Design Considerations for an LLC Resonant Converter

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Abstract: Recently, the LLC resonant converter has drawn a lot of attention due to its advantages over the conventional series resonant converter and parallel resonant converter: narrow frequency variation over wide load and input variation and Zero Voltage Switching (ZVS) of the switches for entire load range. This paper presents an analysis and reviews practical design considerations for the LLC-type resonant converter. It includes designing the transformer and selecting the components. The step-by-step design procedure explained with a design example will help engineers design the LLC resonant converter easily.

I. INTRODUCTION

The growing demand for higher power density and low profile in power converter designs has forced designers to increase switching frequencies. Operation at higher frequencies considerably reduces the size of passive components such as transformers and filters. However, switching losses have been an obstacle to high frequency operation. In order to reduce switching losses, allowing high frequency operation, resonant switching techniques have been developed [1-7]. These techniques process power in a sinusoidal manner and the switching devices are softly commutated. Therefore, the switching losses and noise can be dramatically reduced. Conventional resonant converters use an inductor in series with a capacitor as a resonant network. Two basic configurations are possible for the load connection; series connection and parallel connections.

For the series resonant converter (SRC), the rectifier-load network is placed in series with the L-C resonant network as depicted in Fig.1 [2-4]. From this configuration, the resonant network and the load act as a voltage divider. By changing the frequency of driving voltage V_d, the impedance of the resonant network changes. The input voltage will be split between this impedance and the reflected load. Since it is a voltage divider, the DC gain of an SRC is always lower than 1. At light load condition, the impedance of the load will be very large compared to the impedance of the resonant network; all the input voltage will be imposed on the load. This makes it difficult to regulate the output at light load. Theoretically, frequency should be infinite to regulate the output at no load.

For parallel resonant converter, the rectifier-load network is placed in parallel with the resonant capacitor as depicted in Fig. 2 [5-7]. Since the load is connected in parallel with the resonant network, there inevitably exists large amount of circulating current. This makes it difficult to apply parallel resonant topologies in high power applications.

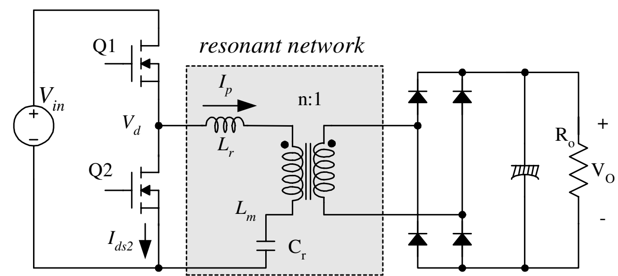

Fig.1. Half-bridge series resonant (SR) converter

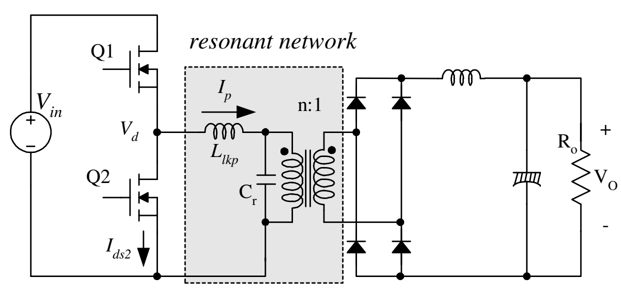

Fig.2. Half-bridge parallel resonant (PR) converter

In order to solve the limitations of the conventional resonant converters, the LLC resonant converter has been proposed [8-12]. The LLC-type resonant converter has many advantages over conventional resonant converters. First, it can regulate the output over wide line and load variations with a relatively small variation of switching frequency. Second, it can achieve zero voltage switching (ZVS) over the entire operating range. Finally, all essential parasitic elements, including junction capacitances of all semiconductor devices and the leakage inductance and magnetizing inductance of the transformer, are utilized to achieve ZVS.

This paper presents an analysis and design considerations for a half-bridge LLC resonant converter. Using the fundamental approximation, the voltage and current waveforms are analyzed and the gain equations are obtained. A design for DC/DC converter with 120W/24V output has been selected as a typical example for describing the design procedure.

