Chapter 5

Improvements of LLC Resonant Converter

From previous chapter, the characteristic and design of LLC resonant converter were discussed. In this chapter, two improvements for LLC resonant converter will be investigated: integrated magnetic design and over load protection.

5.1 Magnetic design for LLC Resonant Converter

From previous discussion, the power stage could be designed according to the given specifications. The outcome of the design is the desired values for the components. For these components, power devices and capacitors are obtained from manufactures, which already reflect the state of the art technology. Within all these components, magnetic is the one need to be physically designed and built by power electronics researcher. In this part, the design of magnetic component for LLC resonant converter will be discussed.

5.1.1 Discrete design and issues

For a LLC resonant converter, the magnetic components need to be designed are shown in Figure 5.1. There are three magnetic components: Lr, Lm and transformer T. From the configuration of Lm and transformer T, it is easy to build

Lm as the magnetizing inductance of transformer. So in fact, we are trying to build one resonant inductor and one transformer with magnetizing inductance.

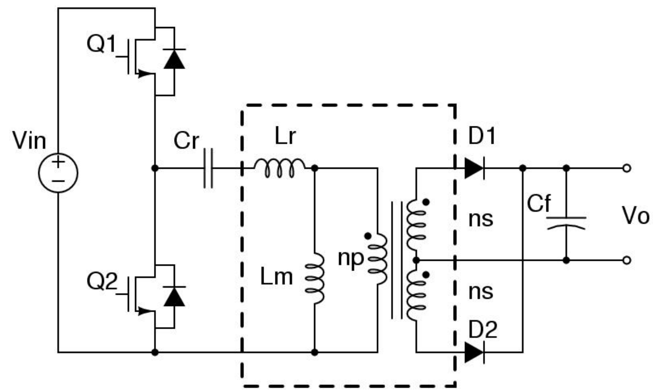


Figure 5.1 Magnetic structure for LLC resonant converter

There are several ways to build them. One is using discrete components, with one magnetic core to build the resonant inductor and one magnetic core to build the transformer and magnetizing inductor Lm. The benefit of this method is that the design procedure is well established.

Next, a discrete design is presented and simulation result is showed to provide a reference for later integrated magnetic designs. For LLC resonant converter, the resonant inductor Lr has pure AC current through it, so we use soft ferrite core for both inductor and transformer.

Figure 5.2 shows the discrete design of the magnetic for LLC resonant converter. Two U cores were used to build the resonant inductor and gapped transformer. Fig.6 shows the simulation results of flux density in the core. For each U core, the cross-section area is $116.5mm^2$. Design result: nl=12, np: ns: ns=16:4:4, gap1=1.45mm and gap2=0.6mm.

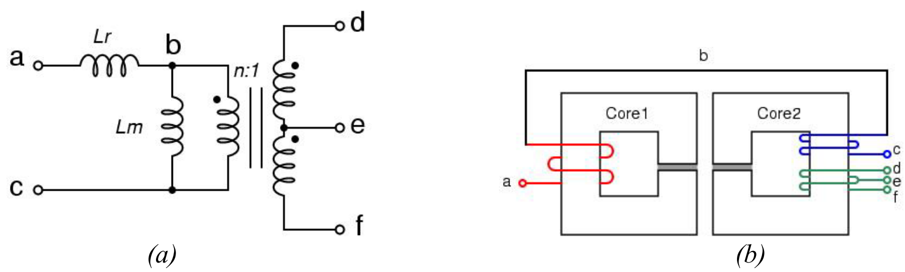
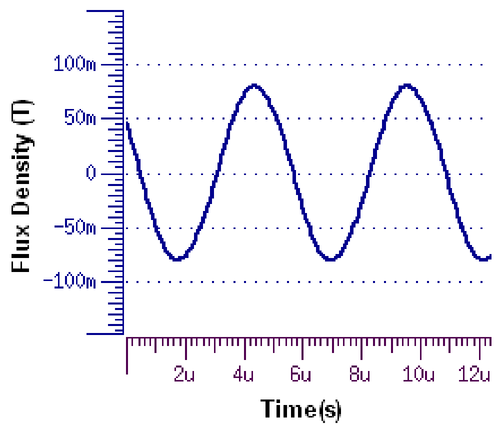
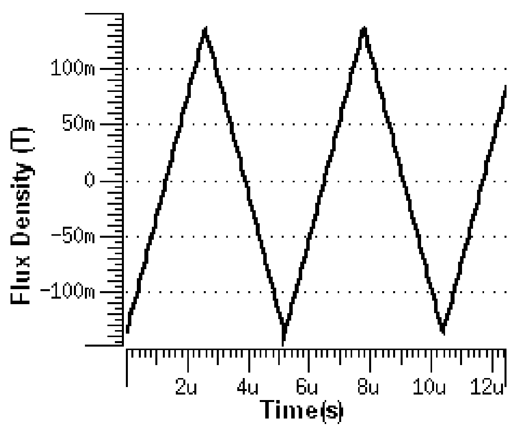


Figure 5.2 Discrete magnetic design (a) schematic (b) physical structure



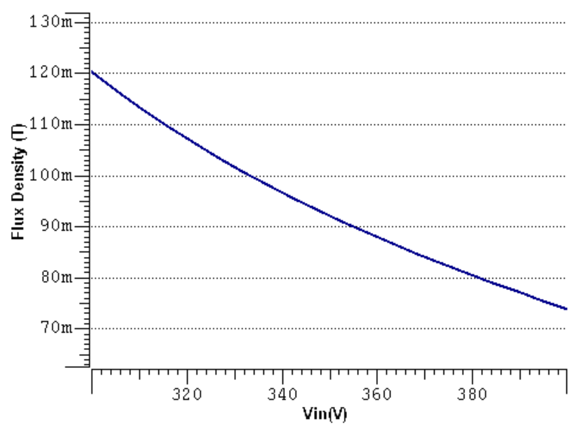
(a) Inductor



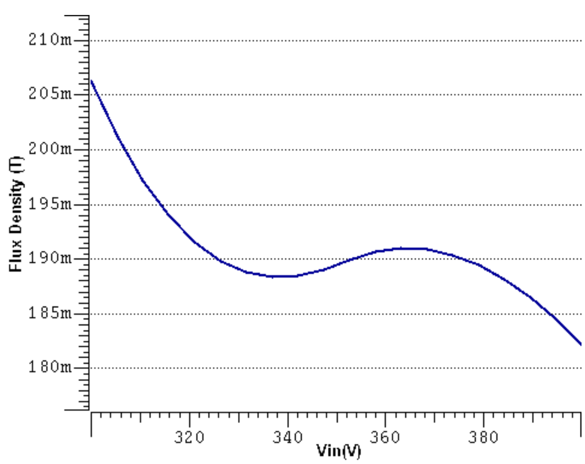
(b) Transformer

Figure 5.3 Flux density simulation result (a) Inductor, and (b) Transformer

Figure 5.3 shows the flux density in each core at 400V input with switching frequency at 200kHz. As seen in the graph, the flux densities in both cores are pretty high. Both cores with high flux density excitation will contribute to the total core loss. For high frequency, core loss is a major limitation on pushing to higher frequency and smaller size. Figure 5.4 shows the peak-to-peak flux density for each core with different input voltage. At low input voltage, the flux density will increase, but it is not critical because of short operating time.



(a) Inductor



(b) Transformer

Figure 5.4 Peak to peak flux density under different input voltage at full load

The drawbacks of this method are: 1. Two magnetic cores are needed, which results in more components count and connections, 2. High magnetic loss caused by high flux ripple in magnetic structure, 3. Large footprint is needed for the whole structure.

In recent years, integrated magnetic has been investigated for many different applications. For asymmetrical half bridge with current doubler, all the magnetic components could be integrated into one magnetic structure with integrated magnetic concept [C1][C5]. In this part, the integrated magnetic structure will be discussed for LLC resonant converter. It integrated all magnetic components into one magnetic core. Through magnetic integration, the component count and footprint are reduced, the connections is also reduced. With proper design; flux ripple cancellation can be achieved, which can reduced the magnetic loss, and reduce the magnetic core size.

In the next part, the integrated magnetic designs for LLC resonant converter will be discussed and compared.

5.1.2 Integrated magnetic design

5.1.2.1 Real transformer with leakage and magnetizing inductance

First structure is just use one transformer and uses the leakage inductance as resonant inductor. The configuration of magnetic components for LLC resonant converter is exactly the same as a real transformer with magnetizing inductance and leakage inductance. It is natural to think about using one real transformer to get all the needed components. The issues with structure are:

1. The leakage inductance cannot be accurately controlled which will determine the operating point of the converter,

2. When we build Lr this way, the leakage inductance will not only exist on primary side, it will also exist on secondary side of the transformer. So the result get from real transformer will be as in Figure 5.5. L_{lp} and L_{ls} have similar value when transferred to same side of the transformer.

