Chapter 4. High Frequency LLC Resonant Converter

4.1 Introduction

With the development of power conversion technology, power density becomes the major challenge for front-end AC/DC converters. Although increasing switching frequency can dramatically reduce the passive component size, its effectiveness is limited by the converter efficiency and thermal management. In present industry implementations, due to the low efficiency PWM type topologies, DC/DC stage switching frequency is limited to 100 kHz range. Therefore, passive components take large portion of converter volume. As shown in Figure 4-1, DC/DC stage transformer and inductor take more than 30% for total converter space.

Figure 4-1. 1200W front-end AC/DC converter.

Beside large passive component size, bulky holdup time capacitor becomes the bottleneck for further power density improvement. As shown in Figure 4-1, to
meet the holdup time requirement, bulky capacitors have to be used to provide the energy during the holdup time. For the 12000W design five 200uF electrolytic capacitors are used.

Holdup time requirement can be explained in Figure 4-2. When the input line exists, PFC stages converters the input AC line into a regulated DC voltage and DC/DC stages changes it into the required output voltage, such as 48V or 12V. Once input line is lost, it is required the converter to keep the regulated output voltage for more than 20mS. During the holdup time, PFC output voltage keeps decreasing and DC/DC keeps regulating the output voltage, until the minimum input voltage the DC/DC converter is reached. After that, DC/DC stage shuts down and output voltage begins to drop.

Due to the loss of input line, energy transferred to load comes from PFC stage output capacitor, which is also known as holdup time capacitor. During holdup
time, the energy that transferred to the load, as well as dissipated in the DC/DC power conversion stage, is

$$E = \frac{1}{2} C \left( V_{bus}^2 - V_{\text{min}}^2 \right)$$

For PFC stage, the output voltage is always set to 400V to ensure its capability of operates at highest input line. Therefore, to provide the required energy during holdup time, both large $C$ with large $V_{\text{min}}$ and small $C$ with small $V_{\text{min}}$ can meet the requirement. The relationship between the holdup time capacitor requirement and the minimum DC/DC stage input voltage for front-end AC/DC converter at different power levels is illustrated in Figure 4-3. It can be seen that the holdup time capacitor requirement decreases with reducing DC/DC minimum input voltage. Therefore, wide input range DC/DC converter is desired for the front-end AC/DC converters.

![Figure 4-3. Holdup time capacitor vs. minimum DC/DC stage input voltage.](image-url)
For conventional PWM converters, it is difficult to achieve both high efficiency and wide operation range simultaneously [D-1]. For instance, the asymmetrical half bridge converter, circuit is desired to operate with 50% duty cycle and achieve maximum efficiency [D-2]. However, to ensure wide operation range, during normal operation, duty cycle has to be reduced to realize regulation capability. Thus, circuit normal operation condition has to be compromised to achieve wide operation range. As demonstrated in Figure 4-4, 1 kW, 48V output asymmetrical half bridge (AHB) is able to achieve 94.5% optimal efficiency at 200 kHz switching frequency when it is dedicatedly designed for 400V input and operates with 50% duty cycle. However, to achieve wide operation range from 300V to 400V, duty cycle at 400V input has to be reduced to 30%, which dramatically reduce the converter efficiency. As demonstrated in Figure 4-4, converter efficiency reduces to 92% when the wide operation range is realized.

![Figure 4-4. Asymmetrical half bridge efficiency.](#)
To allow DC/DC stage accomplishing both wide operation range and high efficiency, different research efforts have been implemented.

By using range winding concept, transformer turns-ratio can be changed during holdup time, as shown in Figure 4-5. In this way, regulation during holdup time can be achieved by using different transformer turns-ratio, instead of compromising the duty cycle [D-3].

![Holdup extension by range winding concept.](image)

Holdup extension circuit is an alternative solution, as shown in Figure 4-6. Instead of modifying DC/DC stage topology, adding an extra stage Boost converter can dramatically reduce the duty cycle range of DC/DC stage, so that better efficiency can be achieved. At normal operation, holdup time extension circuit is bypassed by the parallel diode. During holdup time, the extension circuit begins to work and boost the voltage from holdup time capacitor and maintain DC/DC stage input voltage. Therefore, DC/DC stage sees a relatively stable voltage. Thus, DC/DC stage can be designed without considering wide operation
range. Furthermore, under normal operation condition, the extension circuit doesn't work, and all the load current goes through the bypass diode. In this way, very little loss is generated by the extension circuit and circuit is able to maintain high efficiency [D-4][D-5].

Although these solutions could results in good circuit performance in the aspects of efficiency, due to the complex circuit structure, together with the concern of stability issues during transient period, these solutions are difficult to be adopted by industry.

Instead of using auxiliary circuit, certain circuit topologies are able to achieve wide operation range without sacrificing normal operation condition efficiency. Among these topologies, LLC resonant converter becomes the most attractive solution due to its smaller switching loss, which enables it operates to very high switching frequency [D-6][D-7]. The LLC resonant converter topology is shown in Figure 4-7. By utilizing the transformer magnetizing inductor, LLC converter modifies the gain characteristic of series resonant converter (SRC).
LLC resonant converter can achieve both Buck mode and Boost mode operation. When input AC line exists, its input is generated by PFC stage. At this condition, converter operates at its resonant frequency to minimize the conduction loss and switching loss. During holdup time, its input voltage keeps decreasing, and switching frequency is reduced to realize Boost mode and keeping output voltage regulation.

Thus, LLC resonant operate at resonant frequency and achieves maximum efficiency for most of the time. While during holdup time, circuit operates far away from the resonant frequency, and circuit has lower efficiency, this condition only lasts for 20mS and will not cause thermal issue for the converter design.

The other benefit of using LLC resonant converter is its small switching loss. For PWM converters, large ZVS range is achieved by increasing switching turn-off current, which results in large turn-off loss. However in LLC resonant converter, magnetizing inductor current is used to realize ZVS and ZVS can be achieved for all the load conditions with small turnoff current. Therefore, zero
turn-on loss and small turn-off loss can be achieved. Furthermore, secondary side 
diode of LLC resonant converter turns off with low di/dt. Thus, small reverse 
recovery loss can be achieved on secondary side. Combining smaller switching 
loss on both primary and secondary side, LLC resonant converter efficiency is not 
sensitive to the switching loss and the circuit is able to achieve high switching 
frequency operation.