II. OPERATION PRINCIPLES AND FUNDAMENTAL APPROXIMATION

Fig. 3 shows the simplified schematic of half-bridge LLC resonant converter and Fig. 4 shows its typical waveforms. In Fig. 3, L_m is the transformer magnetizing inductance and L_{lkp} and L_{lks} are the leakage inductances on the transformer primary and secondary sides respectively. Operation of the LLC resonant converter is similar to that of the conventional LC series resonant converter. The only difference is that the value of the magnetizing inductance is relatively small and therefore the resonance between L_m+L_{lkp} and C_r affects the converter operation. Since the magnetizing inductor is relatively small, there exists considerable amount of magnetizing current (I_m) as illustrated in Fig. 4.

In general, the LLC resonant topology consists of three stages as shown in Fig. 3; square wave generator, resonant network and rectifier network.

- The square wave generator produces a square wave voltage, V_d by driving switches, Q1 and Q2 with alternating 50% duty cycle for each switch. The square wave generator stage can be built as a full-bridge or half bridge type.
- The resonant network consists of a capacitor, leakage inductances and the magnetizing inductance of the transformer. The resonant network filters the higher harmonic currents. Thus, essentially only sinusoidal current is allowed to flow through the resonant network even though a square wave voltage is applied to the resonant network. The current (I_p) lags the voltage applied to the resonant network (that is, the fundamental component of the square wave voltage (V_d) applied to the half-bridge totem pole), which allows the MOSFET's to be turned on with zero voltage. As can be seen in Fig. 4, the MOSFET turns on while the current is flowing through the anti-parallel diode and the voltage across the MOSFET is zero.
- The rectifier network produces DC voltage by rectifying the AC current with rectifier diodes and capacitor. The rectifier network can be implemented as a full-wave bridge or center-tapped configuration with capacitive output filter.

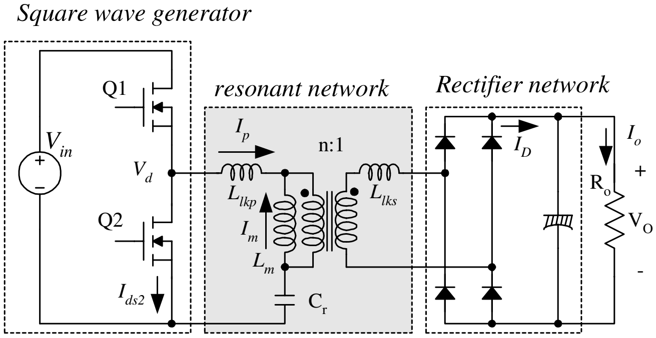

Fig. 3. A schematic of half-bridge LLC resonant converter

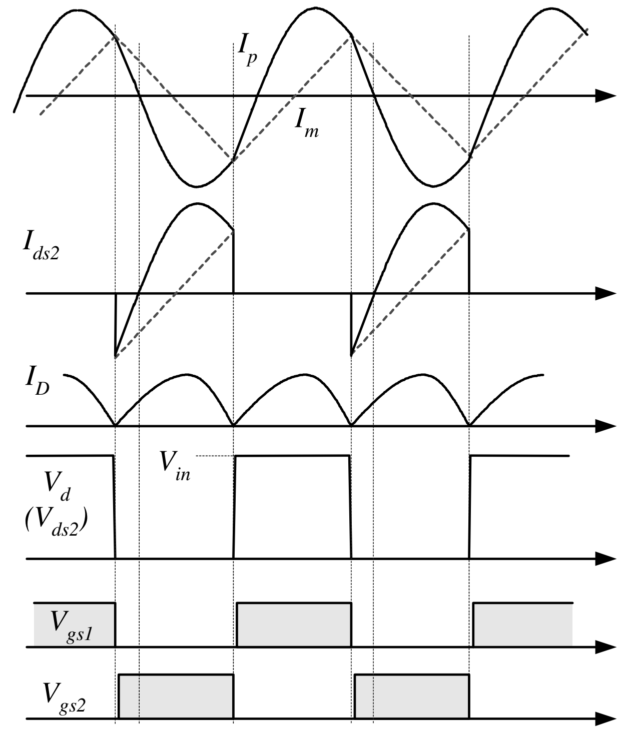

Fig. 4. Typical waveforms of half-bridge LLC resonant converter

The filtering action of the resonant network allows us to use the classical fundamental approximation to obtain the voltage gain of the resonant converter, which assumes that only the fundamental component of the square-wave voltage input to the resonant network contributes to the power transfer to the output. Because the rectifier circuit in the secondary side acts as an impedance transformer, the equivalent load resistance is different from actual load resistance. Fig. 5 shows how this equivalent load resistance is derived. The primary side circuit is replaced by a sinusoidal current source, I_{ac} and a square wave of voltage, V_{RI} appears at the input to the rectifier. Since the average of I_{ac} is the output current, I_o, I_{ac} is obtained as