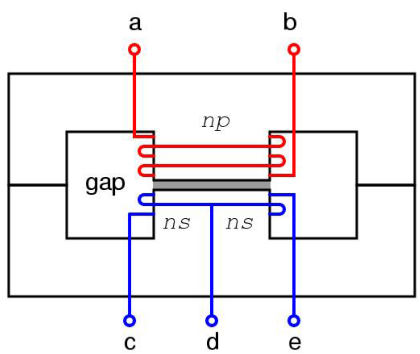


Figure 5.5 Structure with real transformer

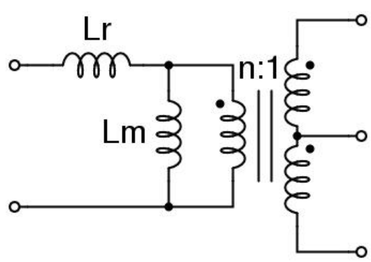


Figure 5.6 Desired magnetic components configuration

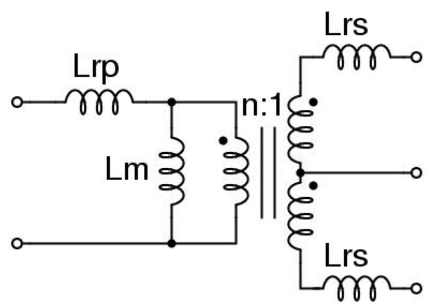
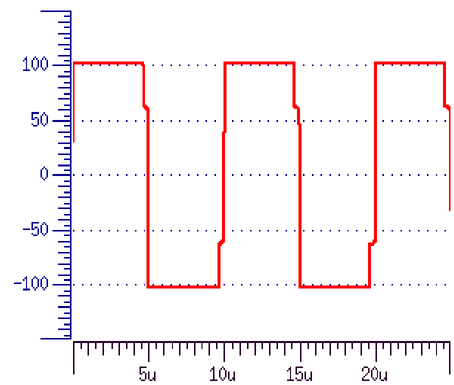
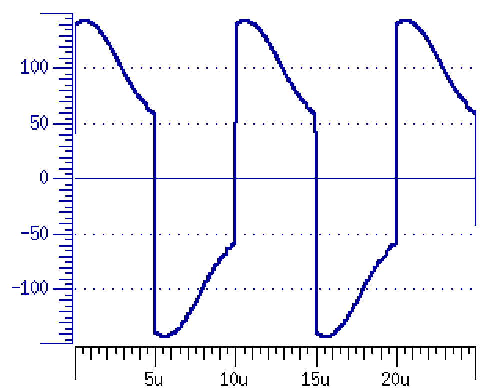


Figure 5.7 Magnetic components configuration from real transformer



(a)



(b)

Figure 5.8 Voltage stress of output diodes D1 D2 (a): desired structure (b) real transformer

When the leakage inductance exists on secondary side, it will increase the voltage stress on secondary rectifier diode. This requires us to use higher voltage rating diode, which will increase the conduction loss of the output rectifier. Figure 5.8 shows the simulate waveforms of secondary diodes voltage stress with magnetic structure in Figure 5.6 and Figure 5.7. We can see that with inductor on the secondary side, the voltage stress of the diodes is much higher.

From above discussion, we can see that the desired magnetic structure will need to provide accurate control of Lr and Lm, at that same time, minimize the inductance on secondary side, which could not be achieved with just a transformer with leakage and magnetizing inductance. Next more complex integrated magnetic structure will be investigated.

5.1.2.2 Integrated magnetic design A

From discrete design, just combine them together with an EE core, we will be able to integrate the two components into one magnetic component as shown in Figure 5.9.

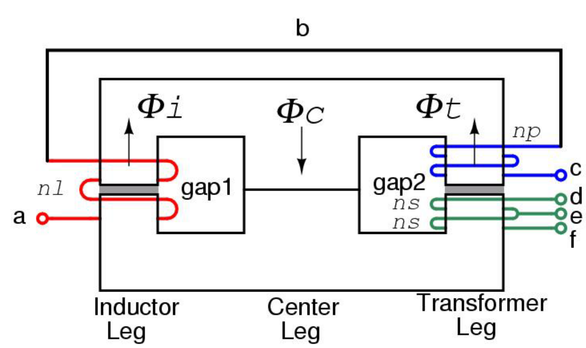


Figure 5.9 Integrated Magnetic Designs A

E42/21/20 core is used. The cross-section area of is 233mm^2. For the outer legs, they have same cross-section area as discrete design. Turn number nl, np and ns is the same as in discrete design. For this design, the inductor and transformer design is decoupled. Discrete design procedure still can be used. Figure 5.10 shows the simulation result of for this structure.

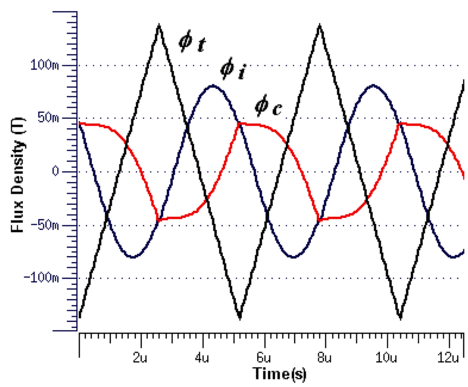


Figure 5.10 Flux density simulation result for Design A

It can be seen from the simulation result: for inductor and transformer leg, the flux density is the same as discrete design. But for center leg, the flux density is much smaller than discrete case. This will greatly reduce the magnetic loss in the big part of the magnetic component.

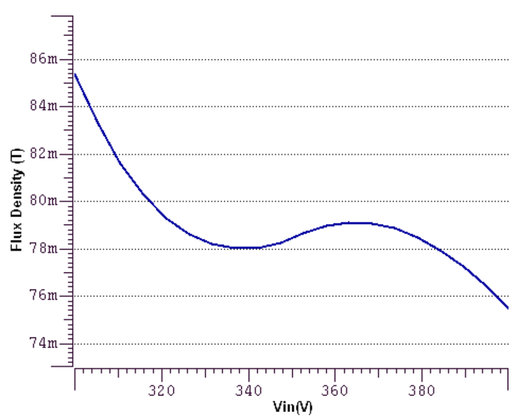


Figure 5.11 Center leg flux density for different input voltage

Figure 5.11 shows the center leg flux density for whole input voltage range. Compare with discrete design, the flux density is only half of the transformer leg and much smaller than inductor leg within all input voltage range.

The problem for this structure is the gapping. In this structure, we are using E cores. The air gap is on two outer legs while there is no air gap on center leg. This structure is not good in several aspects: first, this core structure is not a standard. The standard core normally has air gap on the center leg or no air gap at all. Second, it is not a mechanical stable structure, very accurate gap filling need to be provided. Otherwise, the accuracy of the components value will be impacted. Also, when force is applied which happens when the converter is working, the core tends to vibrate. This vibration will cause broken of the core.

A desired core structure will have air gap on center leg or same air gap for all three legs. Following part will try to establish an electrical circuit model for a general integrate magnetic structure. From the model, we can investigate new core structures.

5.1.2.3 Extraction of Common Structure for Integrated Magnetic

In the past, lot of research was done on integrated magnetic design for power converters. Review those paper, we can find that most of them are based on EE core structure or three legs structure. The difference between different designs is the placement of windings and air gaps.

In this part of the paper, the general circuit model of an EE core with four windings is used as a general structure as shown in Figure 5.12. There are air gaps on each leg. This is a very commonly used structure, many integrated magnetic design for PWM converter also used this structure with some change on the air gap or winding placement [C5].

The reason of choosing this structure for LLC resonant converter is as following:

To integrate two magnetic components, usually we need three magnetic paths. In the LLC resonant converter, although we have three magnetic components, Lm and transformer T can be build with an air-gapped transformer. So in fact we need integrated two magnetic components: series resonant inductor Lr and gapped transformer T. An EE core structure will be a reasonable choice.

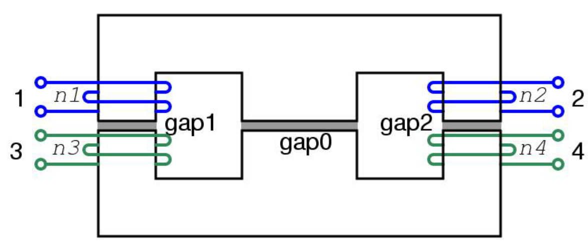
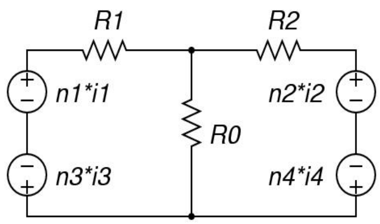


Figure 5.12 general magnetic structures for Integrated magnetic

The model is derived through duality modeling method [E4]. Through this method, we can get the electrical circuit model of a physical magnetic structure. All the components in the model are related to the physical structure of the magnetic structure. Figure 5.13 shows the reluctance model of magnetic structure shown in Figure 5.12. Figure 5.14shows electrical circuit model form this structure. In the structure, we have two sets of ideal transformer and three inductors.



$$R1 = \frac{l_{gap1}}{\mu \cdot Ac}$$

$$R2 = \frac{l_{gap2}}{\mu \cdot Ac}$$

$$R0 = \frac{l_{gap0}}{\mu \cdot Ac}$$

Figure 5.13 Reluctance model of general integrated magnetic structure

For the two ideal transformers, they have same turns ratio as in real physical structure. For the three inductors, they are correspond to each air gap and reflected to first winding n1. They can also be reflected to other windings as necessary. The value of each inductors are as following:

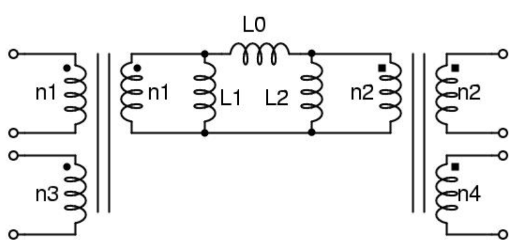


Figure 5.14 Circuit model of general integrated magnetic structure

Base on this circuit model, we will investigate more integrated magnetic structures.

5.1.2.4 Integrated magnetic design B for LLC resonant converter

As discussed in structure A, the air gapping for structure A is not easy to implement. In this part, we will investigate structure with same air gap for all three legs and same winding structure as shown in Figure 5.15.

