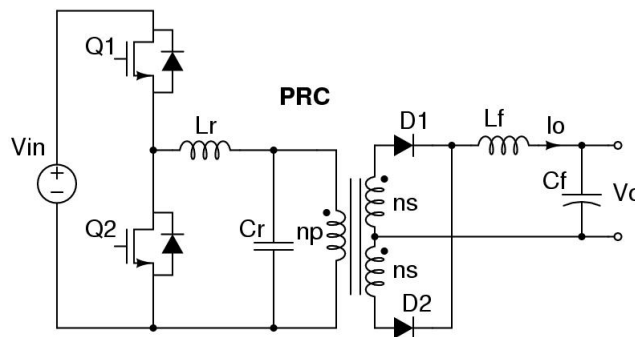


Figure 4.4 Half bridge parallel resonant converter

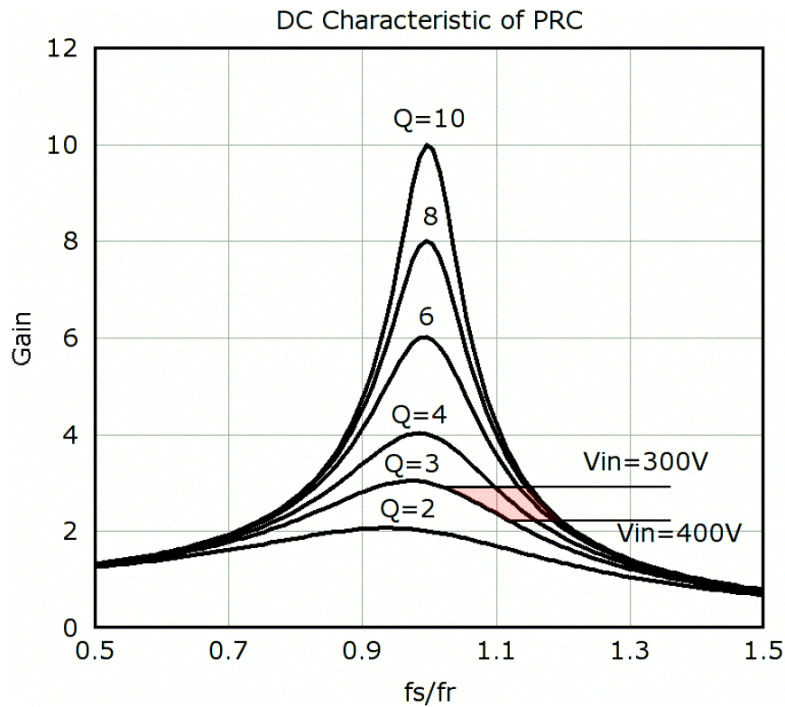


Figure 4.5 DC characteristic and operating region of PRC

The parameters of parallel resonant converter designed for front end DC/DC application are:

Transformer turns ratio: 9:1,

Resonant inductance: 58 $\mu$ H,

Resonant capacitance: 11.7nF.

With above parameters, the range of Q for this converter is 3 (Full load) to  $\infty$  (No load). The operating region of PRC is shown in Figure 4.5 as shaded area. Simulation waveform is shown in Figure 4.6. From the operating region graph and simulation waveforms, several things could be observed:

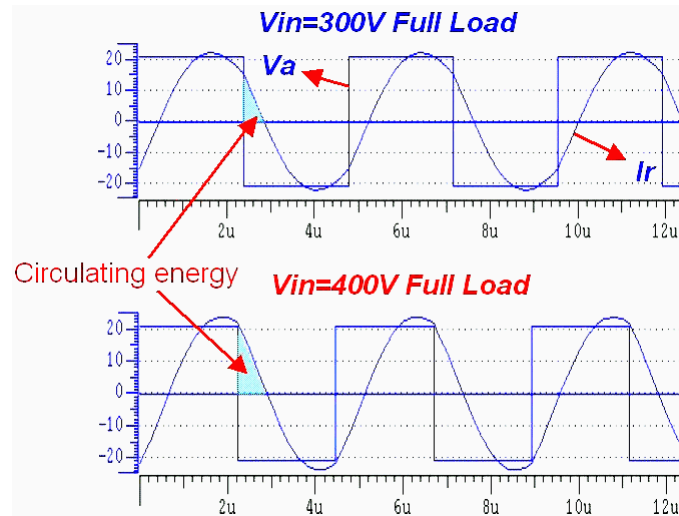


Figure 4.6 Simulation waveforms of PRC

1. Similar to SRC, the operating region is also designed on the right hand side of resonant frequency to achieve Zero Voltage Switching.

Compare with SRC, the operating region is much smaller. At light load, the frequency doesn't need to change too much to keep output voltage regulated. So light load regulation problem doesn't exist in PRC.

Same as SRC for PRC, the converter is working close to resonant frequency at 300V. At high input voltage, the converter is working at higher frequency far away from resonant frequency.

From simulation waveforms, at 300V input, the circulating energy is smaller than 400V input situation. Compare with SRC, it can be seen that for PRC, the circulating energy is much larger. Also from the MOSFET current we can see that the turn off current is much smaller in 300V input. When input voltage



increases to 400V, the turn off current is more than 15A, which is even higher than PWM converter.

For PRC, a big problem is the circulating energy is very high even at light load. For PRC, since the load is in parallel with the resonant capacitor, even at no load condition, the input still see a pretty small impedance of the series resonant tank. This will induce pretty high circulating energy even when the load is zero.

With above analysis, we can see that PRC is not a good candidate for front end DC/DC converter too. The major problems are: high circulating energy, high turn off current at high input voltage condition.

### ***4.2.3 Series parallel resonant converter***

The schematic of series parallel resonant converter is shown in Figure 4.7 [B18]-[B20]. The DC characteristic of SPRC is shown in Figure 4.8. Its resonant tank consists of three resonant components:  $L_r$ ,  $C_s$  and  $C_p$ . The resonant tank of SPRC can be looked as the combination of SRC and PRC. Similar as PRC, an output filter inductor is added on secondary side to match the impedance. For SPRC, it combines the good characteristic of PRC and SRC. With load in series with series tank  $L_r$  and  $C_s$ , the circulating energy is smaller compared with PRC. With the parallel capacitor  $C_p$ , SPRC can regulate the output voltage at no load condition. The parameters of SPRC designed for front end DC/DC application are:

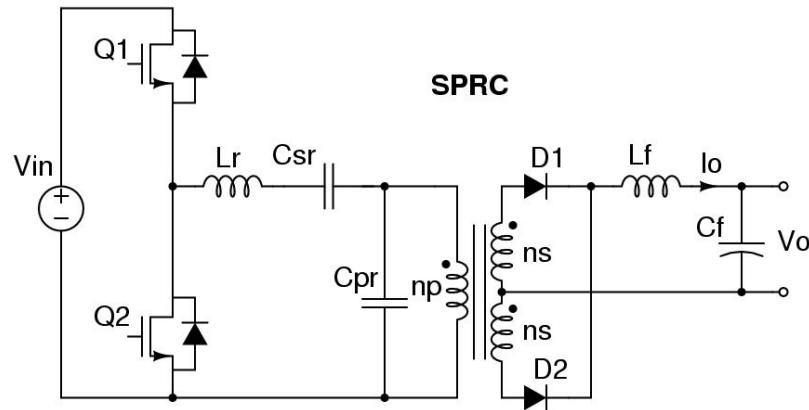


Figure 4.7 Half bridge series parallel resonant converter

Transformer turns ratio: 6:1,

Resonant inductance: 72 $\mu$ H,

Series resonant capacitor Cs: 17.7nF,

Parallel resonant capacitor Cp: 17.7nF,

Range of Q: 1 (Full load) to  $\infty$  (No load)

The DC characteristic and operating region of SPRC are shown in Figure 4.8. Simulation waveform is shown in Figure 4.9. From the operating region graph, several things could be observed:

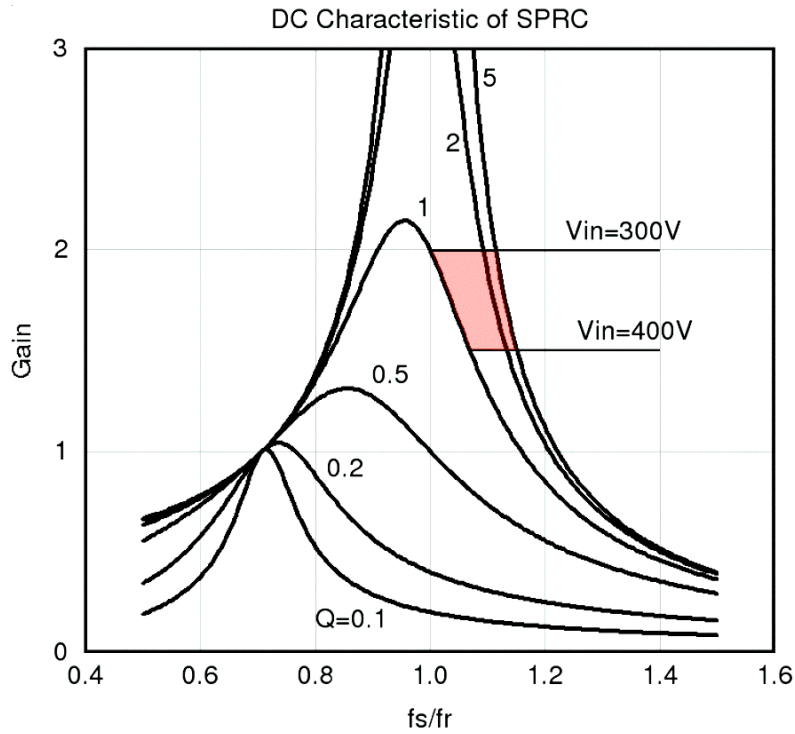


Figure 4.8 DC characteristic and operating region of SPRC

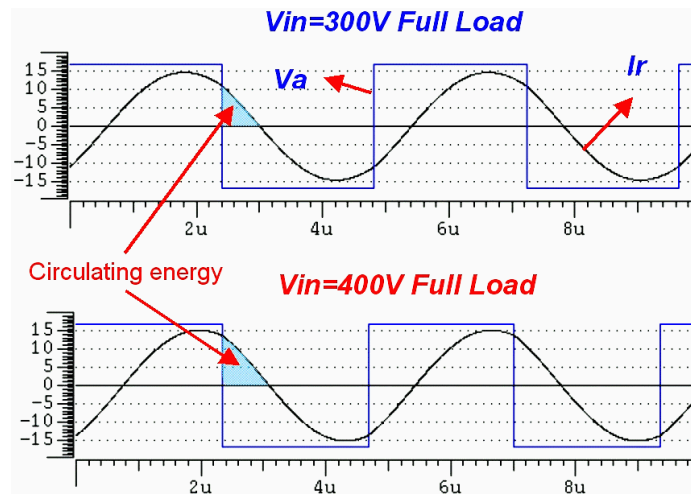


Figure 4.9 Simulation waveforms of SPRC

2. Similar to SRC and PRC, the operating region is also designed on the right hand side of resonant frequency to achieve Zero Voltage Switching.