In this chapter, firstly, LLC resonant converter is analyzed based on the 
normal operation and holdup time operation. Secondly, LLC resonant converter 
design procedure has been developed to achieve high efficiency at normal 
operation and desirable operation range to minimize the holdup time capacitor. 
Furthermore, LLC resonant converter has been implemented using 1 MHz 
switching frequency to demonstrate the high frequency operation capability and 
verify the design process.

4.2 Operation principles of LLC Resonant Converter

Comparing with series resonant converter, LLC resonant largely reduces the 
magnetizing inductance. In this way, the magnetizing inductor can be used to 
realize soft switching for primary side switches. Furthermore, the magnetizing 
inductor participates into the resonance and modifies voltage gain characteristic. 
As shown in Figure 4-8, converter gain can be higher or lower than 1. For these
gain curves, switching frequency is normalized with the resonant frequency, which is determined by the series resonant tank and can be represented as

\[ f_0 = \frac{1}{2\pi \sqrt{L_r C_r}} \]

Due to different operation mode of LLC resonant converter, its operation principles are quite complex. According to its switching frequency, LLC operation modes can be separated into above, below and equal to the resonant frequency [D-8][D-9][D-10][D-11][D-12].

### 4.2.1 Switching frequency equal to resonant frequency

For conventional resonant converters, such as SRC and PRC, it is desired to make the converter operate at resonant frequency to maximize the circuit efficiency. However, when these circuits operate on the left hand side of resonant frequency, circuits work with zero current switching. To ensure ZVS operation,
enough design margins has to be considered during design stage. Thus, circuit would not be able to operate at optimal operation point.

For LLC resonant converter, ZVS switching can be achieved for the switching frequency either higher or lower than the resonant frequency. Thus, there is no requirement for design margin and circuit is able to operate with resonant frequency and achieve optimal efficiency.

For the circuit diagram shown in Figure 4-7, operation principle at resonant frequency can be demonstrated in Figure 4-9. The Equivalent circuit of the circuit at different time instance can be summarized in Figure 4-10.
Figure 4-10. Equivalent circuit for LLC circuit at resonant frequency (a) Equivalent circuit from $t_0$ to $t_1$, (b) Equivalent circuit from $t_1$ to $t_2$, (c) Equivalent circuit from $t_2$ to $t_3$.

Between time $t_0$ and $t_1$, switch Q1 is conducting. In this period, resonant tank current is larger than the magnetizing inductor current. According to the polarity of the transformer, secondary diode D1 is conducting. Therefore, voltage applied to the transformer magnetizing inductance is the output voltage reflected to transformer primary side. Thus, magnetizing current linearly increases. During this period, the different between the input voltage and output voltage is applied to the resonant tank, and resonant tank current is a sinusoidal waveform. At time $t_1$, resonant tank current reaches magnetizing current and Q1 turns off. Between time $t_1$ and $t_2$, the equivalent is shown as Figure 4-10 (b). Secondary diodes are off, because the resonant tank current is the same as magnetizing current and there is no current transferred to load. Due to the junction capacitors of Q1 and Q2, magnetizing current discharges the capacitors and help to achieve ZVS turn on of Q2. Between $t_1$ and $t_2$, both Q1 and Q2 are off. This period is always known as
dead-time, which is used to allow enough time to achieve ZVS, as well as prevent shoot through of two switches.

At time $t_2$, $Q_2$ turn on with zero voltage switching. Between time $t_2$ and $t_3$, the different between the resonant tank current and the magnetizing inductor current is transferred to load. After $t_3$, $Q_2$ is turn off and circuit operates into another half cycle.

At resonant frequency, LLC resonant converter is able to achieve ZVS turn on for the primary side switches. Meanwhile, the switching turn-off current is maximum transformer magnetizing inductor current. By choosing a suitable magnetizing inductor, small turn-off loss can be realized. Moreover, secondary diodes turn off with low $di/dt$, which means smaller reverse recovery loss. Therefore, at resonant frequency, optimal performance of LLC resonant converter is expected.

At resonant frequency, the series resonant tank impedance is equal to zero. Therefore, the input and output voltages are virtually connected together. Thus, the voltage gain at resonant frequency is equal to 1.

### 4.2.2 Switching frequency higher than resonant frequency

When LLC resonant converter operates with switching frequency higher than resonant frequency, the circuit operates as a SRC circuit. The equivalent circuit and key waveforms are shown in Figure 4-11 and Figure 4-12, respectively.
Figure 4-11. Equivalent circuits for LLC switching frequency higher than resonant frequency.

Figure 4-12. LLC resonant converter with switching frequency higher than resonant frequency.
Between time $t_0$ and $t_1$, switch Q1 is conducting and the circuit is transferring energy to load through diode D1. At time $t_1$, Q1 is turned off. Because the switching frequency is higher than resonant frequency, the resonant tank current is higher than the magnetizing current. Between time $t_1$ and $t_2$, both Q1 and Q2 are off, the resonant tank current is charging and discharging the junction capacitors of primary side switches. At time $t_2$, voltage on junction capacitor of switch Q2 is discharged to zero, and Q2 turns on with zero voltage switching. After $t_2$, due to switch Q2 conduction, resonant tank current decreases quickly. At time $t_3$, resonant tank current is equal to magnetizing current and diode D1 turns off. After $t_3$, diode D2 turns on and begins to transfer energy to load.

In this operation mode, ZVS switching on primary side switches can be guaranteed due to the large turn off current. However, the large turn off current generates excessive turn off loss on primary side switches. Moreover, secondary side diode turns off with large $di/dt$, as shown in Figure 4-12 between $t_2$ to $t_3$, which can cause large reverse recovery on the diodes. Furthermore, the high $di/dt$ turn off of the diode causes extra voltage stress on the diode, which makes the circuit less reliable.