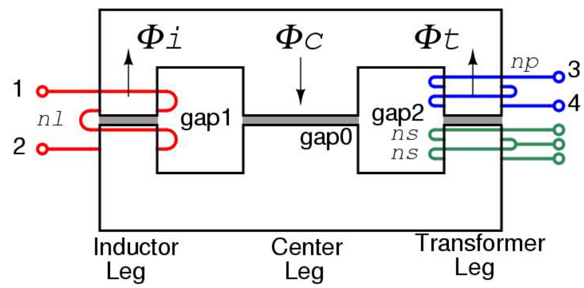


Figure 5.15 Integrated Magnetic Designs B

The electrical model of this structure can be easily got from general structure. Compare this structure with general structure; design B has only one winding on left side leg. By simplify the general model we can get following circuit model of design B as shown in Figure 5.16.

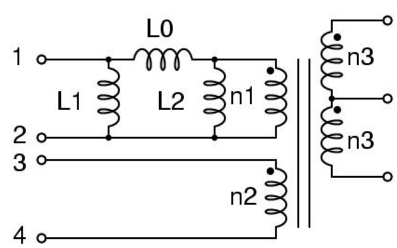


Figure 5.16 Electrical circuit model of integrated magnetic structure B

Base on the electrical circuit model of the structure, next terminal 2 and 3 are connected, which gives following circuit model.

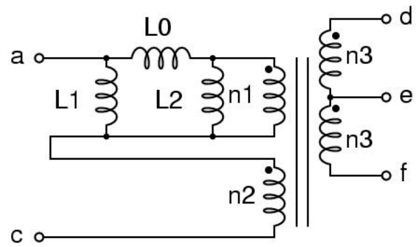


Figure 5.17 Electrical model of connecting dot-marked terminal with unmarked terminal

From circuit model in Figure 5.17, write the input current and voltage equations and solve them, then we can get the equivalent circuit of the structure. For this circuit, it has two modes. One mode is n3 is connected to output voltage. During this mode, the energy is transferred from primary to output. During the other mode, both secondary windings n3 are not connecting. We will derive the equivalent circuit for these two modes separately.

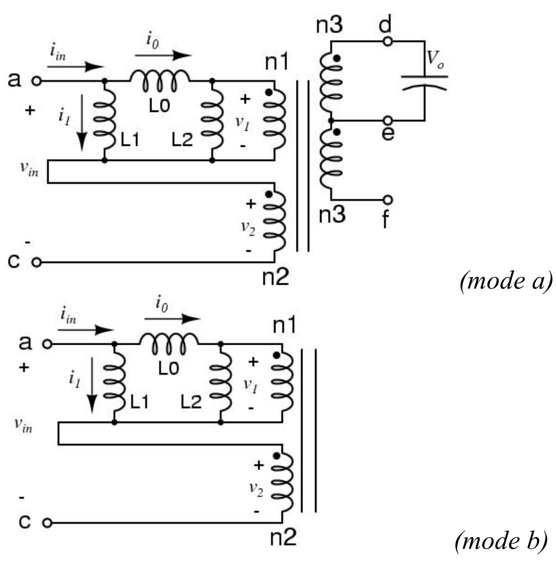


Figure 5.18 Two operation modes for LLC resonant converter

For operation mode (a), we can get following equations:

$$L1\frac{di_1}{dt}+\frac{n2}{n1}v_1=v_{in}$$

$$L0\frac{di_0}{dt}+v_1+\frac{n2}{n1}v_1=v_{in}$$

$$v_1=\frac{n1}{n3}Vo$$

$$i_0+i_1=i_{in}$$

From above equations, we can get the relationship of input voltage, input current and output voltage as following:

$$vin=\frac{L1\cdot L0}{L1+L0}\frac{di_{in}}{dt}+Vo\frac{1}{n3}(n2+n1\frac{L1}{L1+L0})$$

From this equation, we can get the equivalent circuit during this mode as in Figure 5.19.

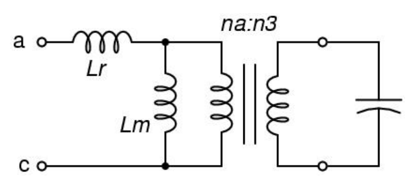


Figure 5.19 Equivalent-circuit for mode (a)

In this circuit, Lr, Lm and na are as following:

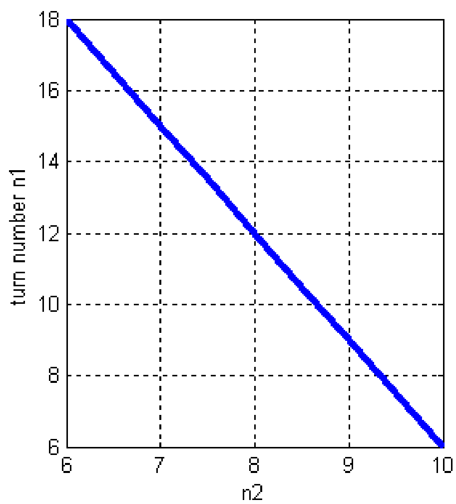
$$Lr=\frac{L1\cdot L0}{L1+L0}$$

$$na = n2 + n1\frac{L1}{L1 + L0}$$

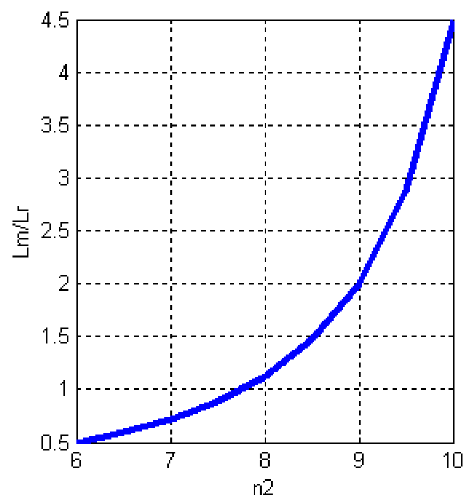
To find out Lm, we need to analyze mode (b). Same as analysis for mode (a), we can get following equations for Lm:

$$Lm = L2 \cdot \frac{na^2}{n1^2} \cdot \frac{L1 + L0}{L1 + L2 + L0}$$

From the equivalent circuit, derive the relationship between terminals; the equivalent circuit above can be simplified into the equivalent circuit, which is the structure we desired. The relationship between resonant inductor, magnetizing inductance and transformer turns ratio is shown also. Base on these equations, the structure can be designed. Following is an example of design: turns ratio 12:3, Lm=14u and Lm = 60uH. The relationship of above equations could be drawn in Figure 5.20. For given turns ratio, there are many different ways to choose n1 and n2 to get the desired na, for example, n1=n2=9, n1=6 and n2=10. The other constrain will be the desired Lm. For this case, the Lm is 4.5 times Lr. To get this Lm, the n2 need to be choosing as 10.



(a)



(b)

Figure 5.20 Design curves for integrated magnetic structure B for LLC converter

From above discussion, n1=6, n2=10 and n3=3 give us turns ratio 12:3, Lm/Lr = 4.5. Next step will be design the air gap, we knows n1 and L1 value. Follow tradition inductor design equations, the air gap can be designed. Here Lr = 14uH, from the structure it can be seen that: L1 = L3 = 0.5 L2. From the relationship above, it can be calculated that we need L1=21uH to give us equivalent Lr=14uH. With the core cross-section area and turns given, the gap can be easily derived.

In this part, the detailed information of the magnetic is described. For this converter, the core used is EE56/24/19 from Phillips. The core material is 3F3. Two outer legs are used to wind the windings. Air gap is 0.55mm for all legs.

Primary windings are built with 8 strands of AWG#27 wires. Secondary side uses 5mil X 0.9inch copper foil.

Figure 5.21 shows the simulated flux density on each of the legs. From simulation result we can see that the flux density on center leg is greatly reduced. So with this integrated magnetic structure, we can reduce the core loss greatly. Also, with this structure, the air gap is the same for all legs, which is easier to manufacture and doesn't have mechanical problem.

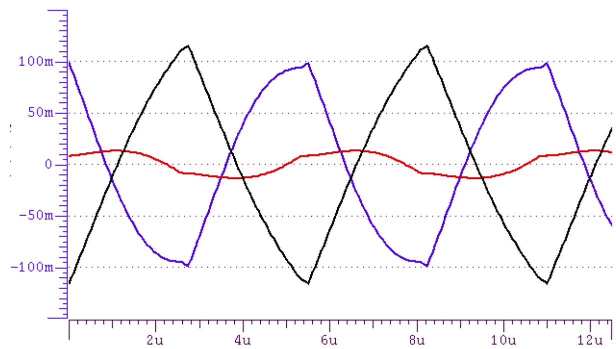


Figure 5.21 Flux density in each leg for integrated magnetic structure B

5.1.3 Test Result

In this part, the test result of integrated magnetic structure B is tested. It is compared with a discrete design. The test efficiency of integrated magnetic and discrete magnetic is shown in Figure 5.22. Because of flux ripple cancellation effect and less turns number, although the size of the magnetic components is reduced, the efficiency is almost the same for these two designs. In Figure 5.23, the sizes of these two designs were compared. With integrated magnetic, the footprint of the magnetic components could be reduced by almost 30%.