$$I_{ac} = \frac{\pi \cdot I_o}{2} \sin(\omega t) \tag{1}$$

And V_{RI} is given as

$$\begin{aligned} V_{RI} &= +V_o \quad \text{if } \sin(\omega t) > 0 \\ V_{RI} &= -V_o \quad \text{if } \sin(\omega t) < 0 \end{aligned} \tag{2}$$

where V_o is the output voltage.

Then, the fundamental component of V_{RI} is given as

$$V_{RI}^F = \frac{4V_o}{\pi} \sin(\omega t) \tag{3}$$

Since harmonic components of V_{RI} are not involved in the power transfer, AC equivalent load resistance can be calculated by dividing V_{RI}^F by I_{ac} as

$$R_{ac} = \frac{V_{RI}^F}{I_{ac}} = \frac{8}{\pi^2}\frac{V_o}{I_o} = \frac{8}{\pi^2} R_o \tag{4}$$

Considering the transformer turns ration ($n=N_p/N_s$), the equivalent load resistance shown in the primary side is obtained as

$$R_{ac} = \frac{8n^2}{\pi^2} R_o \tag{5}$$

By using the equivalent load resistance, the AC equivalent circuit is obtained as illustrated in Fig. 6, where V_d^F and V_{RO}^F are the fundamental components of the driving voltage, V_d and reflected output voltage, V_{RO} (nV_{RI}), respectively.

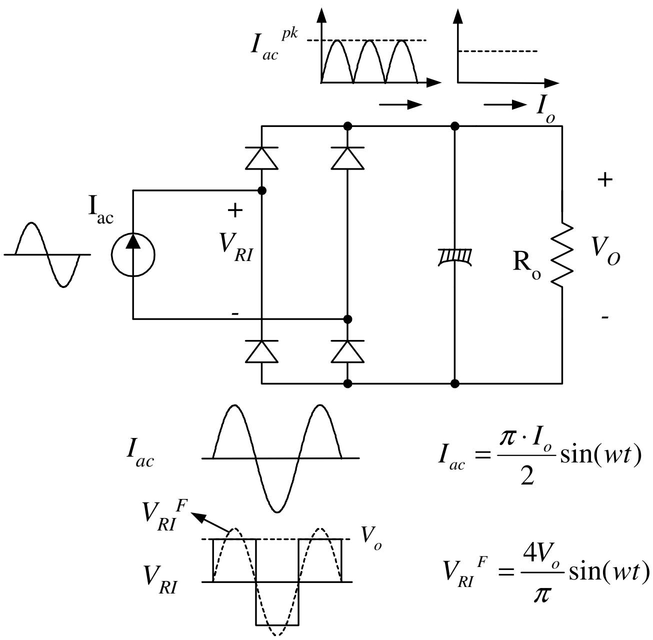

Fig. 5 Derivation of equivalent Load resistance R_{ac}

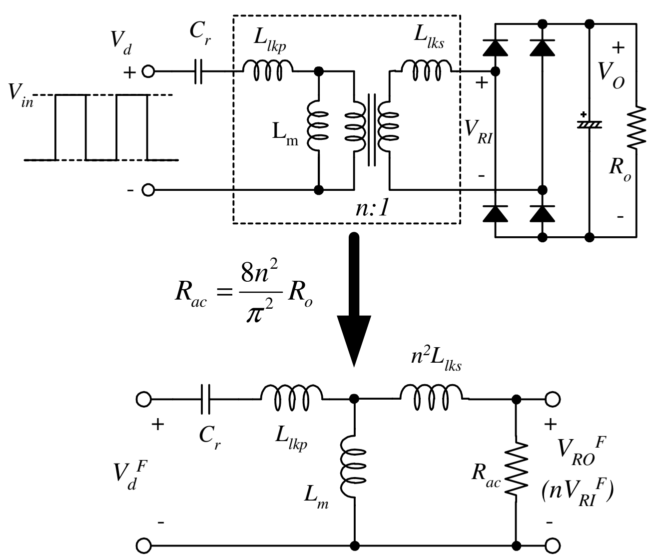

Fig. 6 AC equivalent circuit for LLC resonant converter

With the equivalent load resistance obtained in (5), the characteristics of the LLC resonant converter can be derived. Using the AC equivalent circuit of Fig. 6, the voltage gain, M is obtained as