4.2.3 Switching frequency lower than resonant frequency

When the switching frequency of LLC resonant converter is lower than resonant frequency, magnetizing inductor participate in the circuit operation,
which modifies the converter voltage gain characteristics. The equivalent circuit and key waveforms are shown in Figure 4-13 and Figure 4-14, respectively.

Figure 4-13. Equivalent circuit with switching frequency lower than resonant frequency.

Figure 4-14. LLC resonant converter with switching frequency lower than resonant frequency.
Between time \( t_0 \) and \( t_1 \), switching \( Q_1 \) and diode \( D_1 \) are conducting, and converter delivers energy to load. At time \( t_2 \), resonant tank current resonates back and equals to magnetizing current. After that, magnetizing inductor begins to participate in the resonant. Since resonant current is equal to magnetizing inductor current, diode \( D_1 \) is turned off. Between time \( t_2 \) and \( t_3 \), magnetizing inductor transfers its stored energy to the resonant capacitor. Therefore, in this operation mode, the converter is able to boost gain up. After \( t_3 \), \( Q_1 \) turns off, and resonant tank transfers to the body diode of \( Q_2 \). Thus \( Q_2 \) achieves ZVS turn on. And then circuit enters the other half cycle.

From previous analysis, it can be seen that LLC resonant converter is able to achieve gain larger, smaller or equal to 1. When the circuit operates at resonant frequency, the converter voltage gain is equal to one, and circuit operates optimally.

### 4.3 Performance Analysis of LLC resonant converter

#### 4.3.1 Loss analysis on the operation at resonant frequency

By choosing a suitable transformer turns-ratio, LLC resonant converter could operate with resonant frequency at normal condition and achieve high efficiency [D-13]. Since at resonant frequency, converter voltage gain is equal to 1. To allow LLC converter operating with resonant frequency at normal condition, transformer turns-ratio in turn requires meeting the equation
In this equation, $V_o$ is the desired output voltage, $V_{in}$ is the resonant tank input voltage at normal operation condition, which is equal to the bus voltage for full bridge structure and is equal to half of the bus voltage for half bridge structure.

At resonant frequency, neglecting the dead time, resonant tank current $i_r$ can be simplified as a sinusoidal wave, and the magnetizing current $i_m$ is simplified as a triangle wave. As shown in Figure 4-15.

![Resonant tank current and magnetizing current at resonant frequency.](image)

Thus the resonant tank current can be expressed as

$$i_r = I_{RMS} \sin(\omega_0 t + \phi)$$

$$\omega_0 = 2\pi f_0 = \frac{2\pi}{T}$$

In this equation $I_{RMS}$ is the resonant tank RMS current, and $\omega_0$ is angle frequency of resonant frequency. At resonant frequency, magnetizing inductor
current is a triangle waveform, because magnetizing inductor is charged and discharged by output voltage. Therefore, the magnetizing inductor current can be represented as

\[ i_m(t) = -i_{m \text{ max}} + \frac{nV_o}{L_m} (t - nT) \quad \text{when} \quad nT \leq t < \left( n + \frac{1}{2} \right) T \]

\[ i_m(t) = i_{m \text{ max}} - \frac{nV_o}{L_m} \left[ t - \left( n + \frac{1}{2} \right) T \right] \quad \text{when} \quad \left( n + \frac{1}{2} \right) T \leq t < (n+1)T \]

Here \( n \) is an integer and \( T \) is the switching cycle at resonant frequency. \( i_{m \text{ max}} \) is the peak magnetizing inductor current, which can be calculated as

\[ i_{m \text{ max}} = \frac{nV_o}{L_m} \frac{T}{4} \]

Here \( V_o \) is the output voltage, \( n \) is the transformer turns ration, \( L_m \) is the magnetizing inductance.

According to circuit property, at beginning of each switching cycle, resonant tank current is equal to magnetizing current. Therefore,

\[ \sqrt{2} I_{RMS} \sin(\phi) = -\frac{nV_o}{L_m} \frac{T}{4} \]

Meanwhile, the difference between resonant tank current and magnetizing inductor current flows through the load.

\[ \int_0^{\pi/2} \left[ \sqrt{2} I_{RMS} \sin(\omega t + \phi) + \frac{nV_o}{L_m} t - \frac{nV_o}{L_m} \right] dt = \frac{V_o}{nR_L} \frac{T}{2} \]

In this equation, \( R_L \) is the load resistance. So resonant tank RMS current can be calculated as
Since the resonant tank current continuously flows through the primary side switches, its RMS value determines the primary side conduction loss. After it is normalized with load current reflected to the primary side, the resonant tank RMS current is only related to the magnetizing inductance, the load resistance, and the switching cycle. While the switching cycle and load resistance are predetermined values according to the converter specifications, RMS current is only determined by the magnetizing inductance.

However, if considering the definition of $L_n$ and $Q$, resonant tank RMS current can be transformed into

$$ I_{RMS} = \frac{1}{2\sqrt{2} n R_L} \left( 1 \left[ \frac{1}{L_n Q} \right]^2 \right) $$

From the new equation, for the predetermined output voltage, transformer turns ratio, and load resistance, the primary side RMS current at resonant frequency is determined by the product of $L_n$ and $Q$. The calculated primary side RMS current can be normalized with load current, and used to evaluate the primary side conduction loss of the circuit designs, because the output current is the same for different designs.

$$ I_{RMS \_normal} = \frac{1}{2\sqrt{2}} \left[ 1 \left[ \frac{1}{L_n Q} \right]^2 \right] $$
The relationship between the production of $L_n$ and $Q$ is shown Figure 4-16. Primary side RMS current keeps decreasing with the increasing of $L_nQ$. However, the effectiveness of increasing $L_nQ$ is limited when it is larger than 4. Because when $L_nQ$ is 4, the normalized primary side RMS current is equal to 0.364 and when $L_nQ$ is infinite, the normalized RMS current is reduced to 0.354, only 2.7% improvement can be achieved.