From the operating region graph, it can be seen that SPRC narrow switching frequency range with load change compare with SRC.

Compare the switching waveforms, the input current in much smaller than PRC and a little larger than SRC. This means for SPRC, the circulating energy is reduced compare with PRC.

Same as SRC and PRC,, the converter is working close to resonant frequency at 300V. At high input voltage, the converter is working at higher frequency far away from resonant frequency.

Same as PRC and SRC, the circulating energy and turn off current of MOSFET also increase at high input voltage. The turn off current is more than 10A.

With above analysis, we can see that SPRC combines the good characteristics of SRC and PRC. Smaller circulating energy and not so sensitive to load change. Unfortunately, SPRC still will see big penalty with wide input range design. With wide input range, the conduction loss and switching loss will increase at high input voltage. The switching loss is similar to that of PWM converter at high input voltage.

By analysis, design and simulation of SRC, PRC and SPRC, the conclusion is that these three converters all cannot be optimized at high input voltage. High conduction loss and switching loss will be resulted from wide input range. To

achieve high switching frequency and higher efficiency, we have to look for some other topologies.

### 4.3 LLC resonant converter

Three traditional resonant topologies were analyzed in above part. From the results, we can see that all of them will see big penalty for wide input range design. High circulating energy and high switching loss will occur at high input voltage. They are not suitable for front end DC/DC application.

Although above analysis give us negative results, still we could learn something from it:

For a resonant tank, working at its resonant frequency is the most efficient way. This rule applies to SRC and PRC very well. For SPRC, it has two resonant frequencies. Normally, working at its highest resonant frequency will be more efficient.

To achieve zero voltage switching, the converter has to work on the negative slope of DC characteristic.

From above analysis, LCC resonant converter also could not be optimized for high input voltage. The reason is same as for SRC and PRC; the converter will work at switching frequency far away from resonant frequency at high input voltage. Look at DC characteristic of LCC resonant converter, it can be seen that there are two resonant frequencies. One low resonant frequency determined by

series resonant tank  $L_r$  and  $C_s$ . One high resonant frequency determined by  $L_r$  and equivalent capacitance of  $C_s$  and  $C_p$  in series. For a resonant converter, it is normally true that the converter could reach high efficiency at resonant frequency. For LCC resonant converter, although it has two resonant frequencies, unfortunately, the lower resonant frequency is in ZCS region. For this application, we are not able to design the converter working at this resonant frequency. Although the lower frequency resonant frequency is not usable, the idea is how to get a resonant frequency at ZVS region. By change the LCC resonant tank to its dual resonant network, this is achievable.

As shown in Figure 4.10, by change  $L$  to  $C$  and  $C$  to  $L$ , a LLC resonant converter could be built. The DC characteristics of these two converters are shown in Figure 4.11 and Figure 4.12. The DC characteristic of LLC converter is like a flip of DC characteristic of LCC resonant converter. There are still two resonant frequencies. In this case,  $L_r$  and  $C_r$  determine the higher resonant frequency. The lower resonant frequency is determined by the series inductance of  $L_m$  and  $L_r$ . Now the higher resonant frequency is in the ZVS region, which means that the converter could be designed to operate around this frequency.

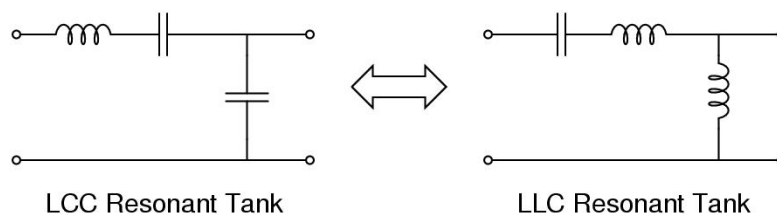


Figure 4.10 LCC and LLC resonant tank

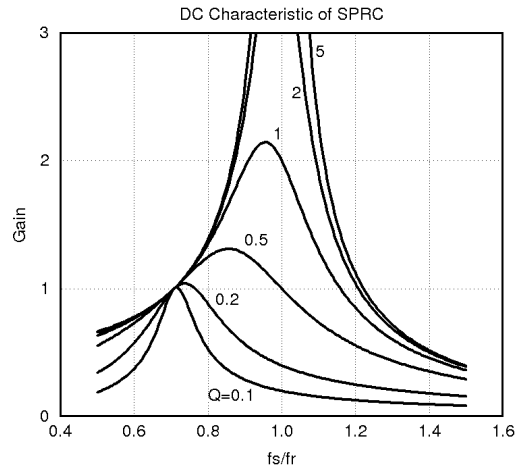


Figure 4.11 DC characteristic of LCC resonant converter

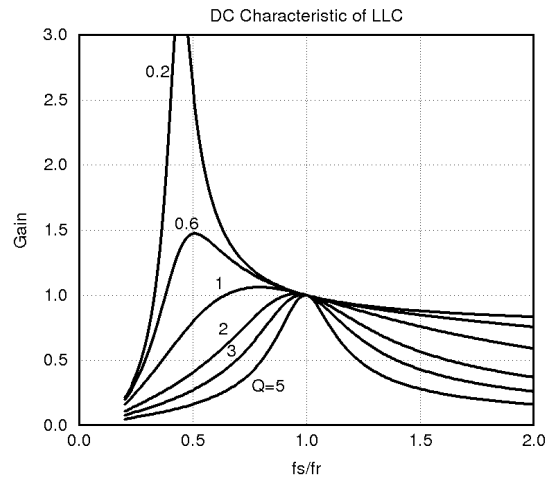


Figure 4.12 DC characteristic of LLC resonant converter

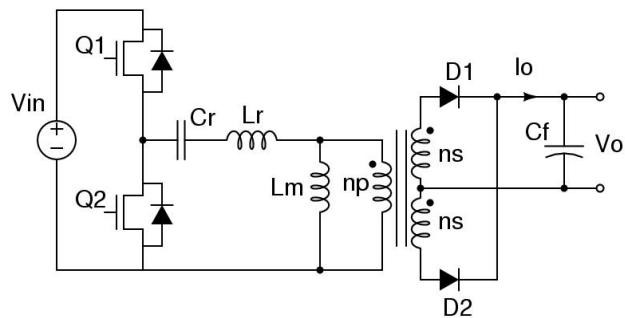


Figure 4.13 Half bridge LLC resonant converter

As a matter of fact, LLC resonant converter existed for very long time [B-10] [B-24]. But because of lack of understanding of characteristic of this converter, it was used as a series resonant converter with passive load. Which means it was designed to operate in switching frequency higher than resonant frequency of the series resonant tank of  $L_r$  and  $C_r$ . When operating in this region, LLC resonant converter acts very similar to SRC. The benefit of LLC resonant converter is narrow switching frequency range with light load and ZVS capability with even no load.

In this dissertation, some unexplored operating region of LLC resonant converter will be investigated. Within these operating regions, LLC resonant converter will have some very special characteristic, which makes it an excellent candidate for front end DC/DC application.

To design LLC resonant converter, DC analysis is essential. The detailed DC analysis is discussed in Appendix B. Two methods were discussed: fundamental simplification method and simulation method. The error is showed too.

#### **4.4 Operation of LLC resonant converter**

The DC characteristic of LLC resonant converter could be divided into ZVS region and ZCS region as shown in Figure 4.14. For this converter, there are two resonant frequencies. One is determined by the resonant components  $L_r$  and  $C_r$ . The other one is determined by  $L_m$ ,  $C_r$  and load condition. As load getting



heavier, the resonant frequency will shift to higher frequency. The two resonant frequencies are:

$$f_{r1} = \frac{1}{2 \cdot \pi \cdot \sqrt{L_r \cdot C_r}}$$

$$f_{r2} = \frac{1}{2 \cdot \pi \cdot \sqrt{(L_m + L_r) \cdot C_r}}$$

With this characteristic, for 400V operation, it could be placed at the resonant frequency of  $f_{r1}$ , which is a resonant frequency of series resonant tank of  $C_r$  and  $L_r$ . While input voltage drops, more gain can be achieved with lower switching frequency. With proper choose of resonant tank, the converter could operate within ZVS region for load and line variation.