$$M = \frac{V_{RO}^F}{V_d^F} = \frac{n \cdot V_{RI}^F}{V_d^F} = \frac{\frac{4n \cdot V_o}{\pi}\sin(\omega t)}{\frac{4}{\pi}\frac{V_{in}}{2}\sin(\omega t)} = \frac{2n \cdot V_o}{V_{in}}$$

$$= \left| \frac{\omega^2 L_m R_{ac} C_r}{j\omega \cdot (1-\frac{\omega^2}{\omega_o^2}) \cdot (L_m + n^2 L_{lks}) + R_{ac}(1-\frac{\omega^2}{\omega_p^2})} \right| \tag{6}$$

where

$$R_{ac} = \frac{8n^2}{\pi^2} R_o$$

$$\omega_o = \frac{1}{\sqrt{L_r C_r}}, \quad \omega_p = \frac{1}{\sqrt{L_p C_r}}$$

$$L_p = L_m + L_{lkp}, \quad L_r = L_{lkp} + L_m // (n^2 L_{lks})$$

As can be seen in (6), there are two resonant frequencies. One is determined by L_r and C_r while the other is determined by L_p and C_r. In actual transformer, L_p and L_r can be measured in the primary side with the secondary side winding open circuited and short circuited, respectively.

One important point that should be observed in (6) is that the gain is fixed at resonant frequency (ω_o) regardless of the load variation, which is given as

$$M_{@\,\omega=\omega_o} = \frac{L_m}{L_p - L_r} = \frac{L_m + n^2 L_{lks}}{L_m} \tag{7}$$

Without considering the leakage inductance in the transformer secondary side, the gain in (7) becomes unity. In previous research, the leakage inductance in the transformer secondary side was ignored to simplify the gain equation [8-12]. However, as observed, there exists considerable error when ignoring the leakage inductance in the transformer secondary side, which generally results in an incorrect design.

By assuming that $L_{lkp}=n^2L_{lks}$, the gain in (6) can be simplified as

$$M = \frac{2n \cdot V_O}{V_{in}} = \left| \frac{(\frac{\omega^2}{\omega_p^2})\frac{k}{k+1}}{j(\frac{\omega}{\omega_o}) \cdot (1-\frac{\omega^2}{\omega_o^2}) \cdot Q \frac{(k+1)^2}{2k+1} + (1-\frac{\omega^2}{\omega_p^2})} \right| \tag{8}$$

where

$$k = \frac{L_m}{L_{lkp}} \tag{9}$$

$$Q = \frac{\sqrt{L_r / C_r}}{R_{ac}} \tag{10}$$

The gain at the resonant frequency (ω_o) of (7) can be also simplified in terms of k as

$$M_{@\,\omega=\omega_o} = \frac{L_m + n^2 L_{lks}}{L_m} = \frac{L_m + L_{lkp}}{L_m} = \frac{k+1}{k} \tag{11}$$

While the gain is expressed in terms of k in (8), a gain expressed with L_p and L_r is preferred when handling an actual transformer since these values can be easily measured with a transformer. Expressing L_p and L_r in terms of k, we can obtain

$$L_p = L_m + L_{lkp} = (k+1)L_{lkp} \tag{12}$$

$$L_r = L_{lkp} + L_m \,//\, L_{lkp} = L_{lkp}\left(1 + \frac{k}{k+1}\right) \tag{13}$$

Using (12) and (13), (8) becomes

$$M = \frac{2n \cdot V_O}{V_{in}} = \left| \frac{\left(\frac{\omega^2}{\omega_p^{\,2}}\right)\sqrt{\frac{L_p - L_r}{L_p}}}{j\left(\frac{\omega}{\omega_o}\right)\cdot\left(1 - \frac{\omega^2}{\omega_o^{\,2}}\right)\cdot Q \frac{L_p}{L_r} + \left(1 - \frac{\omega^2}{\omega_p^{\,2}}\right)} \right| \tag{14}$$

(11) can be also expressed in terms of L_p and L_r as

$$M_{@\,\omega=\omega_o} = \frac{k+1}{k} = \sqrt{\frac{L_p}{L_p - L_r}} \tag{15}$$

By using the gain at the resonant frequency of (15) as a virtual gain of the transformer, the AC equivalent circuit of LLC resonant converter of Fig. 6 can be simplified in terms of L_p and L_r as shown in Fig. 7.