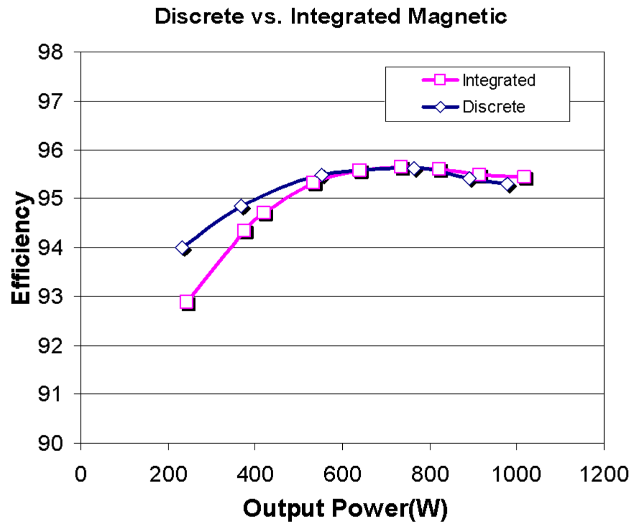


Figure 5.22 Efficiency comparison of integrated and discrete magnetic design for LLC converter

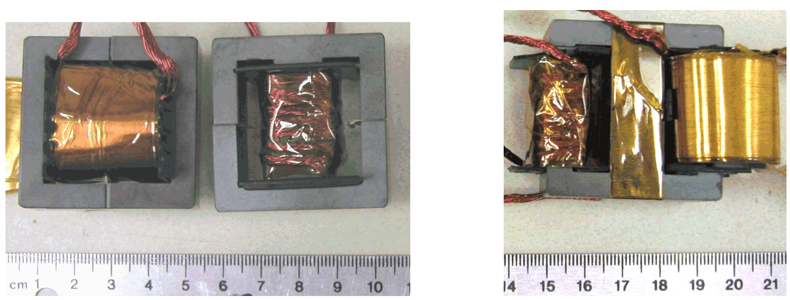


Figure 5.23 Magnetic size comparison of discrete and integrated magnetic

5.1.4 Summary

In this part, the magnetic design for LLC resonant converter is discussed. Discrete design and three method of integrated design were investigated. For discrete design, the footprint is pretty large. Also, there is no flux ripple cancellation effect; the magnetic loss is high in discrete design too. With real transformer, the magnetic components could be built with one magnetic structure. The problem is difficult to control the leakage inductance. Another integrated magnetic structure is to integrate the two U cores used to build discrete magnetic. With this method, the problem is the mechanical structure is not a stable structure. To improve this structure, a general integrated magnetic structure is developed. With the model, another integrated magnetic structure is developed with same air gap on all legs. With this magnetic structure, the manufacture is easy. There is no mechanical problem. Also, flux ripple cancellation could be achieved with this structure. Compare with discrete design, the integrated magnetic structure could provide same efficiency with 30% less footprint.

5.2 Over load protection for LLC resonant converter

In previous part of this chapter, the design of power stage was discussed. Base on these discussions, the power stage of LLC resonant converter could be designed for given specifications. Magnetic design is also investigated for LLC resonant converter. Till now we got a converter could convert 400V DC to 48V DC output with high efficiency and high power density. However, to make practical use of this converter, there are still some issues to be solved. Over load protection is one critical issue, which will be discussed in this part.

The purpose of over load protection is to limit the stress in the system during over load condition. Another function is to limit the inrush current during start up when output voltage is zero so that the power converter can be protected from destructive damage under those conditions.

In some applications, continuous operation in over load condition is required in order to achieve high system availability. In order to achieve this target, other than limit the current, healthy operation is also an important consideration, which means when the converter is running into over load protection mode, the operation of semiconductor and other components should not cause destructive damage too, i.e. lost of Zero Voltage Switching, body diode reverse recovery etc.

For traditional PWM converter, during over load condition, duty cycle is reduced to limit the current. With smaller duty cycle, the current stress could be limited.

For LLC resonant converter, it is working with variable frequency control at constant 50% duty cycle. The over load protection is totally different story. To investigate the over load protection method for LLC resonant converter, following questions need to be answered. First, the intrinsic response of LLC resonant converter to over load situation needs to be understood. Second, methods to improve the intrinsic response need to be developed if the intrinsic response is not safe or healthy.

In this part of the dissertation, first the intrinsic response of LLC resonant converter to over load condition will be investigated. Then three different over load protection methods will be discussed. First method is increasing the switching frequency. The second method, a combination of variable-frequency-control and PWM control is used to achieve over load protection. In the last method, the power stage is modified to include current limiting function into the converter. In following sections, each method will be discussed separately.

The parameters for the LLC converter used in this discussion are:

Lr=12uH, Cr=33nF and Lm=60uH, transformer turns ratio: 4:1.

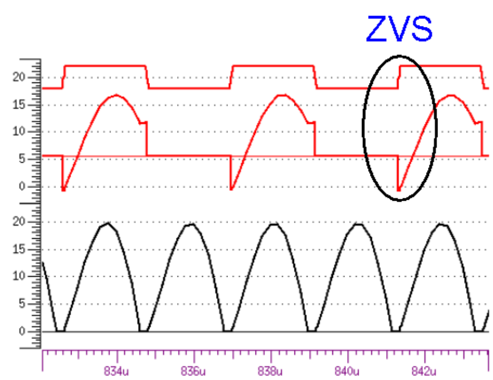
With above specs and parameters, the switching frequency range for the converter is: 170kHz to 250kHz.

5.2.1 Intrinsic response of LLC resonant converter to over load condition

In LLC resonant converter, the impedance of the resonant tank is pretty low during normal operation because it is working close to the resonant frequency of the series resonant tank. This means the current could reach very high level during over load situation. This characteristic makes over load protection design for LLC resonant converter very critical.

During over load condition, the load of the converter increases. The worst scenario will be short circuit of output. In this part, the impact of short circuit output will be investigated for LLC resonant converter.

The simulation waveforms of LLC resonant converter during normal operation and over load operation are shown in Figure 5.24. From the simulation waveforms, lost of ZVS and high current stress could be observed during over load condition. This could be understood through the characteristic of the converter. In Figure 5.25, the DC characteristic of LLC resonant converter is shown. At normal operation, the converter is working at Point A, when over load condition happens, the operating point will move to Point B. As seen in the graph, point A is in ZVS region while point B is in ZCS region. The over load current for different switching frequency is shown in Figure 5.26, the over load current could rise to very high.



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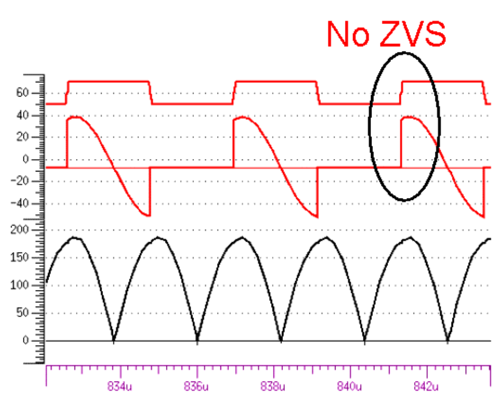


(b)

Figure 5.24 Simulation waveforms of LLC resonant converter at (a) normal operation, and (b) short circuit operation

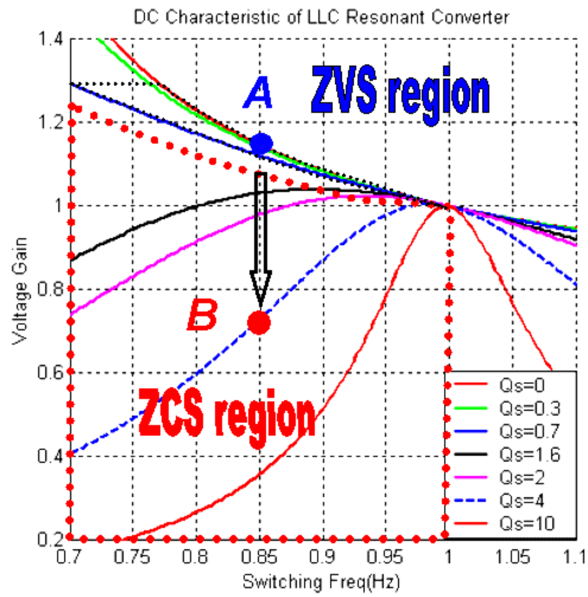


Figure 5.25 Lost of ZVS for LLC resonant converter during over load situation

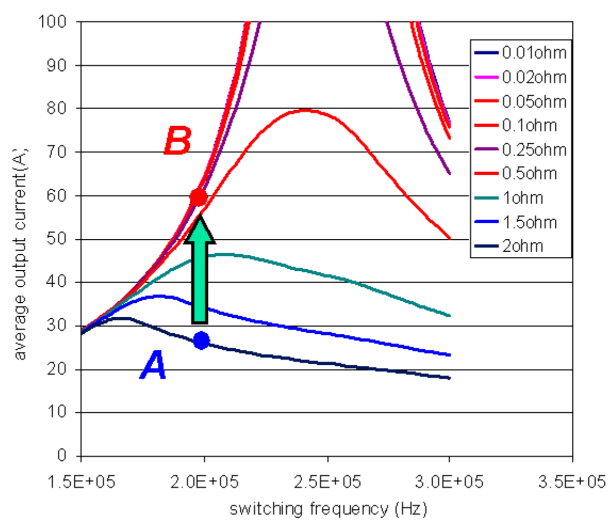


Figure 5.26 High current stress during over load situation for LLC resonant converter

From these results, the major problems for LLC resonant converter during over load condition are: high current stress, and lost of ZVS.

5.2.2 Method 1: Increasing Switching Frequency

When converter running into over load protection condition, there are two ways to limit the current. First way is to reduce the average voltage applied to the converter. For example, in PWM converter, duty cycle is reduced to limit the current. By reducing duty cycle, the average voltage applied to converter is reduced so that the current can be limited. Second way is to increase the impedance of the power stage of the converter so to limit the current. This method is useful for variable frequency controlled converters. By moving the switching frequency away from resonant frequency, the impedance of the resonant tank will increase so that the current can be limited.