![Figure 4-16. Relationship between primary side RMS current with the product of Ln and Q.](image)

Besides the primary side conduction loss, secondary side rectifier conduction loss is also a concern for LLC resonant converter, especially for the low voltage high current applications. Although, diode conduction loss mainly comes from its forward voltage drop and is proportional to the output current, if considering the equivalent resistance of the diode or using synchronous rectification, it is also desirable to minimize the secondary side RMS current.
According to the operation principles of LLC resonant converter, the diode current should be the difference between the resonant tank current and magnetizing inductor current, and its RMS value can be calculated as

\[ I_{RMS_S} = \sqrt{\frac{1}{T} \int_0^T [i_r(t) - i_m(t)]^2 dt} \]

According the equation, secondary RMS can be expressed as

\[ I_{RMS_S} = \frac{1}{4 n R_e} \sqrt{\frac{5 \pi^2 - 48 n^4 R_e^2 T^2}{12 n^2}} + 1 = \frac{1}{4} \sqrt{\frac{5 \pi^2 - 48 \left( \frac{1}{L_n Q} \right)^2}{3}} + 1 \]

In this equation, secondary RMS current is reflected to transformer primary side. Same as the primary side RMS current, secondary side RMS current is also decided entirely by magnetizing inductance. Secondary side RMS can also be normalized with load current, as shown in Figure 4-17. From this curve, it still can be seen that, when \( L_n Q \) is larger than 4, increasing it has very limited effects on RMS current reduction.

![Figure 4-17. Normalized secondary side RMS current.](image)
Through the analysis, both the primary side and secondary side RMS current is only determined by the magnetizing inductor. By increasing the magnetizing inductance, the product of $L_n$ and $Q$ is increased, and the RMS current is reduced, so is the circuit conduction loss. However, when $L_nQ$ is larger than 4, very limited effects can be observed.

The switching loss of LLC resonant converter is caused by the turn-off loss of primary side switches. Because ZVS turn-on can be achieved for primary side switches, there is no turn-on loss generated. However, primary side MOSFETs turning off is hard switching and large turn-off loss is generated. It is desired to have smaller turn-off current. The turn-off current is peak magnetizing current. By choosing a suitable magnetizing inductance, both the ZVS turn-on and small turn-off current can be achieved. Furthermore, secondary diode turn off current has small $\text{di/dt}$ because of the resonant tank, which ensures small reverse recover current and loss on the secondary diode. Low $\text{di/dt}$ turn off on the diodes also ensures low voltage stress on the secondary rectifier, which enables the circuit use low voltage rating rectifiers and further improve the circuit efficiency.

4.3.2 Performance Analysis During holdup time

During holdup time, LLC resonant converter reduces its switching frequency and boosts the gain up to compensate the reducing input voltage and regulates the output voltage. Because the switching frequency is far away from the resonant frequency, its efficiency is getting lower. However, this operation mode only lasts
20mS and low efficiency shouldn't be a concern of extra thermal stress. To maintain wide input voltage range and reduce holdup time, circuit should be able to generate enough voltage gain to maintain regulation during holdup time.

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

From the operation principle, it can be observed that voltage gain is always equal to 1 when the switching frequency is equal to resonant frequency, no matter what $L_n$ or $Q$ values are. However, for different $L_n$ and $Q$ values, the peak gains that the converter can achieve are different. Although the peak gain might be calculated based on theoretical analysis, there is no close form solution can be found. To simplify the analysis, gain characteristics at different $L_n$ and $Q$ combinations are simulated based on the simulation tool Simplis, which can automatically reach the circuit steady state within short simulation time. The simulation results are shown in Figure 4-18.

From the simulation results, it can be observed that converter voltage gain reaches its peak value at certain switching frequency that is below the resonant frequency. This frequency is determined by $L_n$ and $Q$ values. Different peak gain
can be achieved according to different $L_n$ and $Q$ combinations. Reducing $L_n$ or $Q$ value can increase peak gain value. Moreover, when $L_n$ value increases, the distance between the peak-gain frequency to the resonant frequency increases accordingly. Moreover, peak gain happens when the circuit is running at the boundary of zero current switching (ZCS) and zero voltage switching (ZVS) modes.

![Graph showing M=Vo/Vin vs fn for different Q values.](image)

- (a) $L_n=1$
- (b) $L_n=5$
For each $L_n$ and $Q$ combination, there is one corresponding peak gain that the converter can realize. Therefore, peak gains for different $L_n$ and $Q$ values can be summarized as the colored surface in the 3D map in Figure 4-19. Apparently, peak gain is affected by both $L_n$ and $Q$ values. By reducing $L$ or $Q$ value, higher peak gain can be achieved. According to this map, the valid $L_n$ and $Q$ combinations can be easily narrowed down. For instance, if the converter is required to be able to achieve gain higher than 2, a flat plane with gain equal to 2
can be used to intersect with the peak gain surface. Only the portion that is above the flat plane becomes the valid design, as shown in Figure 4-19.

![Figure 4-19. Achievable peak gain for different Ln and Q values.](image)

### 4.4 Design of LLC Resonant Converter

The LLC resonant converter design goal is to achieve minimum loss at normal operation condition together with the capability of achieve required maximum gain to ensure wide operation range. According to previous analysis, the relationships between the design parameters Ln and Q with the converter performance, especially the conduction loss at normal operation condition and the operation range, is revealed. These relationships can be used to develop an optimal design methodology of LLC resonant converter. The proposed design
methodology can be explained by the flow chart shown in Figure 4-20. The keys of successful design relies on choosing the suitable magnetizing inductor and the inductor ratio.

![Diagram of Design Procedure for LLC Resonant Converter](image)

*Figure 4-20. Design procedure for LLC resonant converter.*

In the design process, switching frequency is determined by the desired converter efficiency and power density requirement. Normally higher switching
could results in less passive converter size, but less efficiency. Therefore, trade-off is required to choosing a suitable switching frequency. Because switching frequency of LLC resonant converter is equal to resonant frequency for most of the time, this switching frequency also determines the resonant frequency of resonant tank.