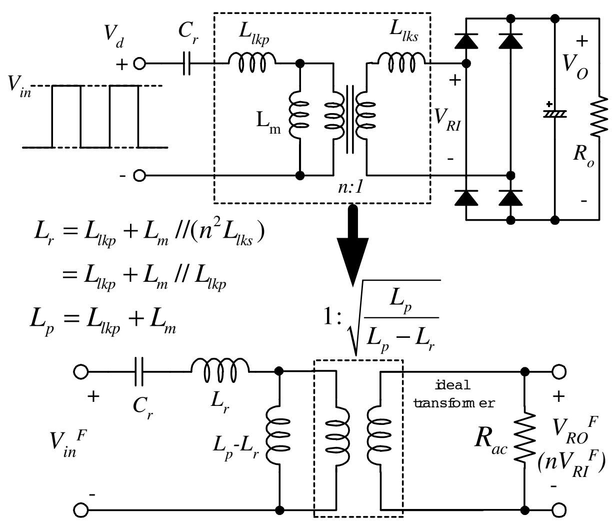

Fig. 7 Simplified AC equivalent circuit for LLC resonant converter

The gain of (8) is plotted in Fig. 8 for different Q values with k=5, f_o=100kHz and f_p=55kHz. As observed in Fig. 8, the LLC resonant converter shows characteristics which are almost independent of the load when the switching frequency is around the resonant frequency, f_o. This is a distinct advantage of LLC-type resonant converter over the conventional series-resonant converter. Therefore, it is natural to operate the converter around the resonant frequency to minimize the switching frequency variation at light load conditions.

The operating range of the LLC resonant converter is limited by the peak gain (attainable maximum gain), which is indicated with '*' in Fig. 8. It should be noticed that the peak voltage gain does not occur at f_o nor f_p. The peak gain frequency where the peak gain is obtained exists between f_p and f_o as shown in Fig. 8. As Q decreases (as load decreases), the peak gain frequency moves to f_p and higher peak gain is obtained. Meanwhile, as Q increases (as load increases), the peak gain frequency moves to f_o and the peak gain drops. Thus, the full load condition should be the worst case for the resonant network design.

Another important factor that determines the peak gain is the ratio between L_m and L_{lkp} which is defined as k in (9). Even though the peak gain at a given condition can be obtained by using the gain in (8), it is difficult to express the peak gain in explicit form. Moreover, the gain obtained from (8) has some error at frequencies below the resonant frequency (f_o) due to the fundamental approximation. In order to simplify the analysis and design, the peak gains are obtained using simulation tool and depicted in Fig. 9, which shows how the peak gain (attainable maximum gain) varies with Q for different k values. It appears that higher peak gain can be obtained by reducing k or Q values. With a given resonant frequency (f_o) and Q value, decreasing k means reducing the magnetizing inductance, which results in increased circulating current. Accordingly, there is a trade-off between the available gain range and conduction loss.

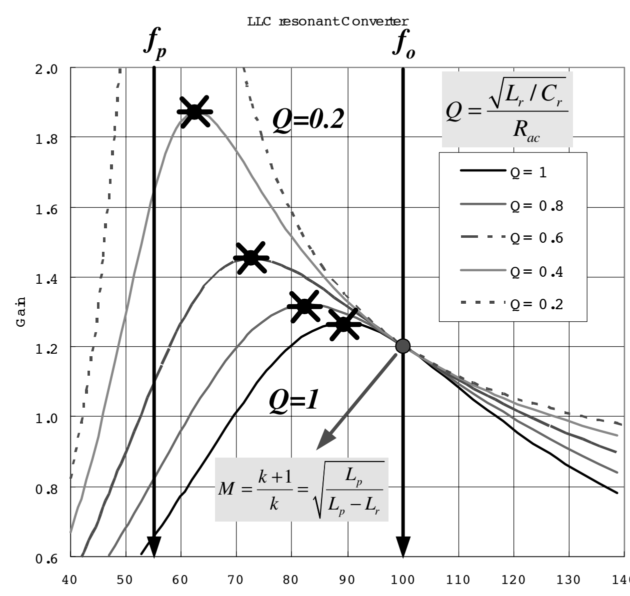

Fig. 8 Typical gain curves of LLC resonant converter (k=5 and f_o=100kHz)