There are some interesting aspects of this DC characteristic. On the right side of  $f_{r1}$ , this converter has same characteristic of SRC. On the left side of  $f_{r1}$ , the image of PRC and SRC are fighting to be the dominant. At heavy load, SRC will dominant. When load get lighter, characteristic of PRC will floating to the top. With these interesting characteristics, we could design the converter working at the resonant frequency of SRC to achieve high efficiency. Then we are able to operate the converter at lower than resonant frequency of SRC still get ZVS because of the characteristic of PRC will dominant in that frequency range.

From above discussion, the DC characteristic of LLC resonant converter could be also divided into three regions according to different mode of operation

as shown in Figure 4.15. Our designed operating regions are region 1 and region 2. Region 3 is ZCS region. The converter should be prevented from entering region 3. The simulation waveform for region 1 and region 2 are shown in Figure 4.16 and Figure 4.17. In fact, there are many other operating modes for LLC resonant converter as load changes. Those different modes are listed in Appendix B.

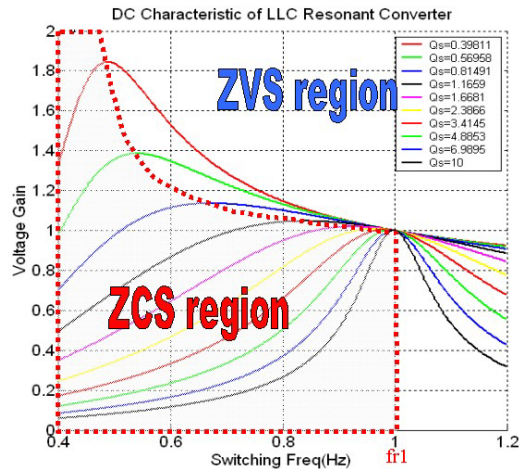


Figure 4.14 DC characteristic of LLC resonant converter

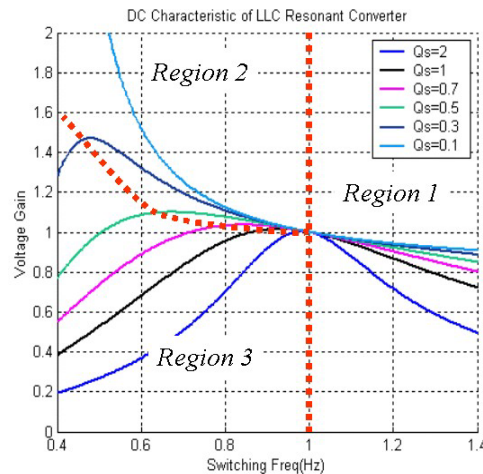


Figure 4.15 Three operating regions of LLC resonant converter

In region 1, the converter works very similar to SRC. In this region,  $L_m$  never resonates with resonant capacitor  $C_r$ ; it is clamped by output voltage and acts as the load of the series resonant tank. With this passive load, LLC resonant converter is able to operate at no load condition without the penalty of very high switching frequency. Also, with passive load  $L_m$ , ZVS could be ensured for any load condition. Here the operation will not be discussed in detail. There are several other modes of operation for light load condition. They will be discussed in Appendix B.

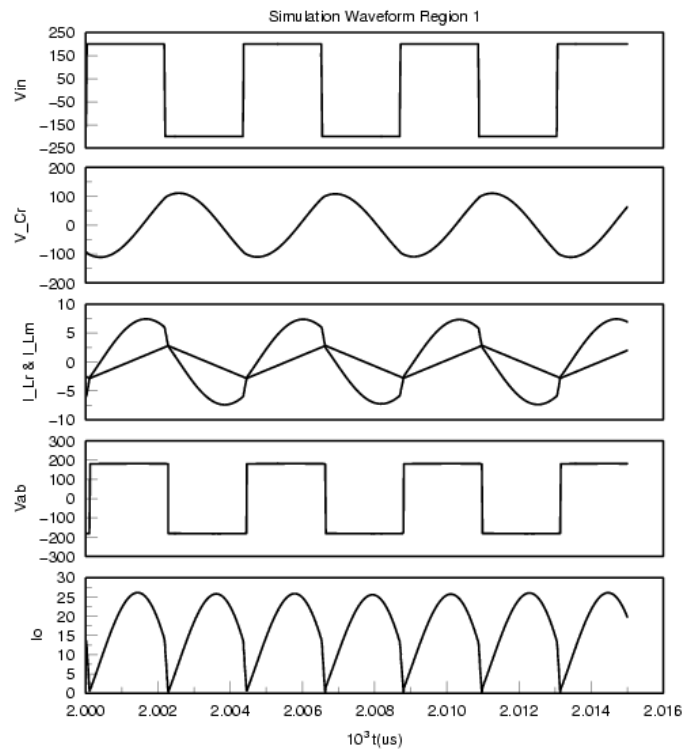


Figure 4.16 Simulated operation waveforms in region 1

In region 2, the operation of LLC resonant converter is more complex and interesting. The waveforms could be divided into clearly two time intervals. In

first time interval,  $L_r$  resonant with  $C_r$ .  $L_m$  is clamped by output voltage. When  $L_r$  current resonant back to same level as  $L_m$  current, the resonant of  $L_r$  and  $C_r$  is stopped, instead, now  $L_m$  will participate into the resonant and the second time interval begins. During this time interval, the resonant components will change to  $C_r$  and  $L_m$  in series with  $L_r$ , which is shown in the waveforms as a flat region. In fact, that is a part of the resonant process between  $L_m+L_r$  with  $C_r$ . From this aspect, LLC resonant converter is a multi resonant converter since the resonant frequency at different time interval is different. Because of the resonant between  $L_m$  and  $C_r$ , a peak on the gain appears at resonant frequency of  $L_m+L_r$  and  $C_r$ . Next the operating of LLC resonant converter in region 2 will be discussed in detail. It is divided into three modes.

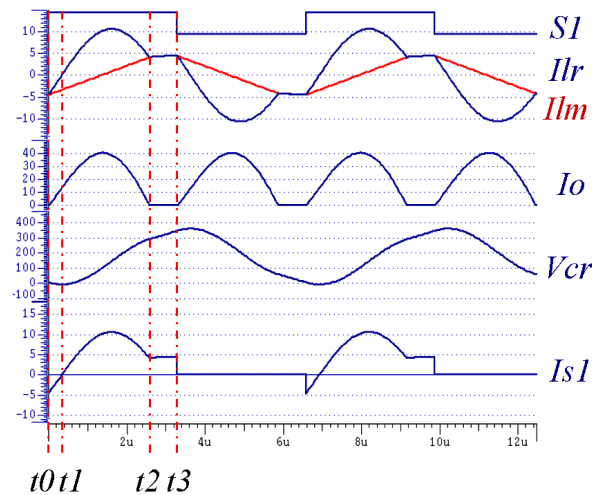


Figure 4.17 Simulation waveforms in region 2

Mode 1 ( $t_0$  to  $t_1$ ):

This mode begins when Q2 is turned off at  $t_0$ . At this moment, resonant inductor  $L_r$  current is negative; it will flow through body diode of Q1, which creates a ZVS condition for Q1. Gate signal of Q1 should be applied during this mode.

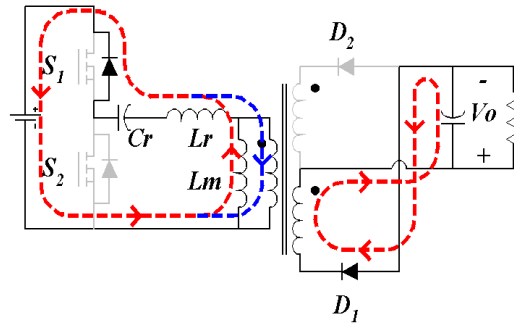


Figure 4.18 Circuit diagram during mode 1 in region 2

When resonant inductor  $L_r$  current flow through body diode of Q1,  $I_{Lr}$  begins to rise, this will force secondary diode D1 conduct and  $I_o$  begin to increase. Also, from this moment, transformer sees output voltage on the secondary side.  $L_m$  is charged with constant voltage.

#### Mode 2 ( $t_1$ to $t_2$ )

This mode begins when resonant inductor current  $I_{Lr}$  becomes positive. Since Q1 is turned on during mode 1, current will flow through MOSFET Q1.

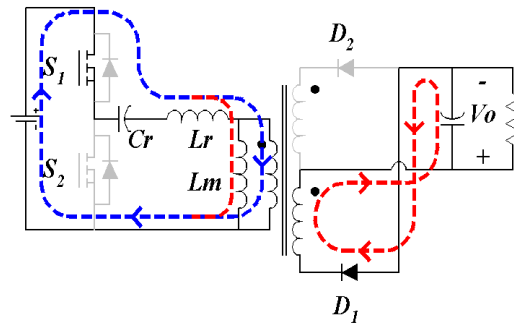


Figure 4.19 Circuit diagram during mode 2 in region 2

During this mode, output rectifier diode  $D_1$  conduct. The transformer voltage is clamped at  $V_o$ .  $L_m$  is linearly charged with output voltage, so it doesn't participate in the resonant during this period. In this mode, the circuit works like a SRC with resonant inductor  $L_r$  and resonant capacitor  $C_r$ .

This mode ends when  $L_r$  current is the same as  $L_m$  current. Output current reach zero.

### Mode 3 ( $t_2$ to $t_3$ )

At  $t_2$ , the two inductor's currents are equal. Output current reach zero. Both output rectifier diodes  $D_1$  and  $D_2$  is reverse biased. Transformer secondary voltage is lower than output voltage. Output is separated from transformer.