4.4.1 Choosing Transformer turns-ratio

The purpose of choosing transformer turns-ratio is to achieve desired voltage gain at normal operation condition. As discussed before, LLC resonant converter could achieve maximum efficiency when its switching frequency is the resonant frequency. Therefore, transformer turns-ratio should be chosen so that, when the input voltage is 400V, the switching frequency should be resonant frequency, which means converter voltage gain is equal to one. Thus

\[ n = \frac{V_{\text{in,normal}}}{V_o} \]

In this equation, \( V_{\text{in,normal}} \) is the resonant input voltage, and it is 400V for a full bridge structure and 200V for a half bridge structure.

4.4.2 Choosing of magnetizing inductor

According to the circuit analysis during normal operation condition, the relationship between the conduction loss and circuit parameters has been revealed. The results show that the conduction loss is purely determined by the
magnetizing inductor value. Larger the inductor value results in a smaller conduction loss.

Beside the conduction loss, switching loss also needs to be minimized. From the circuit operation analysis, LLC converter is able to achieve ZVS turn-on for all the load conditions, as long as the turn-off current meets the ZVS requirement. The soft switching transient happens during the dead-time between two primary side switches gate signals. At resonant frequency, during dead-time, both the primary side switches and secondary side diodes are off. Resonant tank is series with magnetizing inductor and charging the junction capacitors of primary side switches. The equivalent circuit during dead-time can be illustrated in Figure 4-21.

![Figure 4-21. Equivalent circuit during dead-time.](image)

Primary side switch turns off with magnetizing inductor peak current. During the transient period, magnetizing inductor current almost keeps constant. To ensure ZVS, peak magnetizing inductor current should be able to discharge the junction capacitors. Thus,
Here \( V_{\text{bus}} \) is bus voltage, \( I_{m\text{.max}} \) is peak magnetizing inductor current, \( t_{\text{dead}} \) is the dead time, and \( C_{eq} \) is equivalent MOSFET output capacitor. Because of the nonlinearity of MOSFET junction capacitor, this equivalent capacitor is based on the stored charge.

Considering at resonant frequency, converter voltage gain is equal to one, together with the ZVS criteria, magnetizing inductor should fulfill the equation

\[
I_{m\text{.max}} t_{\text{dead}} \geq 2V_{\text{bus}} C_{eq}
\]

This equation is based on half bridge structure and voltage gain is equal to 1. When the circuit changes into full bridge structure, the equation changes into

\[
L_m \leq \frac{T \cdot t_{\text{dead}}}{16C_{eq}}
\]

From this equation, it can be observed that magnetizing inductor is proportional to the switching cycle and dead-time, and it is inverse proportional to the equivalent junction capacitor. Increasing dead-time or reducing MOSFET junction capacitor could result in a larger magnetizing inductor and less loss.

Based on the analysis on the RMS currents, the conduction loss increases with the increase of magnetizing inductor, which means the magnetizing inductor is required to be as large as possible. On the other hand, the soft switching requirement gives the maximum magnetizing inductor. Therefore, the optimal
design should be the magnetizing inductor that just meets the soft switching requirement.

However, according to the ZVS criteria, magnetizing inductor is determined by the switching cycle, dead-time and MOSFET junction capacitor. By increasing the dead-time, magnetizing inductor can be increased accordingly, which results in a smaller turn-off current and less turn-off loss. For certain stitching frequency and MOSFET junction capacitor, it is essential to choose a suitable dead-time to achieve the trade-off between the conduction loss and switching loss.

When the dead-time is larger, larger magnetizing inductor can be used while circuit still maintains ZVS capability. In this way, magnetizing inductor could be smaller and turn-off loss is smaller. However, due to the large dead-time, less time is used to transfer energy to the load. Thus, primary side current has to be increased to compensate the duty cycle loss caused by large dead-time.

On the other hand, when the dead-time is chosen too small, more turn-off current is required to maintain soft switching requirement. Therefore, less magnetizing inductor is needed, which results in more magnetizing current and more conduction. Meanwhile due to the large turn-off current, more switching loss is generated.

Apparently, the magnetizing inductor should be designed based on the trade off between the switching loss and conduction loss.
4.4.3 Choosing inductor ratio $L_n$

Besides high efficiency at normal operation condition, LLC resonant converter is required regulating the output voltage during holdup time by reducing its switching frequency. Input voltage range of LLC converter is desired to be larger so that the holdup time capacitor can be reduced. Once input voltage range is decided, resonant tank can be designed accordingly. As analyzed before, the minimum operation voltage of LLC resonant converter is determined by its peak voltage gain. Therefore, to achieve wide operation range of LLC resonant converter, it is essential to achieve enough voltage gain. By using simulation tool, peak gains for different $L_n$ and $Q$ combinations have been demonstrated in Figure 4-19. It can be redrawn into contour curves as in Figure 4-22, which can be used to locate valid design parameters. For instance, if the minimum voltage is required to be 250V, the peak gain of the designed converter has to be higher than $\frac{400}{250}=1.6$. Thus, all the $L_n$ and $Q$ combinations that allow the converter peak gain higher than 1.6 could be the valid design. Comparing with the contour curve, all the $L_n$ and $Q$ combinations beneath line 1.6 become the valid designs.

Apparently, there are infinite choices of solutions, to further narrow down the design parameters, magnetizing inductor is considered. The magnetizing inductor is determined by soft switching and conduction loss requirement. However, once magnetizing inductor is chosen, the relationship between $L_n$ and $Q$ has been fixed. From the definition of $L_n$ and $Q$, their product is
Once converter specification is defined and switching frequency is chosen, the product of $L_m$ and $Q$ is only determined by the magnetizing inductor. Therefore, for the designed magnetizing inductor, the product of $L_m$ and $Q$ is set.

$$L_m Q = \frac{L_m}{L_r} \sqrt{\frac{C_r}{n^2 R_L}} = \frac{L_m}{n^2 R_L \sqrt{L_r C_r}} = \frac{2 \pi f_0 L_m}{n^2 R_L}$$

For instance, for a 200 kHz switching frequency, 1kW output DC/DC converter with 100nS dead-time, if it chooses IXFH21N50 as the switching device, according to the equivalent junction capacitor, the magnetizing inductor is required to be 70uH. Thus, the product of $L_m$ and $Q$ should be
\[ L_n Q = 2.2 \]

Since the product is a constant, it can be draw in Figure 4-22 as a line, which represent the relationship between \( L_n \) and \( Q \) for a constant magnetizing inductor value. As shown in Figure 4-23, the marked line represents all the \( L_n \) and \( Q \) combinations that have the product of 2.2.