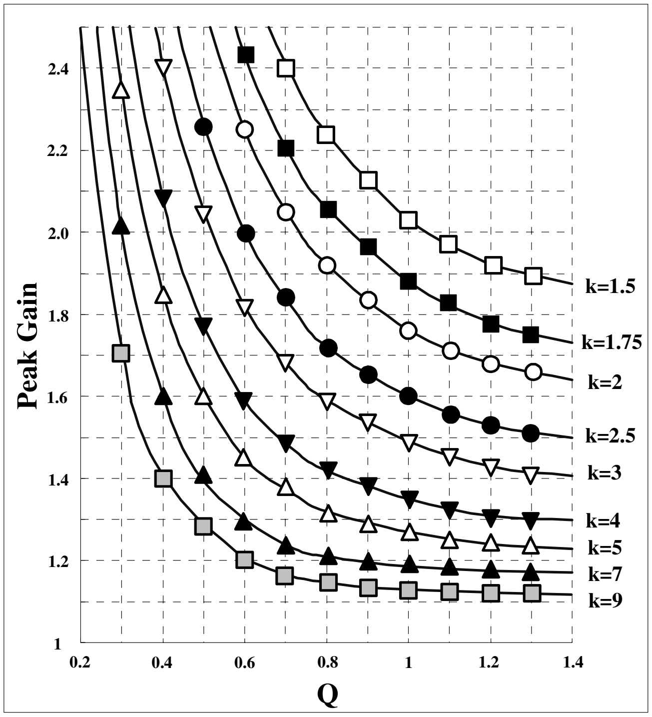

Fig. 9 peak gain (attainable maximum gain) versus Q for different k values

III. DESIGN PROCEDURE

In this section, a design procedure is presented using the schematic of Fig.10 as a reference. A dc/dc converter with 125W/24V output has been selected as a design example. The design specifications are as follows:

- Input voltage: 380Vdc (output of PFC stage)
- Output: 24V/5A (120W)
- Holdup time requirement: 17ms
- DC link capacitor of PFC output: 100uF

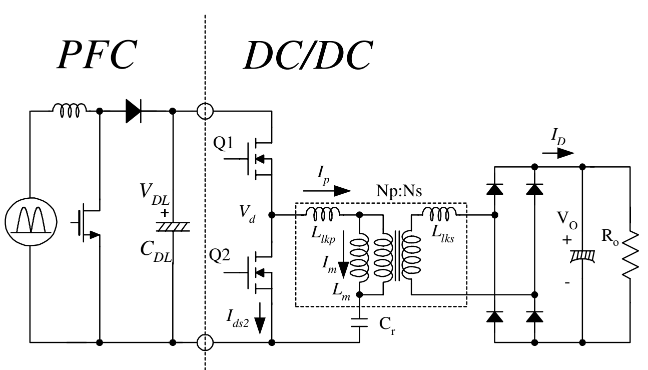

Fig.10 Schematic of half-bridge LLC resonant converter with power factor pre-regulator

[STEP-1] Define the system specifications

As a first step, the following specification should be defined.

-Estimated efficiency (E_{ff}): The power conversion efficiency must be estimated to calculate the maximum input power with a given maximum output power. If no reference data is available, use E_{ff} = 0.88~0.92 for low voltage output applications and E_{ff} = 0.92~0.96 for high voltage output applications. With the estimated efficiency, the maximum input power is given as

$$P_{in} = \frac{P_o}{E_{ff}} \tag{16}$$

-Input voltage range (V_{in}^{min} and V_{in}^{max}) : Typically, it is assumed that the input voltage is provided from Power Factor Correction (PFC) pre-regulator output. When the input voltage is supplied from PFC output, the minimum input voltage considering the hold-up time requirement is given as

$$V_{in}^{min} = \sqrt{V_{O.PFC}^{\ 2} - \frac{2P_{in}T_{HU}}{C_{DL}}} \tag{17}$$

where $V_{O.PFC}$ is the nominal PFC output voltage, T_{HU} is a hold up time and C_{DL} is the DC link bulk capacitor.
The maximum input voltage is given as