To simplify the problem, let's look at the worst scenario: short circuit of output. Under such condition, the LLC resonant converter could be simplified into a simple series resonant tank as shown in Figure 5.27.

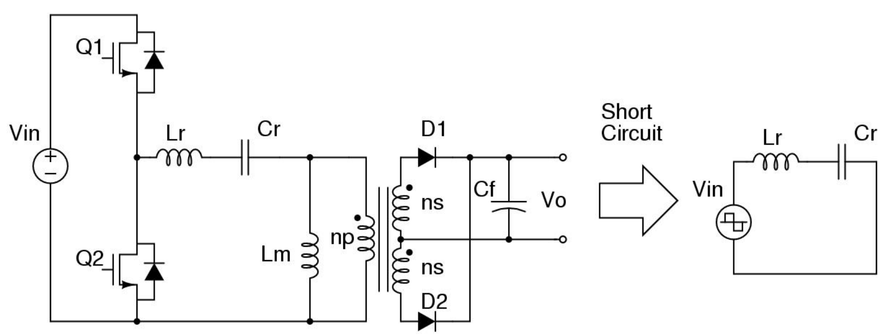


Figure 5.27 Simplified model of LLC resonant converter during short circuit condition

With this model, the switching frequency needed to limit the output current during short circuit situation could be derived. It is shown in Figure 5.28. As seen in the graph, if the desired over load current is 27A, then the switching frequency need to increase to about 400kHz.

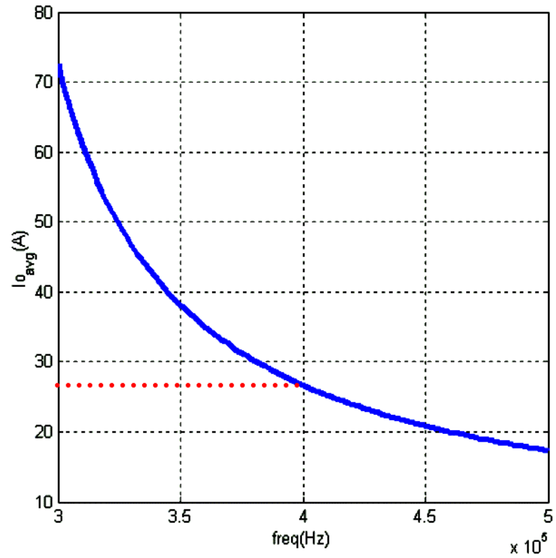


Figure 5.28 Short circuit output current at different switching frequency

Figure 5.29 shows the average for different over load condition. From Fig.2 we can see, by moving switching frequency away from resonant frequency (250kHz), the output current can be limited. There are two directions to move switching frequency: move to higher frequency or lower frequency in relationship to resonant frequency. Since the lower frequency will result in ZCS condition as shown in Figure 5.30, which is not a desirable working condition for MOSFET, here we will move switching frequency to higher frequency.

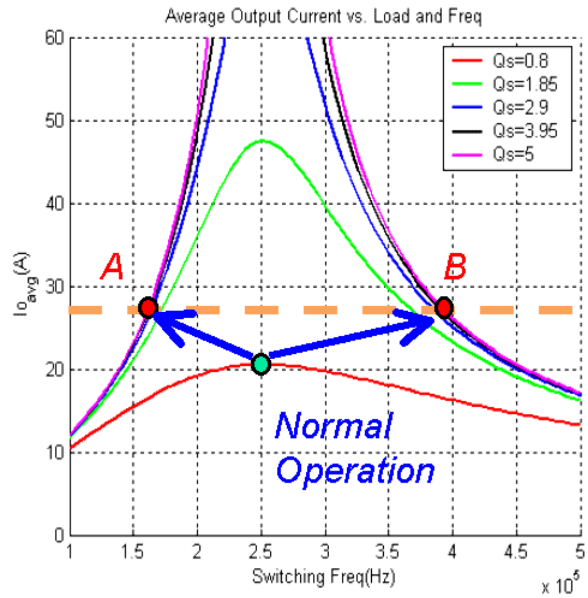


Figure 5.29 Average output current vs. switching frequency under short circuit

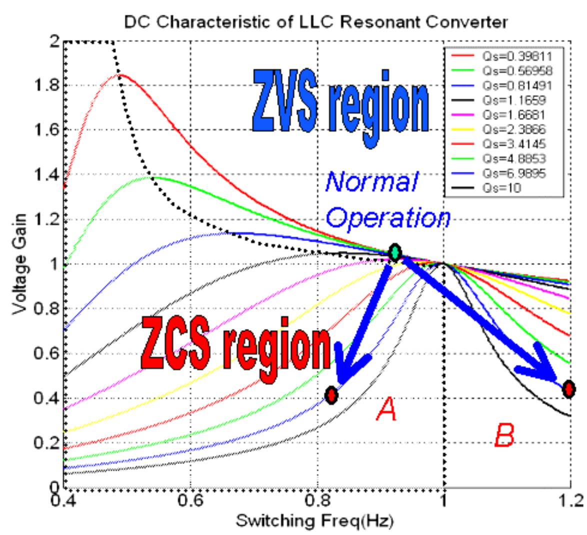


Figure 5.30 Change of operating mode with different switching frequency under protection mode

Figure 5.31 shows the test waveforms for this condition. In the real test, because of the parasitic parameters, with 358kHz switching frequency, the output current can be limited under 27A under short circuit condition.

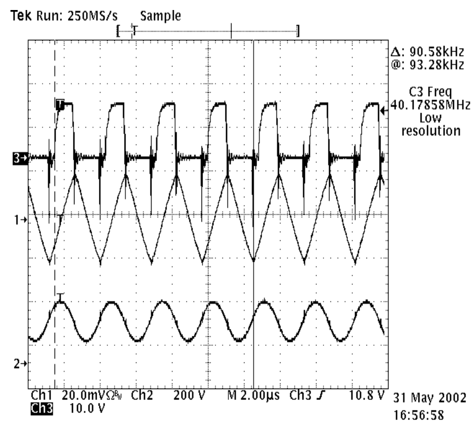


Figure 5.31 Test waveform (top to bottom: Q1 gate signal, Transformer primary current and resonant cap voltage)

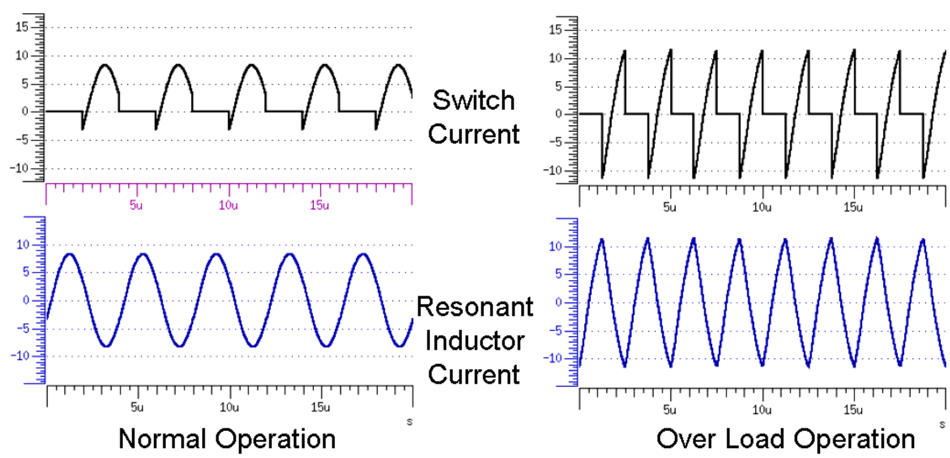


Figure 5.32 Problems with high switching frequency protection mode

For this method, the converter will be working at pretty high switching frequency during over load protection mode compare with normal operation condition. With high switching frequency, there are several considerations:

First the switching loss will increase. As shown in Figure 5.32, during short circuit condition, current stress reaches the highest. Turn off current also reaches the highest. With so high switching frequency, the loss on the device will be very high which will increase the thermal management requirement.

Second, the stress on the magnetic components will be very unbalanced. During over load protection, switching frequency reaches highest level while all the volt-second is applied to the inductor, which means inductor has to be designed according to this highest. For LLC resonant converter, this frequency will be almost double of normal operation frequency; this will make the size of the inductor to be larger.

5.2.3 Method 2: Variable frequency control plus PWM Control

From previous discussion, reduce the voltage applied to the converter can limit the current too. In the second method, variable frequency control and PWM control method are combined.

For this method, the converter has two modes: normal operation mode and protection mode. During normal operation mode, variable frequency control is used to get high efficiency. During over load protection mode, first switching frequency is increased to limit the current, when switching frequency reaches the limit we set, PWM control mode will be used to reduce the voltage applied to resonant tank as shown in Figure 5.33. With this method, the output current can be effectively limited. As shown in Figure 5.34, the current can be limited with duty cycle change. In this graph, when switching frequency is lower than 300kHz, variable frequency control is used. When switching frequency reaches 300kHz, duty cycle control cut in.