![Figure 4-23. Design example of LLC converter.](image)

Therefore, all the \( L_n \) and \( Q \) combinations along the marked line would give the same conduction loss and switching loss at normal operation condition, because of the same magnetizing inductor. But along the same line, due to different \( L_n \) and \( Q \) values, the maximum gain can be achieved are different. According to the
maximum gain requirement from the desired operation range of LLC resonant converter, only certain range on the line could be a valid design. From Figure 4-23, all the \( L_n \) and \( Q \) combinations below line 1.6 could meet the gain requirement. Combining the magnetizing inductor line, \( L_n \) has to be larger than 5.5 to meet the gain requirement.

After going through the loss analysis at normal operation and peak gain analysis during holdup time, the choices of circuit parameters have been narrowed down to a portion of constant magnetizing inductor line. Along the line, still infinite \( L_n \) and \( Q \) combinations can be chosen as valid design. Therefore, the trade off design is required to find a suitable \( L_n \) vale. The trade-off should be based on the converter parameters impacts on converter performance.

To evaluate circuit performance, normalizing method is always used for resonant circuit. The normalization method is summarized in Table 4-1.

<table>
<thead>
<tr>
<th></th>
<th>Base</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>Frequency</strong></td>
<td>( 1/(2\pi \sqrt{L_n C_p}) )</td>
</tr>
<tr>
<td><strong>Current on primary side</strong></td>
<td>( V_o / (nR_L) )</td>
</tr>
<tr>
<td><strong>Current on secondary side</strong></td>
<td>( V_o / R_L )</td>
</tr>
<tr>
<td><strong>Voltage on resonant capacitor</strong></td>
<td>( V_o )</td>
</tr>
<tr>
<td><strong>Volt-second on transformer</strong></td>
<td>( 2\pi \sqrt{L_n C_p} nV_{out} )</td>
</tr>
</tbody>
</table>
For LLC resonant converter, no matter it is during normal operation condition or holdup time, the input voltage is varying while its output voltage and current keeps the same. Therefore, it is more make sense to use output voltage and current as the normalization base. By using the proposed normalization method, LLC resonant converter can be evaluated with different $L_n$ and $Q$ value for different switching frequency and different power levels.

The primary side current, secondary side current, transformer volt-second and resonant capacitor voltage stress are evaluated with different voltage gains, which are summarized in Figure 4-24. The x-axis is the voltage gain of converter instead of switching frequency, because under the same voltage gain, different converters have the same input voltage and output voltage, which means a fair comparison.

For different $L_n$ value, it can be observed that, when the gain is equal to one, circuit operates at resonant frequency, primary side and secondary side RMS current keeps the same. Because the conduction loss at resonant frequency is mainly determined by magnetizing inductor, different $L_n$ value has no impact on the conduction loss.

However, the capacitor voltage stress reduces with increasing $L_n$ value. Because, according to the definition of $Q$, for the same magnetizing inductor, larger $L_n$ value means smaller $Q$ value. Thus, the resonant capacitor is lager. Therefore, the voltage stress on the resonant capacitor is reduced.
When the voltage gain is larger than one, which means the circuit is at holdup time, larger Ln value results in higher secondary side RMS current and lower voltage stress on the resonant capacitor. For the transformer volt-second, due to large design margin, saturation is not a concern. Besides, the holdup time is only requires 20mS, the extra loss caused by large transformer volt-second won’t cause extra thermal stress.

![Figure 4-24. Impacts of Ln.](image)

By summary, larger Ln value is desired to achieve smaller resonant capacitor voltage stress. Smaller Ln value is desired to have smaller conduction loss during holdup time. Therefore, the Ln value should be chosen based on the available resonant capacitors voltage rating and its value should be minimized to achieve higher efficiency during holdup time. Based on the chosen Ln and Q values, Lr and Cr can be calculated accordingly.
The design procedure can be summarized as: According to the converter specifications, especially the input and output voltage at normal operation condition, transformer turns-ratio can be design accordingly. By choosing certain switching frequency, the optimal dead-time can be chosen. According to the dead-time and selected MOSFET junction capacitor, magnetizing inductor can be designed. For the designed magnetizing inductor, designer is able to judge if the converter has enough gain to achieve desired operation range. If the converter peak gain is not enough, magnetizing inductor is required to be reduced to meet the gain requirement. Or the trade-off between the holdup time capacitor and converter efficiency can be performed to meet the design goal. After that, Ln value can be chosen based on resonant capacitor voltage stress, and Q can be designed accordingly.

4.5 Mega Hz LLC Resonant Converter

After deriving the design procedure for LLC resonant converter, LLC resonant converter with less conduction loss and switching loss can be achieved [D-1][D-14].

From the previous research results, it can be demonstrated that LLC resonant converter is able to achieve around 95.5% efficiency at 200 kHz switching frequency with 1kW output power. The prototype is shown in Figure 4-25.
Because of the low switching frequency, although the converter is able to achieve high efficiency, only 28W/in$^3$ power density is achieved.

To achieve higher power density, LLC converter switching frequency was pushed up to 400 kHz and still maintain high efficiency. As demonstrated in Figure 4-26, 48W/in$^3$ power density can be achieved and circuit still maintains efficiency around 94.7%. Comparing with 200 kHz switching frequency, the converter achieves much higher power density with small efficiency drop. Thus, the circuit efficiency is less sensitive to the switching frequency, which means the circuit is able to achieve much higher switching frequency but still maintains high efficiency.
Besides, previous LLC converter design was based on try and error approach, by comparing different designs and picking up a better one. And the operation range of LLC resonant converter is chosen to be from 300V to 400V. Although comparing with the conventional PWM converter, it shows some holdup time capacitor reduction capability. By using the proposed design methodology, LLC resonant converter is able to achieve even wider operation range and further reduces holdup time capacitor, without affecting the converter normal operation condition efficiency.