$$V_{in}^{max} = V_{O.PFC} \tag{18}$$

(Design Example) Assuming the efficiency is 95%,

$$P_{in} = \frac{P_o}{E_{ff}} = \frac{120}{0.95} = 126W$$

$$V_{in}^{min} = \sqrt{V_{O.PFC}^{\ 2} - \frac{2P_{in}T_{HU}}{C_{DL}}}$$

$$= \sqrt{380^2 - \frac{2 \cdot 126 \cdot 17 \times 10^{-3}}{100 \times 10^{-6}}} = 319V$$

$$V_{in}^{max} = V_{O.PFC} = 380V$$

[STEP-2] Determine the maximum and minimum voltage gains of the resonant network

As discussed in the previous section, it is typical to operate the LLC resonant converter around the resonant frequency (f_o) in normal operation to minimize switching frequency variation. When the input voltage is supplied from the PFC output, the input voltage has the maximum value (nominal PFC output voltage) in normal operation. Designing the converter to operate at f_o for the maximum input voltage condition, the minimum gain should occur at the resonant frequency (f_o). As observed in (11), the gain at f_o is a function of the ratio ($k=L_m/L_{lkp}$) between the magnetizing inductance and primary side leakage inductance. Thus, the value of k should be chosen to obtain the minimum gain. While a higher peak gain can be obtained with a small k value, too small k value results in poor coupling of the transformer and deteriorates the efficiency. It is typical to set k to be 5~10, which results in a gain of 1.1~1.2 at the resonant frequency (f_o).

With the chosen k value, the minimum voltage gain for maximum input voltage (V_{in}^{max}) is obtained as

$$M^{\min} = \frac{V_{RO}}{\frac{V_{in}^{\max}}{2}} = \frac{L_m + n^2 L_{lks}}{L_m} = \frac{L_m + L_{lkp}}{L_m} = \frac{k+1}{k} \quad (19)$$

Then, the maximum voltage gain is given as

$$M^{\max} = \frac{V_{in}^{\max}}{V_{in}^{\min}} M^{\min} \quad (20)$$

(Design Example) The ratio (k) between L_m and L_{lkp} is chosen as 7, which results in the minimum and maximum gains as

$$M^{\min} = \frac{V_{RO}}{V_{in}^{\max}/2} = \frac{k+1}{k} = \frac{7+1}{7} = 1.14$$

$$M^{\max} = \frac{V_{in}^{\max}}{V_{in}^{\min}} M^{\min} = \frac{380}{319} \cdot 1.14 = 1.36$$

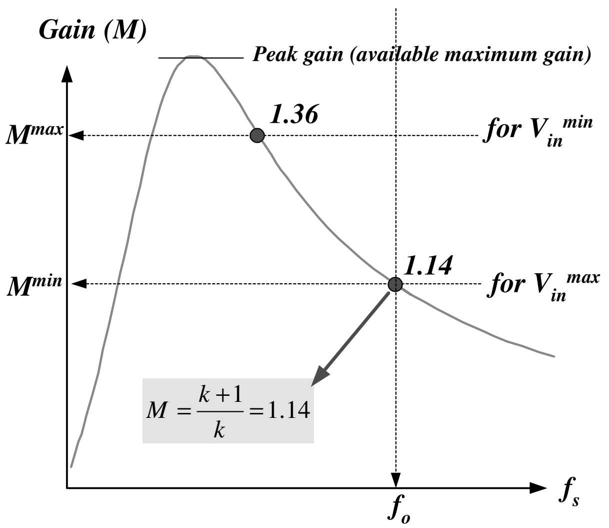

Fig. 11 Maximum gain and minimum gain

[STEP-3] Determine the transformer turns ratio ($n=N_p/N_s$)

Since the full-wave bridge rectifier is used for the rectifier network, the transformer turns ratio is given as

$$n = \frac{N_p}{N_s} = \frac{V_{in}^{\max}}{2(V_o + 2V_F)} \cdot M^{\min} \quad (21)$$

where V_F is the secondary side rectifier diode voltage drop.