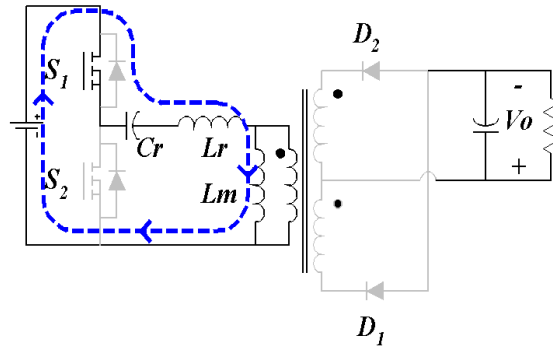


Figure 4.20 Circuit diagram during mode 3 in region 2

During this period, since output is separated from primary,  $L_m$  is freed to participate resonant. It will form a resonant tank of  $L_m$  in series with  $L_r$  resonant with  $C_r$ .

This mode ends when  $Q_1$  is turned off. As can be seen from the waveform,  $Q_1$  turn off current at  $t_3$  is small compare with peak current. For next half cycle, the operation is same as analyzed above.

From the simulation waveform we can see, the MOSFETs are turned on with ZVS. The ZVS is achieved with magnetizing current, which is not related to load current, so ZVS could be realized even with zero-load. Since this magnetizing current is also the turn off current of MOSFET. Choosing different magnetizing inductance could control it. The turn off current could be much smaller than load current, so turn off loss can be reduce. Also, the secondary side diode current reduce to zero and stay off, the reverse recovery is eliminated also. With all these, the switching loss of this converter is very small.

## 4.5 Design of LLC resonant converter

### 4.5.1 Design of power stage parameters

From above analysis, the DC characteristic of LLC resonant converter could be derived. Based on the DC characteristic, parameters in power stage can be designed. The parameters need to be designed are:

- Transformer turns ratio:  $n$
- Series resonant inductor:  $L_r$
- Resonant capacitor:  $C_r$
- Resonant inductor ratio:  $L_m/L_r$

The specifications for the design are:

Input voltage range: 300V to 400V, normal operating region (360-400V)

Output voltage: 48V

Maximum load: 2.5Ohm

Maximum switching frequency: 200kHz

With above information, we can begin to choose the parameters.

For front-end application, the target is to optimize the performance at high input voltage. From previous analysis results, the optimal operating point for this



converter is when switching frequency equals to the resonant frequency of  $L_r$  and  $C_r$ . At this point, the voltage gain of LLC resonant converter is 1. Base on this, the transformer turns ratio can be choose. For Half Bridge LLC resonant converter with 400V input and 48V output, the transformer turns ratio can be choose base on following equations:

$$n = V_{in} / (2 \cdot V_o)$$

For Full Bridge LLC resonant converter, the turn's ratio will be:

$$n = V_{in} / V_o$$

In our design, a half bridge LLC resonant converter is used; the turns ratio was choose to be 4.

After the transformer turns ratio, the resonant tank can be designed. To determine the resonant tank, lot of trade offs are involved. Three design examples will be shown to demonstrate the trade offs.

#### Design 1:

In this design, the ratio of two resonant inductors is 1, which means the two resonant inductors are with same value. The characteristic and operating region are shown in Figure 4.21. The region of  $Q$  is from 1(Full load) to 0(no load). Here  $Q$  is defined as:  $Q = Z_o / R_l$ . Resonant inductor  $L_r$  is 27.8uH, resonant inductor  $L_m$  is 27.8uH and  $C_r$  is 22.8nF.

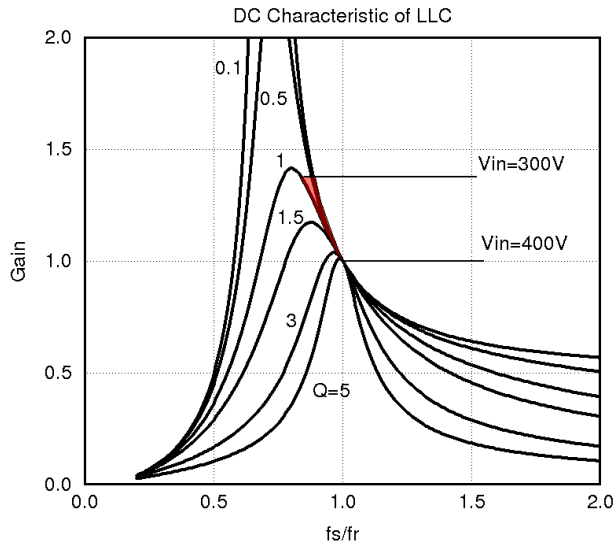


Figure 4.21 Operating region for design 1

Simulation waveform is shown in Figure 4.22.

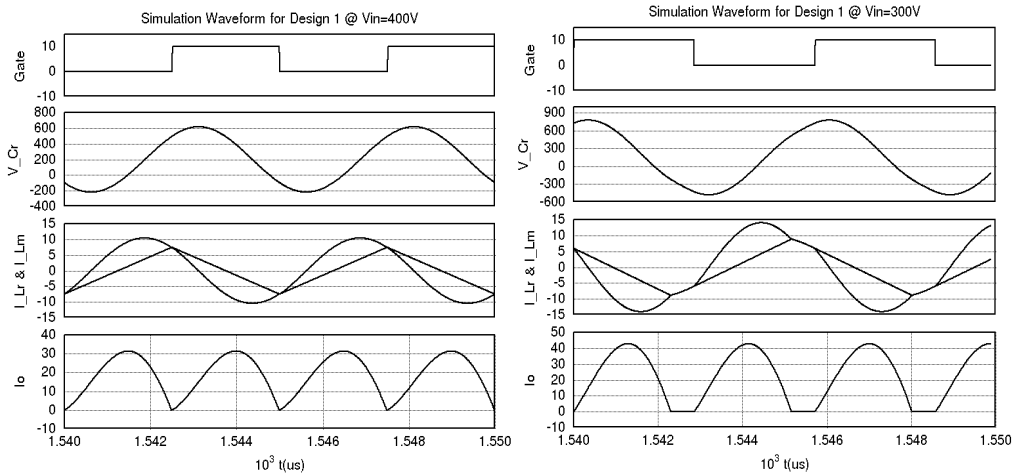


Figure 4.22 Simulation waveforms of design 1 with 300V and 400V input voltage

Design 2:

In this design, the ratio of two resonant inductors is 4, which means  $L_m$  is four times  $L_r$ . The characteristic and operating region are shown in Figure 4.23.

The region of  $Q$  is from 0.5 (Full load) to 0 (no load). Resonant inductor  $L_r$  is 17 $\mu$ H and  $L_m$  is 70 $\mu$ H. Resonant capacitor is 24nF.

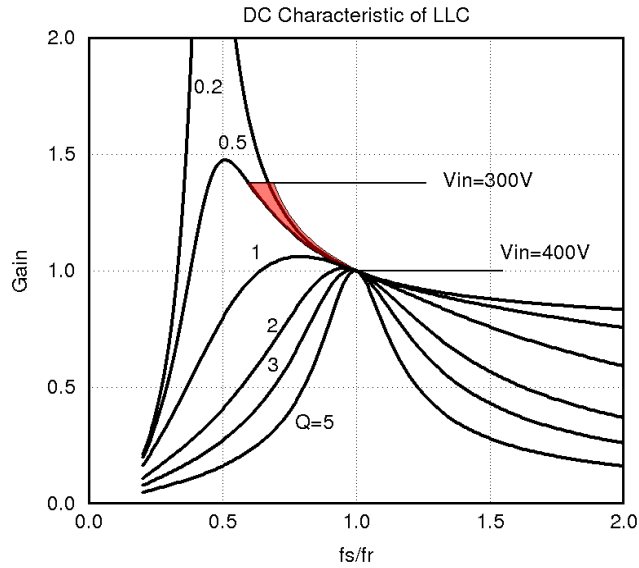


Figure 4.23 Operating region for design 2

Simulation waveform is shown in Figure 4.24.

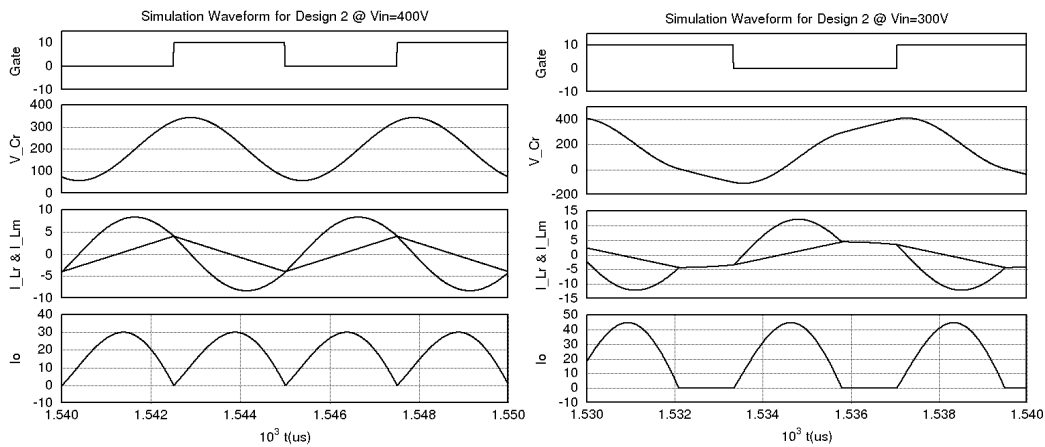


Figure 4.24 Simulation waveforms of design 2 with 300V and 400V input voltage

Design 3:

In this design, the ratio of two resonant inductors is 16, which means  $L_m$  is sixteen times  $L_r$ . The characteristic and operating region are shown in Figure 4.25. The region of  $Q$  is from 0.25 (Full load) to 0 (no load). Resonant inductor  $L_r$  is  $7.95\mu\text{H}$ ,  $L_m$  is  $127\mu\text{H}$  and  $C_r$  is  $79.6\text{nF}$ .