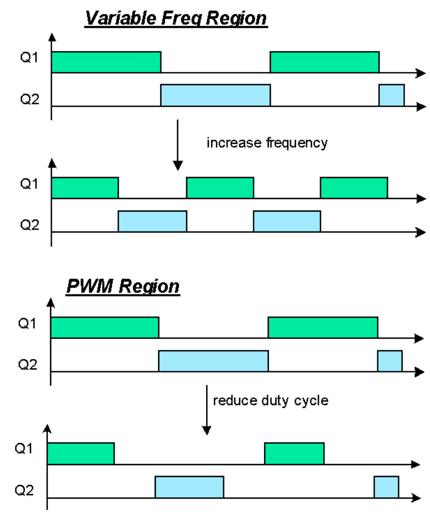


Figure 5.33 Control Method of Variable freq + PWM control

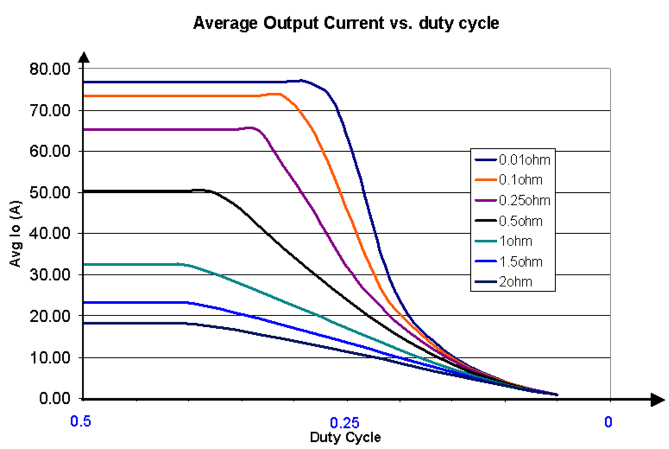


Figure 5.34 Average output current of LLC converter with variable frequency + PWM control

In Figure 5.34, a flat area is observed when duty cycle is close to 0.5. In this flat area, the duty cycle change cannot change the current. The reason is for each switching cycle, the body diode of the MOSFET will conduct for some time; duty cycle change must be larger than the body diode conduction time as shown in Figure 5.35.

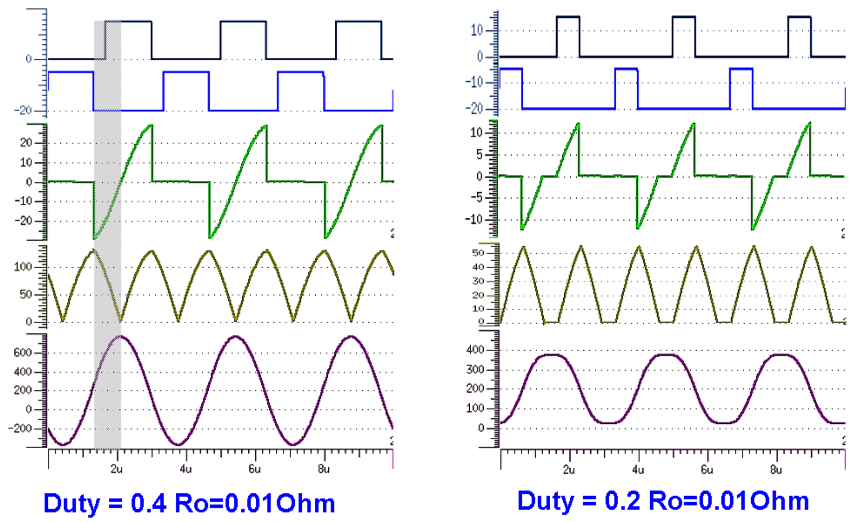


Figure 5.35 Simulation waveform for D=0.5, 0.4 and 0.2 at short circuit condition

Figure 5.36 shows the simulation result with consideration of output capacitance of MOSFET. The current will resonant instead of stay at zero. Also can be seen from Fig.10 that the ZVS condition of MOSFET is lost because of DCM operation of primary current. Figure 5.37 shows the test result.

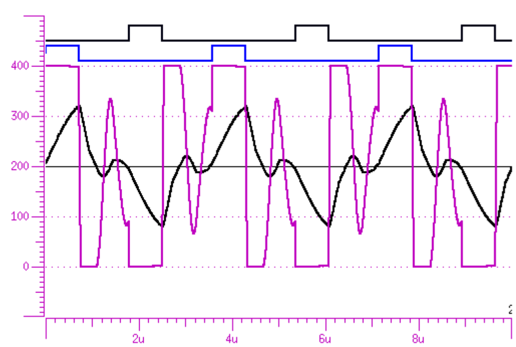


Figure 5.36 Simulation waveform with PWM control

(from top: gate signal of Q1 and Q3, Vds of Q1 and primary current Ip)

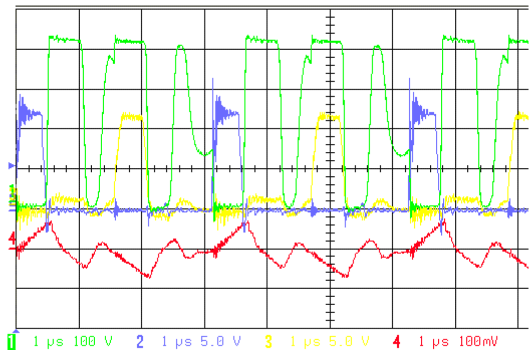


Figure 5.37 Test waveform for PWM control

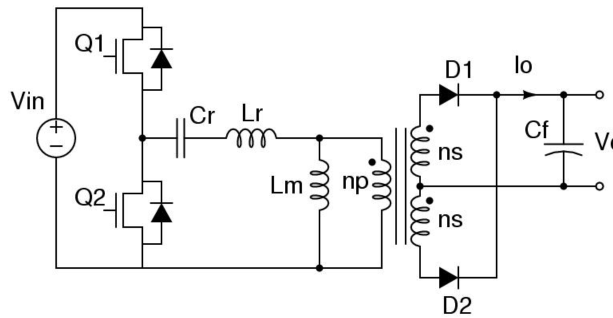
(Top: Vds of Q1, middle: gate signal of Q1, and Q2, bottom: primary current)

This method can achieve the current limiting function. The concern is operating condition. Since ZVS is lost during over load protection mode, the switching loss will increase and noise on gate driver will be a problem too. Another issue will be how fast the transition between different modes could be. Since the current could ramp up very fast, a very fast protection is necessary.

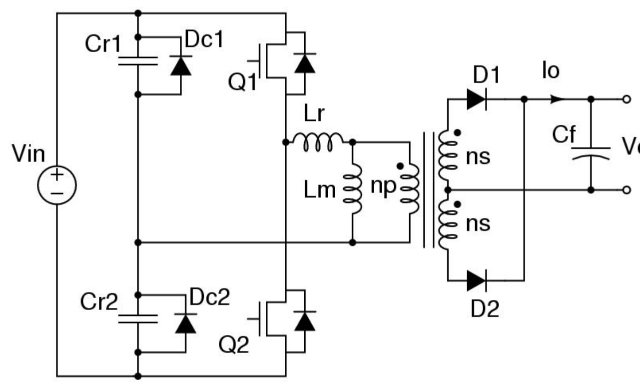
5.2.4 Method 3: LLC resonant converter with clamping diode

In this method, the current limit function is built in the power stage. This method can provide cycle-by-cycle current limiting function without control interference. Also, this method provides some other benefits too. Next the detail of this method will be discussed.

Figure 5.38 shows the original LLC resonant converter and proposed LLC resonant converter. The proposed LLC resonant converter is different from previous discussed LLC resonant converter in following aspects: first, the resonant capacitor is spited into two capacitors. Then, two diodes are put in parallel with the resonant capacitor. With this modifications, there are several benefits could be achieved.



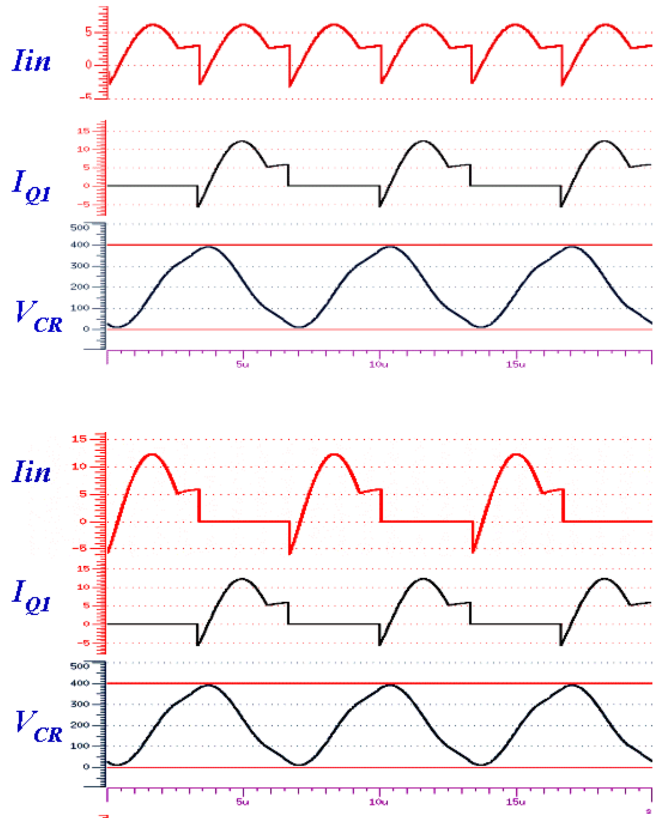
(a)



(b)

Figure 5.38 Two LLC resonant converter topologies: (a) Original LLC converter and (b) proposed clamped LLC converter

First benefit is achieved through splitting the resonant capacitor. Figure 5.39 shows the simulation waveforms of these two topologies. As seen from the simulation waveform, with splitting resonant capacitor, the input current will have lower ripple. This will alleviate the stress put on the high voltage bus capacitor.