To demonstrate the high switching frequency operation capability of LLC resonant converter, together with the effectiveness of the proposed design methodology, a 1 MHz switching frequency 1kW LLC resonant converter with 48V output is designed.

According to the analysis on the holdup time capacitor requirement, input voltage range of LLC converter is desired to be wider to achieve smaller holdup time capacitor. However, according to the capacitor manufacturers, the capacitance can be used as holdup time capacitor is limited. The commercially available 450V electrolytic capacitor is 330uF, 270uF, 220uF, and 180uF. To ensure even distribution of ripple current, identical capacitors are used when the parallel is required. Therefore, by considering this effect, the relationship between the holdup time capacitor and the minimum DC/DC stage input voltage can be refined as shown in Figure 4-27.
The holdup time capacitor requirement changes dramatically around certain voltages. Slightly reduce the minimum input voltage could result in a much smaller holdup time capacitor. For conventional PWM converter to achieve high efficiency at normal operation condition, minimum operation voltage is normally set to around 330V. Thus, 880uF capacitors are required to meet the holdup time requirement. However, if the minimum operation voltage can be reduced to around 230V, the holdup time capacitor can be reduced to 440uF and achieve 50% size reduction. Thus, the peak gain of LLC resonant converter has to be higher than 1.8 to meet the operation range requirement.

Because the input voltage at normal operation condition is 400V and output voltage is 48V, to ensure the converter operating at resonant frequency at normal operation condition, the transformer turns-ratio is required to be the ratio between input and output voltage. Thus,
Because of the half bridge structure, half of the input voltage is used to calculate the turns-ratio.

According to the 400V input voltage and 1 kW power level, IXFH21N50F from IXYS is chosen as the switching device, because of its small input capacitor and fast recovery body diode. Due to the high switching frequency, driver circuit for the MOSFET is a big challenge. To ensure small driver loss and fast driving speed, it is desirable to have MOSFET with smaller input capacitor. Moreover, because MOSFET body diode is used to achieve ZVS turn-on, it is required to be fast recovery diode, so that it can block voltage after MOSFET turning off. For secondary side rectifiers, MBR20200CT from Onsemi was used. Although its voltage rating is 200V, it shows the better performance comparing with other 150V rating diode from other vendors.

After the selection of switching devices, the magnetizing inductor can be chosen according to the dead-time. As discussed before, the chosen of dead-time is based on the trade-off between the conduction loss and switching loss. For 1 MHz switching frequency, the relationship between the dead-time and converter conduction loss and switching loss can be represented in Figure 4-28. The switching loss is represented by primary side switch turn-off current and conduction loss is represented by primary side RMS current. It can be observed that the switching loss keeps decreasing with increasing dead-time. While for the
conduction loss, it firstly decreases with increasing dead-time. After dead-time is larger than 150nS, the conduction loss increases again. By combining the conduction loss and switching loss, it can be observed that the total loss reaches its minimum value around 120nS. Thus, 100nS dead-time is chosen as the design. According to the dead-time selection, together with the junction capacitor of IXFH21N50F, the magnetizing inductor can be chosen as 13uH.

(a) Normalized primary side RMS current

(b) Normalized primary side turn-off current
To realize operation range from 230V to 400V, according to the peak gain map of LLC resonant converter, the valid design of $L_n$ and $Q$ combinations are required to below the line that represents peak gain equal to 1.8, as demonstrated in Figure 4-29. Combining the designed magnetizing inductor, $L_n$ value can be chosen as the intersection of two curves and equal to 10. To leave enough design margin, chosen as 15. Therefore, the resonant inductor is 0.85uH. According to 1MHz resonant frequency, the resonant capacitor needs to be 30nF.
The circuit parameters have been summarized in Table 4-2 and compared with 200 kHz and 400 kHz switching frequency. It can be observed that the transformer size can be dramatically reduced by using 1MHz switching frequency. Although at 1MHz switching frequency, the resonant capacitor keeps the same value as 200 kHz switching frequency, the designed converter allows the minimum input voltage lower down to 200V and reduces the holdup time. Comparing with conventional design shown in Figure 4-1, the holdup time capacitor could be reduced from 11000uF to 440uF, and about 60% size reduction can be achieved. The prototype of 1MHz LLC resonant converter is shown in Figure 4-30. Its dimension is 5”×1.5”×1.75”. Thus, 76W/in$^3$ power density can be achieved.
Table 4-2. Design Parameter for LLC Resonant converter.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>200kHz</th>
<th>400kHz</th>
<th>1MHz</th>
</tr>
</thead>
<tbody>
<tr>
<td>Switching frequency</td>
<td>200kHz</td>
<td>400kHz</td>
<td>1MHz</td>
</tr>
<tr>
<td>Magnetizing inductor</td>
<td>70u</td>
<td>40u</td>
<td>12u</td>
</tr>
<tr>
<td>Resonant inductor</td>
<td>14u</td>
<td>7u</td>
<td>0.85u</td>
</tr>
<tr>
<td>Resonant capacitor</td>
<td>48n</td>
<td>24n</td>
<td>30n</td>
</tr>
<tr>
<td>Transformer core size</td>
<td>1 Set EE55/28/21</td>
<td>2 Sets of E43/10/28</td>
<td>1 Set E43/10/28</td>
</tr>
</tbody>
</table>

Comparing with 400 kHz prototype, 70% power density improvement has been achieved. Furthermore, by using the proposed design method, the minimum input voltage is around 200V, which reduces the holdup time capacitor to 440uF and achieves 30% size reduction comparing with 400 kHz LLC. The experimental waveforms at resonant frequency and during holdup time are shown in Figure 4-31 and Figure 4-32, respectively.

From the experimental results, at resonant frequency, soft switching can be achieved with minimum turn off current. By changing Lr and Cr combination, the
waveform keeps the same, which verifies the analysis that the loss purely relies on the magnetizing inductance. During holdup time, the converter has to reduce switching frequency to boost up voltage gain. As shown in Figure 4-32, most of the energy stored in the magnetizing inductor is transferred back to the resonant capacitor and gain can be achieved higher than 2.