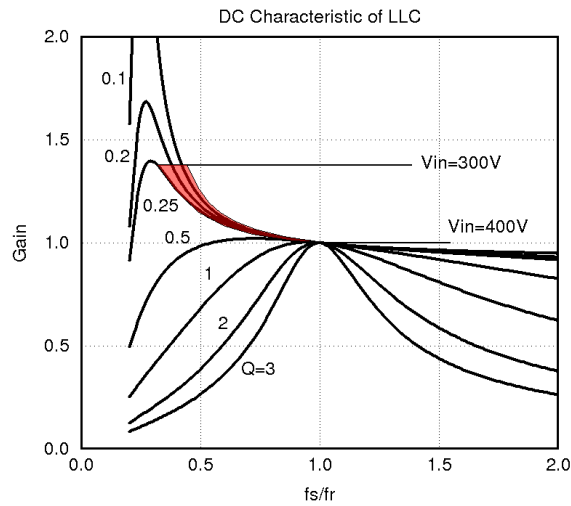


Figure 4.25 Operating region for design 3

Simulation waveform is shown in Figure 4.26.

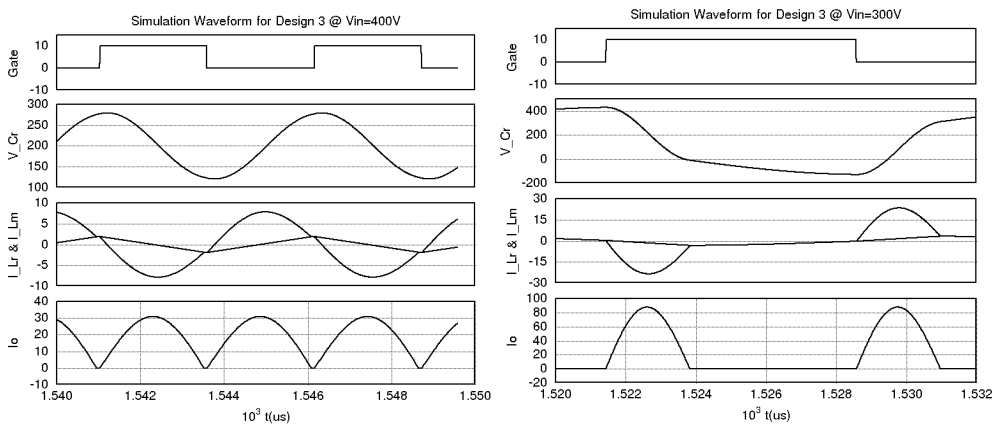


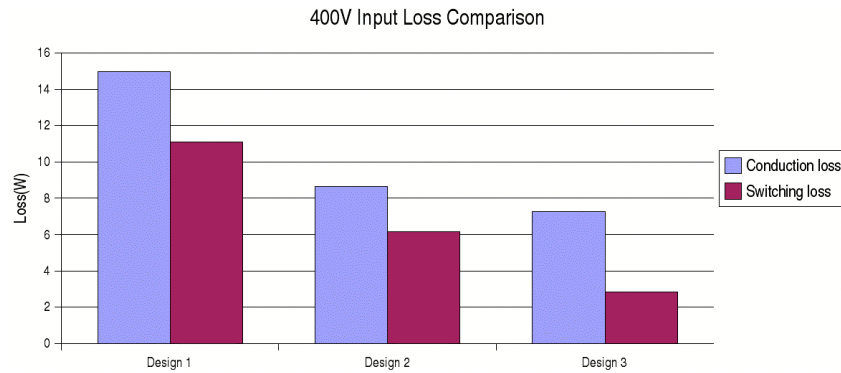
Figure 4.26 Simulation waveforms of design 3 with 300V and 400V input voltage

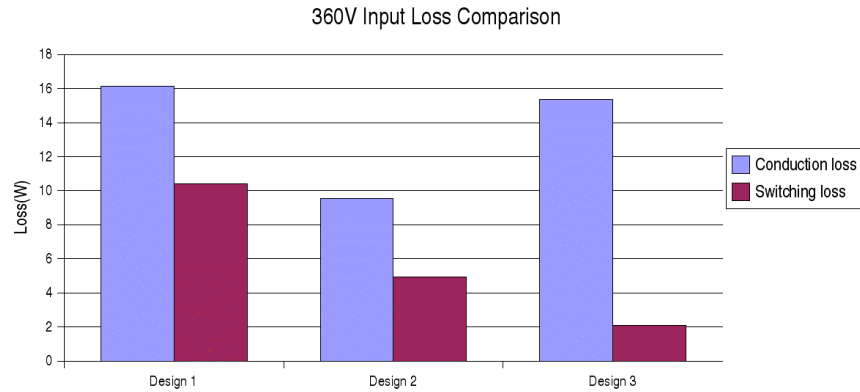
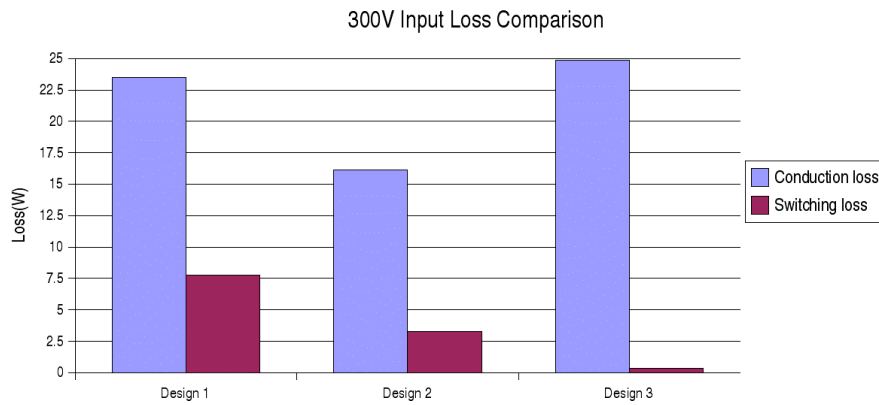
Summary of three designs:

Table 4-1 Summary of three LLC resonant converter designs

	<i>Q</i> range	<i>F</i> <sub>s</sub> range	Primary RMS Current	Switch turn off current	Resonant Cap Voltage	Peak Output current
Design 1	1 to 0	175k to 200k	8.1A to 9.2A	7.8A to 5.8A	800V	31A to 43A
Design 2	0.5 to 0	135k to 200k	6.0A to 8.3A	4.1A to 3.2A	440V	31A to 49A
Design 3	0.25 to 0	72k to 200k	5.7A to 10.2A	1.9A to 0.24A	430V	31A to 89A

From the summary, design 3 provides best performance at 400V input, but the switching frequency range will be much larger. For design 1, the performance at 400V is compromised; the benefit is very narrow switching frequency range. For this application, since the output voltage of PFC circuit is not tightly regulated, it has a range from 360 to 400V. The performance at 360V is also a concern. In Figure 4.27 to Figure 4.29, primary switching and conduction loss are compared for three designs with different input voltage.

Figure 4.27 Primary loss for three designs with  $V_{in}=400V$

Figure 4.28 Primary loss for three designs with  $V_{in}=360V$ Figure 4.29 Primary loss for three designs with  $V_{in}=300V$ 

From these comparisons, although design 3 could provide better performance at 400V input, its performance degrades very fast as input voltage drops. Design 1 could provide more balanced performance for whole range, but the performance at 400V input is greatly impaired. Design 2 is choosing for front-end application with 200kHz design. With design 2, the performance is balanced within input range. Stress on different devices is reasonable.

### 4.5.2 Design trade offs:

In design of power stage, there are some trade offs that will affect the final results.

First trade off is switching frequency range and switching loss. With smaller magnetizing inductance, narrower switching frequency range can be achieved, but switching loss and conduction loss will increase because of high magnetizing current.

Another trade off to make is switching frequency range and resonant tank impedance. For same specification,  $L_r$  and  $C_r$  can have different values, which will work. Although there is a limit on how small  $C_r$  can be in order to keep series resonant tank work in constant gain region. With larger  $C_r$ , the voltage stress on  $C_r$  will be smaller. The problem is that the impedance of the resonant tank will be small too, which will affect the short circuit performance. With smaller tank impedance, the higher the shorts circuit current will be and higher switching frequency is needed to limit the output current.

The problem with low switching frequency is the conduction loss will increase as switching frequency drops. As shown in first part, the conduction loss can be doubled when switching frequency change from 200kHz to 150kHz.

From these trade offs, the optimized design should choose as small resonant capacitor as possible to get enough voltage gain at heavy load. Then  $L_m$  should

be as large as possible to get the voltage gain with desired switching frequency range.

### 4.5.3 Test results

Base on the design, a LLC resonant converter is built with those parameters. The test circuits are shown in Figure 4.30 with the part number of the devices. Test waveforms are shown in Figure 4.31. The test efficiency is shown in Figure 4.32. Compare with asymmetrical half bridge converter, LLC resonant converter could improve the efficiency at normal operation point by more than 3%. Figure 4.33 shows the test efficiency at different input voltage. LLC resonant converter could cover wide input range with much higher efficiency compared with PWM converter.