(a)

(b)

Figure 5.39 Simulation waveforms for two LLC resonant converter topologies: (a) original LLC converter and (b) clamped LLC converter

Another benefit will be over load protection, which is provided by the clamping diodes. Figure 5.40 shows the simulation waveforms of original LLC resonant converter and the clamped LLC resonant converter at over load condition. For original LLC resonant converter, it can be seen that during over load condition, input current is very high and the peak voltage across resonant capacitor will increase to very high too. This is because during over load condition, more current is going through the resonant tank, which will charge the resonant cap to higher voltage. For LLC resonant converter with clamping diodes, first the voltage stress on resonant cap is limited so that a low voltage cap can be used; another benefit is that by limit the voltage on resonant cap, the energy could be absorbed by resonant tank is limited as shown in the state plane in Figure 5.41.

Also could be observed from the simulation waveform of clamped LLC resonant converter, under clamped operating mode, ZVS is still achieved.

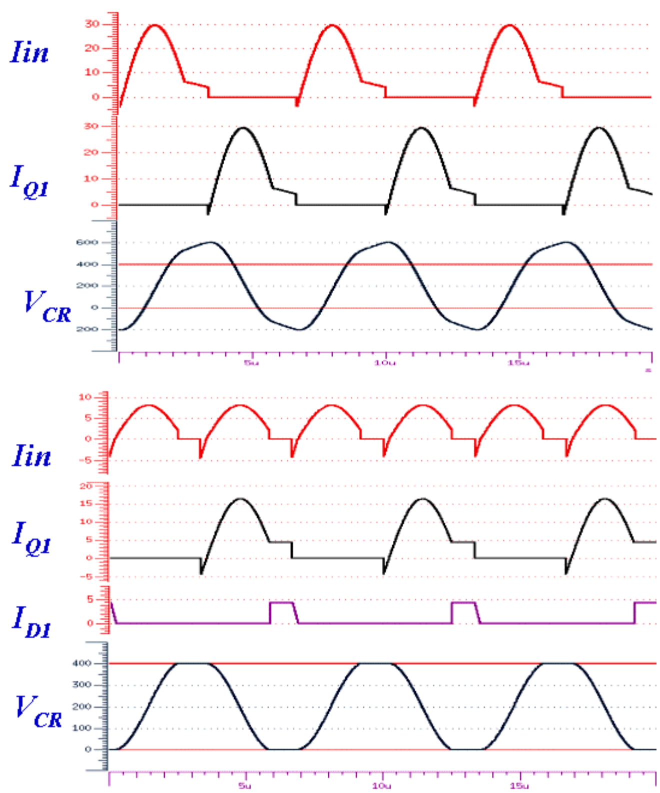


Figure 5.40 Simulation waveforms under over load condition for (a) original LLC converter and (b) clamped LLC converter

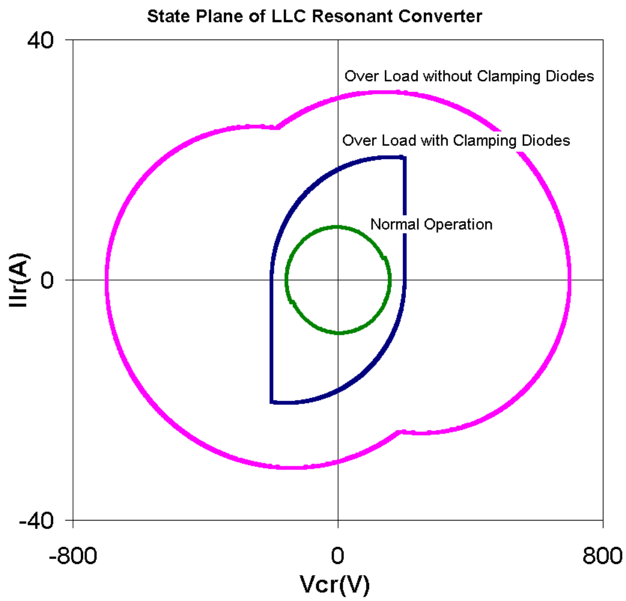


Figure 5.41 State plane of original and clamped LLC resonant converter

The over load current for both topologies are shown in Figure 5.42 and Figure 5.43.

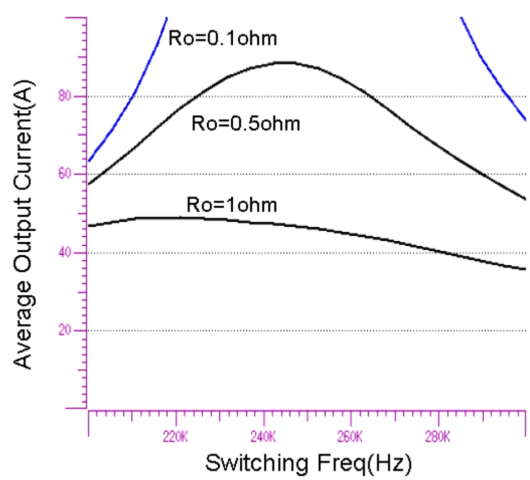


Figure 5.42 Average output current under over load condition for original LLC converter

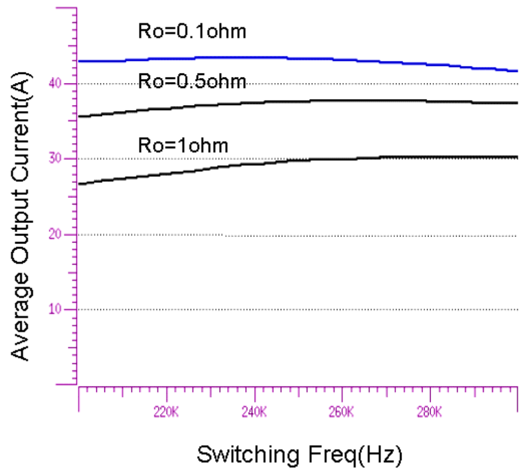


Figure 5.43 Average output current under over load condition for clamped LLC converter

Another benefit of this method is that this method doesn't need active control; it is very simple to implement. Its response speed is fast, which can provide cycle-by-cycle current protection. During normal operation, these two diodes will not conduct, the clamped LLC converter operates exactly same as original LLC resonant converter in every aspects.

In order to avoid the clamp diodes to impact normal operation condition, the design is chosen as shown in Figure 5.39. Within the expected operating region of the converter, the voltage stress on resonant capacitor is designed to be lower than the clamping voltage. Figure 5.44 shows the design region for clamped LLC resonant converter. During normal operation condition, the voltage stress on the resonant capacitor is always lower than the clamp voltage, which is the input voltage. Figure 5.45 shows the test waveforms with this method. With this method, the converter is tested with short circuit with output current at 32A at switching frequency at 200kHz for over 5 minutes.

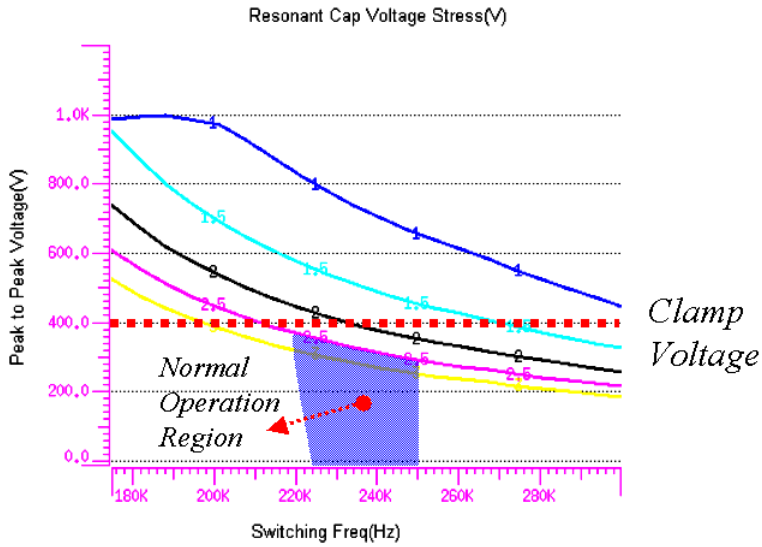


Figure 5.44 Design region of clamped LLC resonant converter

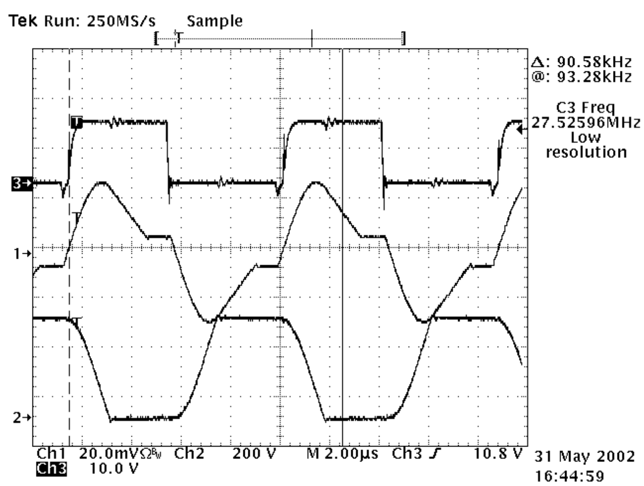


Figure 5.45 Test waveform of LLC converter under clamping mode

To use this method, there are several concerns. As described before, because of these clamping diodes, the current is limited for each switching cycle. The current can be passed through the resonant tank is related to the input voltage. Also, since this method limit the amount of current flow through resonant tank in each switching cycle, when switching frequency is changing, the average output current will change too. Let's look at an example next.

For the given application, when input voltage is 300V, we set the maxim output current at 27A. When input voltage is 400V, two things changes: input voltage is higher, switching frequency is higher too. From above analysis, this will increase the maxim output current. Instead of 25A at 300V, the maxim output current at 400V will be 34A.