![Figure 4-31. Experimental waveform at resonant frequency.](image1)

![Figure 4-32. Experimental waveform during holdup time.](image2)

The efficiency of 1MHz LLC resonant converter with different load conditions is shown in Figure 4-33. The efficiency reaches its maximum
efficiency 94.5% at about 700W output power. At full load condition, the efficiency drops down to 92.5%. Comparing with lower switching frequency cases, slightly efficiency drop can be observed.

![Figure 4-33. Efficiency comparison of LLC resonant converter.](image)

Although the developed prototype is able to achieve high efficiency with 1 MHz switching frequency, it is important to get detailed loss breakdown and estimate the further switching frequency increasing possibility. According to the circuit operation waveform, loss breakdown of the designed converter under 400V input and 1kW output is summarized in Figure 4-34. Because of the diode rectification, secondary diode conduction loss is the main loss. Moreover, considering the switching loss on primary side, it reaches 5.6W at 1MHz switching frequency, which is about 0.5% of the total power. Moreover, the loss density of the transformer reaches 700kW/m$^3$, for $\Delta B$ is equal to 50mT.
There is about 2% efficiency error between the theoretical analysis and experimental results, which is mainly caused by the loss estimation for the transformer copper loss and secondary diode switching loss. Because the planar transformer structure, skin effects and approximate effects can largely increase the conduction loss, which can be further improved by better winding structure design through 3D finite element analysis (FEA). For the diode switching loss, it is getting more and more severe when the switching frequency is higher because di/dt of turn-off current is getting higher for high frequency case. Thus the switching loss is no longer neglectable at higher switching frequency.

The major benefit of higher switching frequency for DC/DC converter is to shrink the passive component size. But its effectiveness is limited by the thermal management. For the present transformer design, to achieve low loss density, high performance 3F4 material was used, which is suitable for the switching frequency 1 to 2 MHz. Although $\Delta B$ is chosen as 50mT, $700 \text{kW/m}^3$ loss density is generated.
in the ferrite core. The relationship between 3F4 material loss density and switching frequency, peak flux density can be represented as

\[ P_h = 0.0001005 \cdot f^{2.8} \Delta B^{2.4} \]

When the switching frequency is further increased, transformer flux density has to be decreased to maintain similar loss density. Thus, the transformer size decreasing is getting less to maintain similar loss density, which means the diminishing return on the high switching frequency. Therefore, more detailed analysis is required to estimate the switching frequency impacts on the transformer size.

Furthermore, when switching frequency is getting higher, switching loss increases linearly. In the designed prototype, switching loss takes 1~2% efficiency. Apparently, further increasing switching frequency needs to consider the converter efficiency requirement.

4.6 Future work for LLC resonant converter

For LLC resonant converter, the key components are the resonant inductor, capacitor and magnetizing inductor. To achieve the performance as theoretical analysis, the desired equivalent circuit is shown in Figure 4-35.
When the switching frequency is low, parasitics of transformer, especially the leakage inductance on transformer primary and secondary side can be ignored. However, when switching frequency is pushed up, resonant components values keep decreasing. Thus transformer parasitic components can't be ignored. They begin to participate into resonance. Because the leakage inductances exist on both primary and secondary side, the equivalent circuit for resonant tank changes in to the equivalent circuit shown in Figure 4-36. Therefore, these leakage inductances should be controlled as small as possible. Further more, because of diode turn-off current has higher di/dt at high switching frequency, leakage inductances of secondary windings are desired to be well coupled together, so that the extra switching loss caused by diode reverse recovery current can be recovered.
Therefore, different transformer winding structures need to be investigated, so that these parasitic components can be controlled. Besides, because of the low output voltage and high output current interleaving winding structure also need to be considered to achieve smaller winding loss.

Furthermore, because the parasitic components couldn't be totally removed, it might be new research opportunities to find a suitable resonant tank that can absorb the transformer parasitics while maintain same gain characteristic of LLC resonant converter.

4.7 Summary

Front-end AC/DC converter is under the pressure of continuous increasing power density requirement. By using high switching frequency PFC stage, Boost inductor size and EMI filter size can be reduced to achieve higher power density. However, to further improve the power density, bulky holdup time capacitor becomes the bottleneck for further improvement.

To reduce the holdup time capacitor, wide operation range DC/DC converter is required. For conventional PWM converters, increasing operation range is achieved by sacrificing the normal operation efficiency. Although by using auxiliary holdup time extension circuit, high efficiency can be achieved together with smaller holdup time capacitor size. Due to the complex structure, it becomes less attractive comparing with LLC resonant converter.
By reducing the magnetizing inductor, LLC resonant converter modifies SRC and able to achieve gain higher or lower than 1. At normal operation condition, LLC resonant converter operates with resonant frequency and achieves maximum efficiency. When the input voltage drops, switching frequency can be reduced to boost converter gain up and maintain regulated output voltage. During holdup time, less efficiency is achieved, which is not a concern for the converter design. Furthermore, the smaller switching loss makes the converter suitable for operating with high switching frequency and maintains high efficiency.

Although smaller switching loss and the capability of wide input range operation makes LLC resonant converter quite attractive for front-end AC/DC converters, lack of design methodology makes the converter less valuable, because of its complexity. In this chapter, based on the theoretical analysis on the circuit operation at resonant frequency and maximum gain point, the relationship between the converter loss and operation range has been revealed. Based on the relationship, the converter efficiency can be optimized with a good magnetizing inductance design. Moreover, by choosing $L_n$ and $Q$ from the peak gain curves, the operation range of the LLC resonant converter can be ensured.

The developed methodology has been implemented into a 1MHz 1kW LLC resonant converter design. And the experimental results verify the theoretical analysis. By pushing the switching frequency up to 1MHz, LLC resonant converter is able to achieve 92.5% efficiency at full load and 94.5% efficiency at
700W output. 76W/in\(^3\) power density can be achieved. Comparing with 400 kHz switching frequency, 70% power density improvement is achieved. Furthermore, by using the proposed design method, the minimum input voltage can be reduced to 200V, which results in 440uF holdup time. Comparing with conventional PWM converter design with 330V minimum input voltage, 50% size reduction of holdup time capacitor can be achieved by using LLC resonant converter.