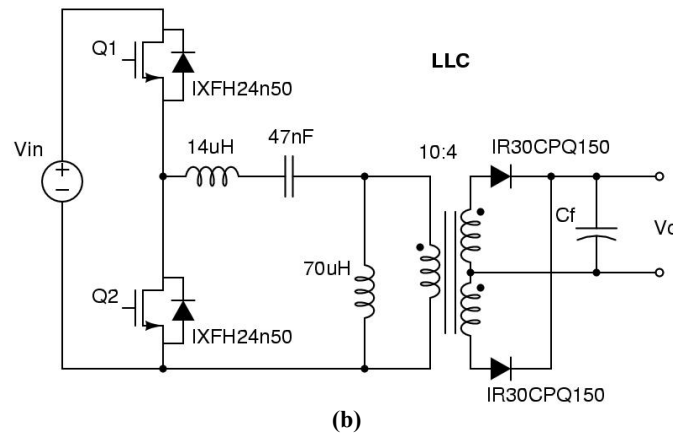


Figure 4.30 Test circuit for 200kHz LLC resonant converter



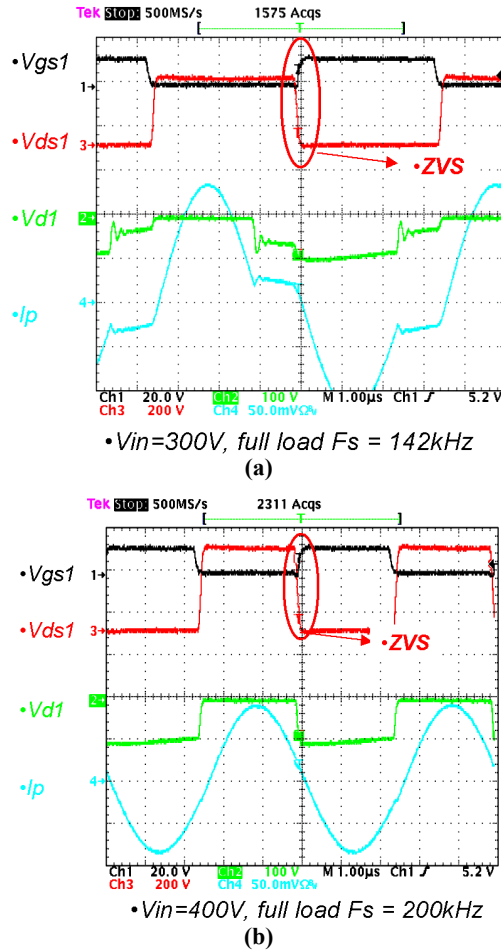


Figure 4.31 Test waveform of LLC converter at full load and (a) 300V input, and (b) 400V input

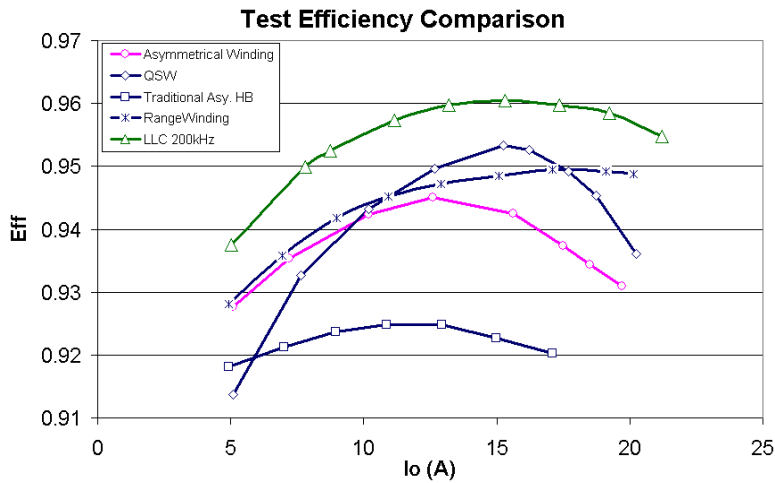


Figure 4.32 Test efficiency of LLC converter and HB converter at 400V input

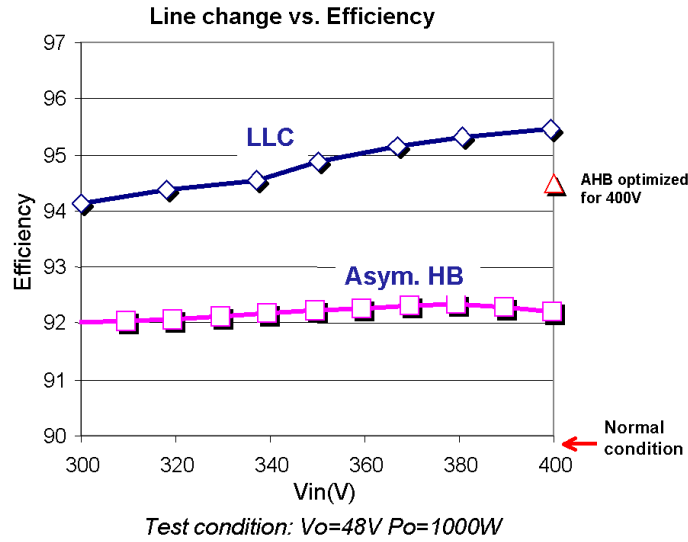


Figure 4.33 Test efficiency of LLC and AHB converter at different input voltage and full load

#### 4.5.4 Loss comparison

Since our target is to achieve high efficiency, and high frequency. In this part, the loss of LLC resonant converter is compared with Asymmetrical Half Bridge and asymmetrical winding asymmetrical half bridge. Comparison is based on simulations and calculations. All converters are designed with same specifications:

$V_{in}=300$  to  $400V$ ,  $V_{out}=48V$ , and  $P_{out}=1kW$ ;

Switching frequency for PWM  $200kHz$  and switching frequency for LLC is from  $140kHz$  to  $200kHz$ .

The design of Asymmetrical half bridge just follows the traditional design procedure. LLC resonant is as shown in Figure 4.30. The circuit diagrams of both converters with devices listed are shown in Figure 4.34.

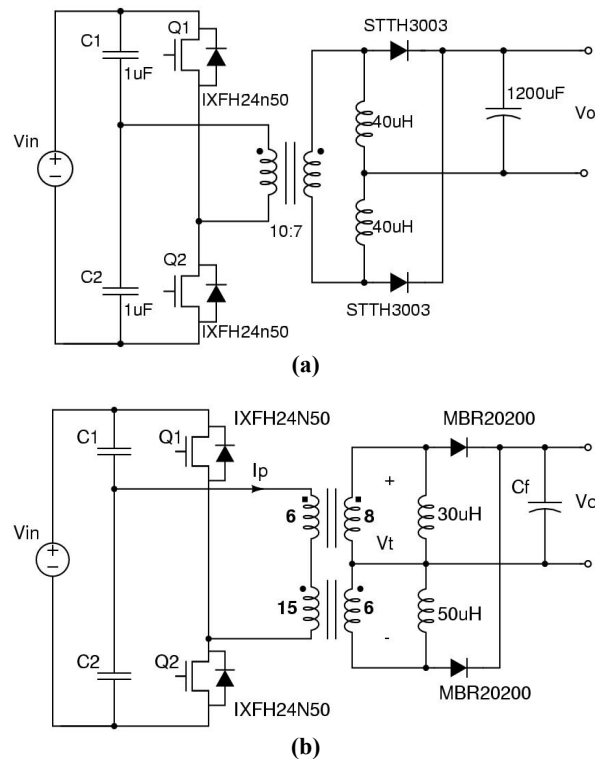


Figure 4.34 Two PWM topologies to be compared (a) AHB, and (b) asymmetrical winding AHB

The devices we used for these converters are:

Primary switches for both converters: IXFN24N50 500V 21A MOSFET

Asymmetrical Half Bridge secondary rectifier: STTH3003, 300V ultra fast diode, forward voltage drop 1.2V.

Asymmetrical winding AHB: MBR20200, 200V shottky diodes with forward voltage drop 0.95V.

LLC resonant converter secondary rectifier: IR30CPQ150, 150V shottky diodes with forward voltage drop 0.65V.

For PWM Half Bridge converter, the current waveform is highly asymmetrical at high input voltage. This will increase both the conduction loss and switching loss when the converter works in this condition. So the efficiency of the converter will be hurt by wide input range. The small duty cycle will also increase the voltage stress of the secondary rectifier. Higher voltage rated devices have to be used which have higher forward voltage drop. Secondary conducting loss will be a large part of total loss.

For LLC resonant converter, at high input voltage, the input current have lower peak value and RMS value, so the conduction loss is much lower at high input voltage. Also, the secondary side voltage stress is fixed at two times output voltage for LLC resonant converter. Low voltage schottky diodes can be used to reduce the secondary conduction loss.

For resonant converter, high current stress or voltage stress is always a concern since the increase of conduction loss will annihilate the benefit get from reduced switching loss.

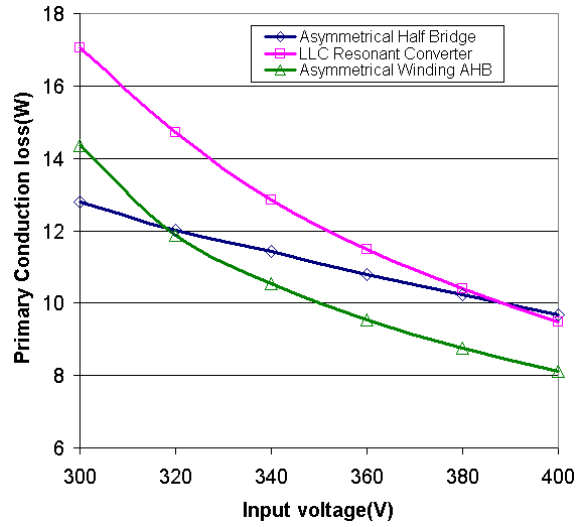


Figure 4.35 Primary side conduction loss comparison

Figure 4.35 shows the primary conduction loss comparison. Because for PWM converter, at high input voltage, the duty cycle is small, so the RMS current is even higher than LLC resonant converter. When the input voltage decrease, the duty will become more symmetrical and the resonant converter will show higher RMS current and higher conduction loss. But this is not a problem for this case since our normal working condition input voltage will be within 360V to 400V. Only during fault condition would the circuit work at so low input voltage. With LLC resonant converter, primary conduction loss is not a problem.