Although with this drawback, the clamping diode is still an effective way to protect the converter. With these clamping diodes, ZVS is ensured at all condition. At high input voltage, although the setting point increased, still it gives us enough time to let the controller to take over and limit the current.

Base on this information, the compensator could be designed and the front end DC/DC converter is a complete system now.

5.3 Integrated power electronics module for LLC

From above analysis and test results, LLC resonant converter demonstrated significant improvements over PWM topologies. With high frequency and high efficiency, the power density of LLC resonant converter is improved by 3 times compared with asymmetrical half bridge.

As seen in chapter 3, with advanced packaging technology, the power density and performance of asymmetrical half bridge converter could be improved significantly too. In this part, integrated power electronics module for LLC resonant converter will be discussed.

For active IPEM, it is the same for both asymmetrical half bridge converter and LLC resonant converter. With smaller turn off current and loss on the active IPEM, the thermal stress on active IPEM in LLC resonant converter will be much less. This could results to reduced thermal requirement.

For the passive IPEM for LLC resonant converter, it is different from asymmetrical half bridge converter. As discussed in previous part, with splitting resonant capacitor and clamping diodes as shown in Figure 5.46, current limiting and smooth input current could be achieved. From here, the passive IPEM for LLC resonant converter could be identified. The passive IPEMs for asymmetrical half bridge and LLC resonant converter are shown in Figure 5.47.

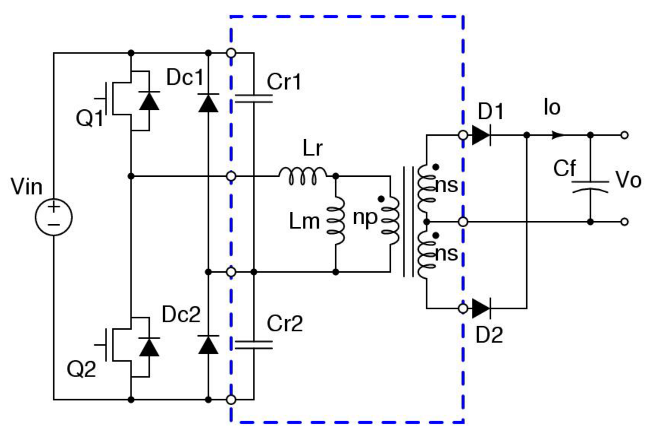


Figure 5.46 LLC resonant converter with splitting resonant cap and clamping diodes

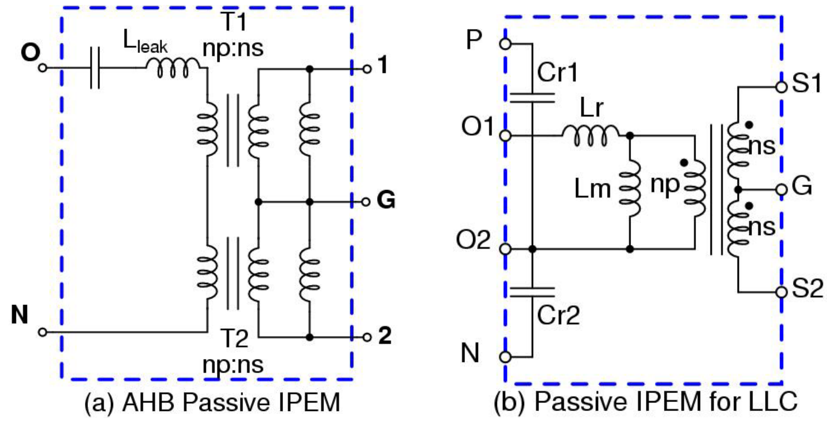


Figure 5.47 Schematics of passive IPEM for AHB and LLC r converter

Comparing these two passive IPEMs, they are very different in several ways. First, passive IPEM for AHB consists two transformers. In LLC passive IPEM, only one transformer with center-taped secondary is needed. Second, the series inductor and capacitor have very different value. For LLC resonant converter, the capacitor is around 40nF while AHB need 1uF capacitor. Third, for LLC passive IPEM, two capacitors are needed to utilize clamping LLC topology. For asymmetrical half bridge, only one capacitor is integrated. To integrate two resonant capacitors into the structure, another dielectric layer is used as shown in Figure 5.48

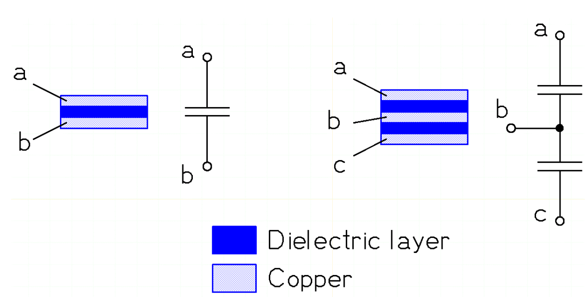


Figure 5.48 Capacitor integration for LLC resonant converter

There are different structures to build passive IPEM for LLC resonant converter. Two structures are shown in Figure 5.49.

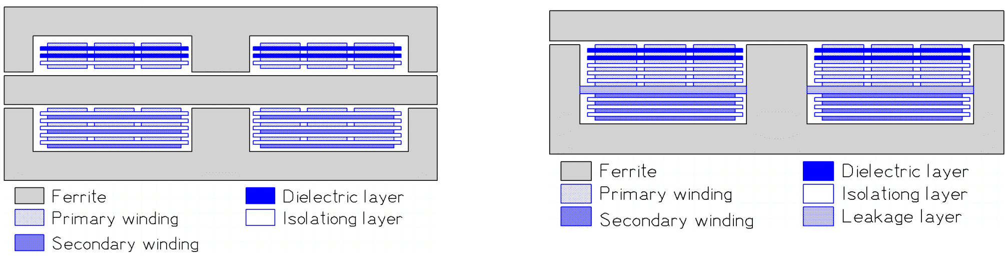


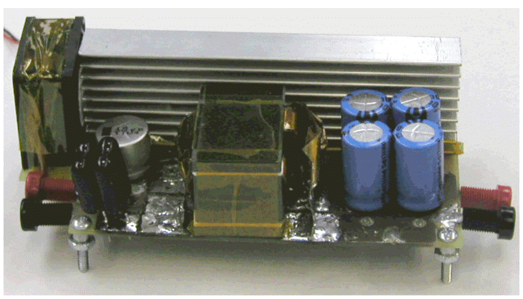
Figure 5.49 Two passive IPEM structures for LLC resonatn covnerter

For the fist method, it uses one planar E core and an I-core. It is build with transformer with controlled leakage by inserting a leakage layer between primary winding and secondary winding. The resonant capacitor could be constructed by building primary winding on dielectric material.

The other method use similar structure as asymmetrical half bridge. With two planar E core and one I core, with integrated magnetic concept, all the magnetic components could be integrated into this structure. The resonant inductor winding will be built on dielectric layer to provide resonant capacitors.

Comparing these two methods, first method is simpler. The issues of this structure are: first, accurately control the leakage inductance is not easy. For resonant converter, resonant inductance value to the operating point of the converter. The value of the inductance needs to be accurately controlled. With this method, the leakage inductance could not be very accurately controlled. Second, with this structure, interleaving of winding is impossible. For high frequency operating, this could introduce high winding loss in the structure. With method two, these problems could be solved with more complex structure.

With advanced packaging technology, all the passive components except output filter capacitor could be integrated into the planar structure. Also, the active IPEM will reduce the size for primary switches. With advanced integration, the power density of 400kHz LLC resonant could be further improved.



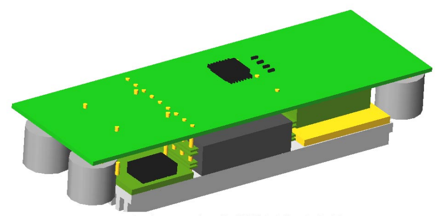


Figure 5.50 Power density of discrete LLC and projected integrated LLC

For PWM converter, the high thermal stress and requirement of snubber prevented from integrate secondary rectifier diodes into passive IPEM. With LLC resonant converter, first the thermal stress on secondary rectifier diodes is greatly reduced; also, with shottoky diodes and natural commutation, there is no need for snubber. This gives us opportunity to integrate the secondary switches into passive IPEM. With this integration, first the system will be built with just two blocks, which makes the system very simple. With this method, the parasitic between rectifier and transformer could be minimized which will be beneficial for high switching frequency operation.

5.4 Summary

In this part, the over load protection issue of LLC resonant converter is been investigated. Three ways of over load protection methods are discussed. For each method, there are some benefits and concerns.

Increasing switching frequency is simple to use and implement. The main concern is that magnetic design will be greatly affect by how high the frequency will be. Also, during protection, current stress is very high for primary switches. The thermal design for primary switch will be suffered to deal with this condition.

For variable frequency + PWM control, some modification on the control circuit is necessary to implement it. This method will prevent the issues of high frequency operation in method a. We can choose a lower frequency and use PWM control to limit the current so that magnetic and semiconductor doesn't to be over designed.

The problem of this method is that during protection mode, primary switches will loss ZVS.

LLC resonant converter with splitting cap and clamping diodes is a very effective way to limit the output current during over load condition. Basically this is a passive method to limit the current. With splitting cap, input current ripple can be reduced greatly. With clamping diodes, the current at over load condition can be automatically limited. The voltage stress on the resonant cap is also kept under a given voltage. ZVS is achieved during over load protection mode.

The problem of this method is that the setting point is a function of input, output voltage. So for different operating point, this setting value will change. It reaches minimal at low line and high output voltage and reaches maximum with high line and low output voltage.

Since each of these three methods has its pros and cons, for different requirement, different over load protection method should be choose. In some case, combination of different method could be used to get better performance too.