Figure 4.36 shows the primary switching loss comparison of Half Bridge and LLC resonant converter based on the datasheet. It can be seen that switching loss of LLC is 40% lower than PWM converter. This calculation is for 200kHz switching frequency, if we increase the switching frequency further, this will make even bigger difference.

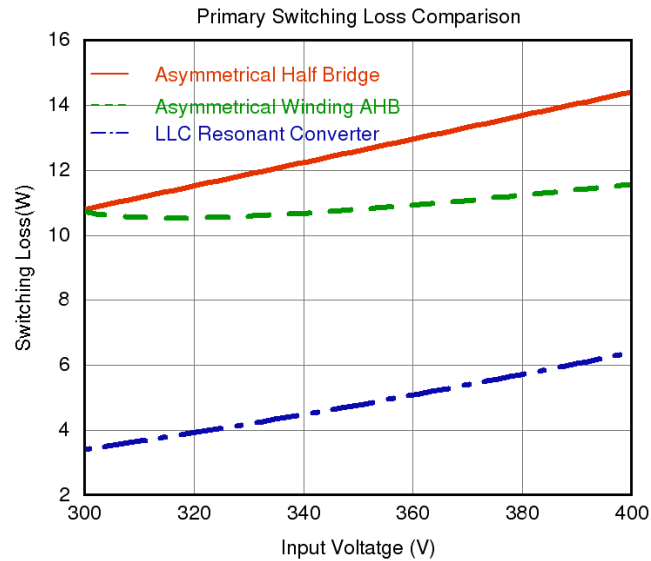


Figure 4.36 Primary switches turn off loss comparison

Another benefit of LLC resonant converter is that the voltage stress on the secondary rectifier is much lower than asymmetrical half bridge. For Asymmetrical half bridge converter, the highest voltage stress on secondary diode will be higher than 250V; 300V diode has to be used. With asymmetrical winding, the voltage stress could be reduced so that 200V diodes could be used, still the forward voltage drop is pretty high at 0.95V. For LLC converter, the voltage stress is limited to two times output voltage, so 150V shottky diode can be used. In this comparison, STTH3003, which has forward voltage drop of 1.2V, is used for PWM converter. For LLC converter, we choose 30CPQ150, which has forward voltage drop of 0.65V. This will reduce the secondary conduction loss from 25W to 13W. This means 1% efficiency improvement.

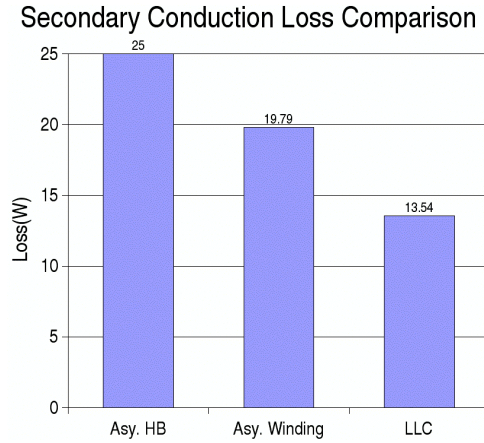


Figure 4.37 Secondary conduction loss comparison

There is still something not included into this analysis: the rectifier diodes reverse recovery loss. In PWM converter, the diodes commutate at load current with high  $dv/dt$ ; this will cause pretty high reverse recovery loss. In LLC resonant converter, since the diode current resonant to zero, this part of loss is eliminated. From the test results, there is no ringing on the secondary diodes without any snubber circuits.

Table 4-2 Loss breakdown comparison of AHB and LLC

	Asy. HB	Asy. Winding	LLC
Primary Conduction Loss	9.7	8.1	9.5
Primary Switching Loss	14.4	11.6	6.4
Secondary Conduction Loss	25	19.7	13.5
Secondary Switching Loss	?	?	0
<b>Total</b>	<b>49.1 + ?</b>	<b>39.4 + ?</b>	<b>29.4</b>

Table 4-2 shows the loss breakdown for asymmetrical half bridge and LLC resonant converter. Without considering secondary reverse recovery loss, LLC

resonant converter can provide about 2% efficiency improvement over asymmetrical half bridge.

For both topologies, we can achieve zero voltage turn on, so we assume turn-on loss could be neglected. In fact, for PWM converter, we are using the energy stored in leakage inductance to achieve ZVS, so at light load, the converter will loss ZVS capability. But for LLC resonant converter, we are using magnetizing inductor to achieve ZVS, so we are able to get ZVS in whole load range.

From previous discussion we can see that with LLC resonant converter, we can get high efficiency and high switching frequency, which is the key to meet the trend for this application.

#### **4.6 Extension of LLC resonant topology**

As demonstrated in previous part, the benefit of LLC resonant converter comes from the fact that it has two resonant frequencies and the operating point of normal operating condition could be positioned at resonant frequency.

In three elements resonant tanks, there are 36 different configurations. The question is if there is any other resonant tank configuration could provide similar characteristic so that could result to similar benefits. To answer this question, a thorough search is performed for three resonant components resonant tank as shown in Appendix A. Since our application is a voltage source input application. Only resonant tank could be used for voltage source input are look in detail,



which consists 23 resonant tank configurations. With in these 23 resonant tank configurations, three resonant tanks show the desired characteristic as shown in Figure 4.38. Tank G is the same resonant tank as we used for LLC resonant converter. It is interesting to notice that the other two resonant tanks also have two resonant inductors and one resonant capacitor.

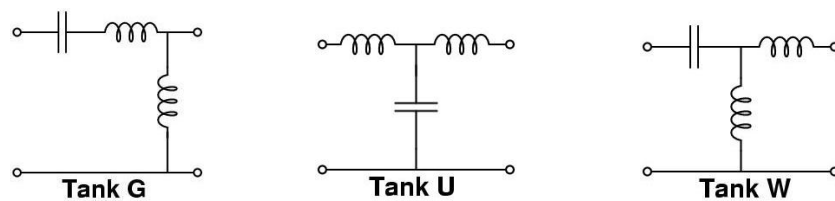


Figure 4.38 Resonant tanks with desired DC characteristic

For resonant tank W, a resonant converter could be constructed as shown in Figure 4.39. With this converter, the series resonant inductor will be on secondary side. This topology operates same as LLC resonant converter discussed before except the voltage stress on the output rectifier diodes will be different as will be discussed in magnetic design in next chapter. With resonant tank U, a half bridge resonant topology could be constructed as shown in Figure 4.40.

These three topologies have similar characteristic. They will be able to operate at resonant frequency when input voltage is high while still cover wide input range.

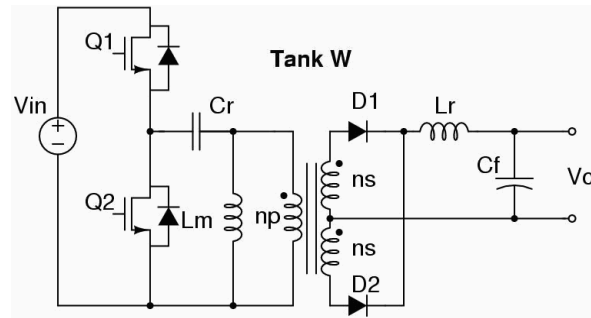


Figure 4.39 Half bridge resonant converter with tank W

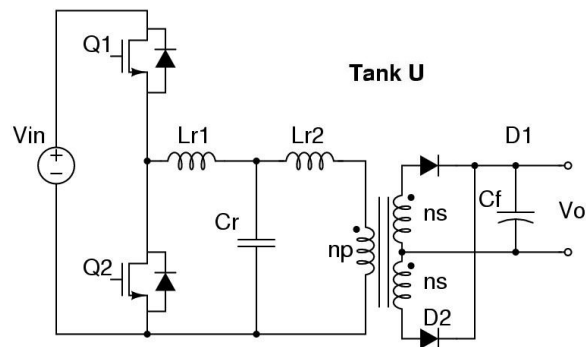


Figure 4.40 Half bridge resonant converter with tank U

## 4.7 High frequency operation

LLC resonant converter shows much lower switching loss compared with PWM converter as shown in previous part. Compare with PWM converter, the primary switching loss is reduced by more than 50%. Secondary diode commutate naturally so there is no reverse recovery problem at all. All these enable LLC resonant converter to be able to operate with higher switching frequency. Another significant difference of LLC resonant converter and PWM converter is that now switching loss could be controlled.

For PWM converter, switching loss is determined by device and load condition. To deliver the power, the switches have to turn off significant amount

of current. For asymmetrical half bridge, it is around 13A of current. This will introduce very high switching loss. This is also one limitation why PWM converter could not operate with higher switching frequency.

For LLC resonant converter, switching loss is a controlled parameter. As shown in Figure 4.41, the turn off current of primary switching is determined by the choice of resonant inductor  $L_m$ . With larger  $L_m$ , turn off current will be smaller, vice versa. This could also be observed from the design examples shown in previous section.

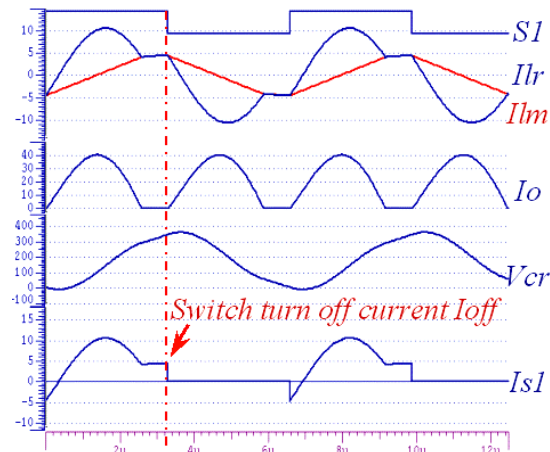


Figure 4.41 Simulation waveform of LLC resonant converter

In this part, a 400kHz LLC resonant converter is designed to demonstrate the benefits of LLC resonant converter with high switching frequency. The resonant tank is scale down as designed for 200kHz LLC resonant converter. For design of  $L_m$ , balance of switching loss and switching frequency range is considered. As shown in Figure 4.42, with smaller  $L_m$ , switching loss will increase. In this case, 40uH  $L_m$  is chosen which gives less than 10W switching loss.

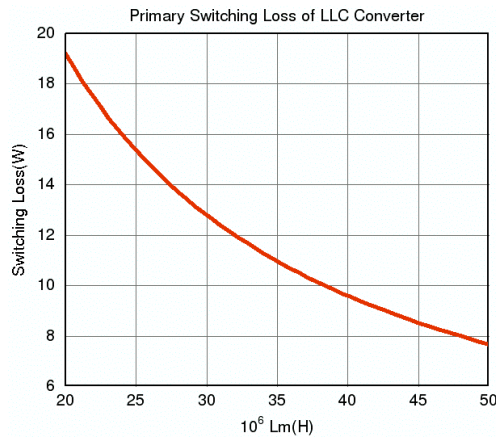


Figure 4.42 Switching loss with different Lm for 400kHz LLC resonant converter

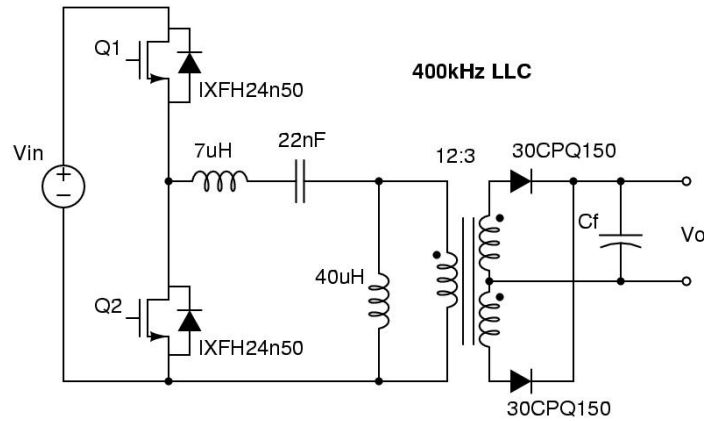


Figure 4.43 Circuit diagram of 400kHz LLC resonant converter

The magnetic design use integrated magnetic concept, which will be discussed in next chapter.

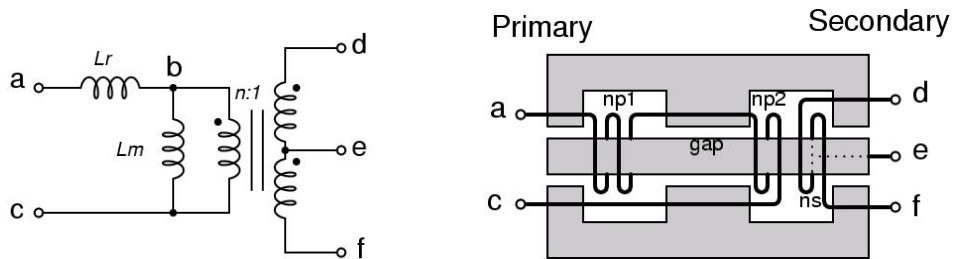


Figure 4.44 Integrated magnetic structure for 400kHz LLC resonant converter

The parameters are as following:

Turns ratio: 12:3;  $L_r$ : 7 $\mu$ H;  $L_m$ : 40 $\mu$ H.

Design results are:  $n_{p1} = 6$ ,  $n_{p2} = 10$ ,  $n_s = 3$  and gap = 0.5mm.

The magnetic components for 200kHz LLC and 400kHz LLC resonant converter are shown in Figure 4.45. Both magnetic components use integrated magnetic concept, which will be discussed in next chapter. With 400kHz switching frequency, volume of the magnetic component is reduced by more than 40%. The test circuits are shown in Figure 4.46. Compare 200kHz LLC resonant converter with 200kHz AHB, the power density could be improved by almost 100%. With high frequency, the power density is improved by 200%.

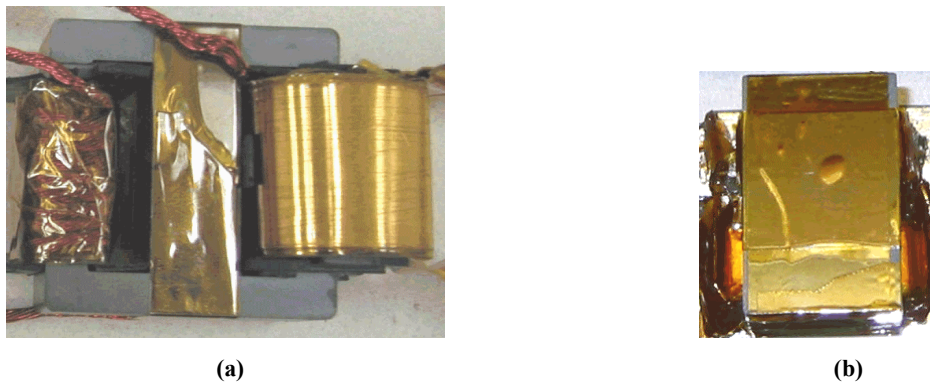
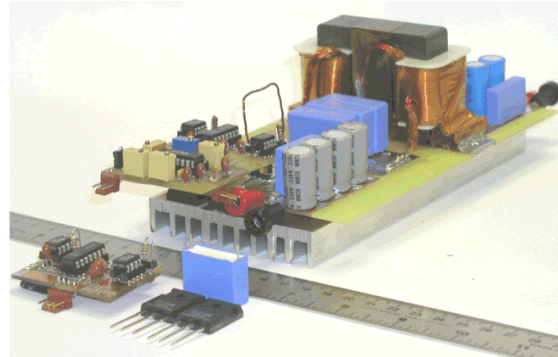
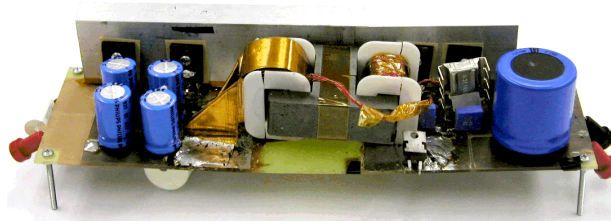


Figure 4.45 Magnetic components size comparison (a) 200kHz and (b) 400kHz



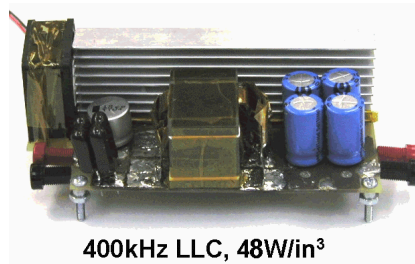
**AHB with Integrated Magnetic 13W/in<sup>3</sup>**

(a)



**200k LLC Converter 28W/in<sup>3</sup>**

(b)



**400kHz LLC, 48W/in<sup>3</sup>**

(c)

Figure 4.46 Test circuits of front end DC/DC converter (a) 200kHz AHB, (b) 200kHz LLC, and (c) 400kHz LLC

Figure 4.47 shows the test efficiency. With 400kHz switching frequency, the efficiency is lower than 200kHz. But compare 400kHz LLC resonant converter with 200kHz PWM converter, the efficiency of LLC resonant converter is still better. The projected 400kHz PWM converter uses the test data of 200kHz PMW converter with consideration of switching loss of primary switches at 400kHz. In

fact, the efficiency will be even lower with consideration of secondary diode reverse recovery loss.

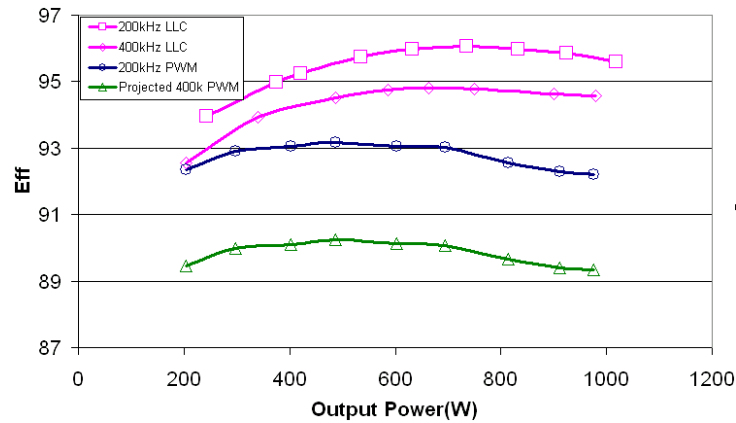


Figure 4.47 Test efficiency of 400kHz LLC resonant converter

## 4.8 Summary

In this chapter, resonant topologies were investigated for front end DC/DC application. The target is to find a topology, which could be optimized at high input voltage with low switching loss. First three traditional resonant topologies were designed and simulated for this application. They are: series resonant converter, parallel resonant converter and series parallel resonant converter. Unfortunately, all three converters suffered from wide input range problem and could not be optimized for high input voltage.

LLC resonant topology was introduced by observing the tank characteristic of series parallel resonant converter. With LLC resonant converter, performance at high input voltage could be optimized and the converter still could cover wide input voltage range. Test results shows that with LLC resonant converter, the

efficiency could be improved by more than 3% compared with PWM converter. With high switching frequency, LLC resonant converter could achieve  $45\text{W/in}^3$  compared with  $13\text{W/in}^3$  of PWM converter.

From analysis and test, LLC resonant converter is proved to be able to improve the performance of front end DC/DC converter significantly. Another question is that if this is the only resonant tank or there is more resonant tank, which could provide similar characteristic.