(Design Example)

$$n = \frac{N_p}{N_s} = \frac{V_{in}^{\max}}{2(V_o + 2V_F)} \cdot M_{\min} = \frac{380}{2(24 + 2 \cdot 0.6)} \cdot 1.14 = 8.6$$

[STEP-4] Calculate the equivalent load resistance (R_{ac})

With the transformer turns ratio obtained from (21), the equivalent load resistance is obtained as

$$R_{ac} = \frac{8n^2}{\pi^2} \frac{V_o^2}{P_o} E_{ff} \quad (22)$$

(Design Example)

$$R_{ac} = \frac{8n^2}{\pi^2} \frac{V_o^2}{P_o} = \frac{8 \cdot 8.6^2 \cdot 24^2}{\pi^2 \cdot 120} = 288\Omega$$

[STEP-5] Design the resonant network

With k chosen in STEP-2, read proper Q value from the peak gain curves in Fig. 9 that results in enough peak gain. 10~15% margin on the peak gain is typical.
Then, the resonant parameters are obtained as

$$C_r = \frac{1}{2\pi Q \cdot f_o \cdot R_{ac}} \quad (23)$$

$$L_r = \frac{1}{(2\pi f_o)^2 C_r} \quad (24)$$

$$L_p = \frac{(k+1)^2}{(2k+1)} L_r \quad (25)$$

(Design Example)
As calculated in STEP-2, the maximum voltage gain (M^{max}) for the minimum input voltage (V_{in}^{min}) is 1.36. With 10% margin, a peak gain of 1.5 is required. k has been chosen as 7 in STEP-2 and Q is obtained as 0.43 from the peak gain curves in Fig. 12. By selecting the resonant frequency as 85kHz, the resonant components are determined as

$$C_r = \frac{1}{2\pi Q \cdot f_o \cdot R_{ac}} = \frac{1}{2\pi \cdot 0.43 \cdot 85\times10^3 \cdot 288} = 15nF$$

$$L_r = \frac{1}{(2\pi f_o)^2 C_r} = \frac{1}{(2\pi \cdot 85\times10^3)^2 \cdot 15\times10^{-9}} = 234uH$$

$$L_p = \frac{(k+1)^2}{(2k+1)} L_r = 998uH$$

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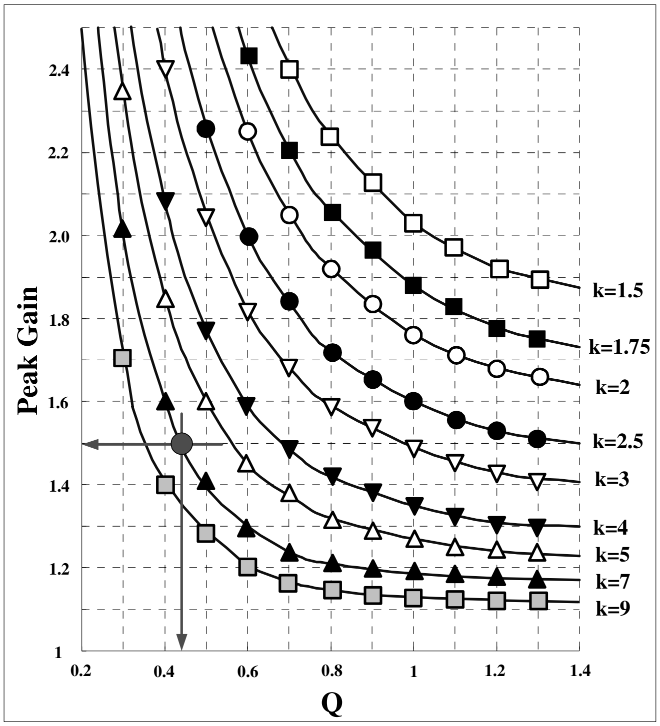

Fig. 12 Resonant network design using the peak gain (attainable maximum gain) curve for k=7

[STEP-6] Design the transformer

The worst case for the transformer design is the minimum switching frequency condition, which occurs at the minimum input voltage and full load condition. To obtain the minimum switching frequency, plot the gain curve using the gain equation of (8) and read the minimum switching frequency. Then, the minimum number of turns for the transformer primary side is obtained as

$$N_p^{\min} = \frac{n(V_o + 2V_F)}{2f_s^{\min} \cdot \Delta B \cdot A_e} \tag{26}$$

where A_e is the cross-sectional area of the transformer core in m^2 and ΔB is the maximum flux density swing in Tesla. If there is no reference data, use ΔB =0.25~0.30 T.

Then, choose the proper number of turns for the secondary side that results in primary side turns larger than $N_p^{\min}$ as

$$N_p = n \cdot N_s > N_p^{\min} \tag{27}$$

(Design Example) EER3541 core (A_e=107mm^2) is selected for the transformer. From the gain curve of Fig .13, the minimum switching frequency is obtained as 66kHz. Then, the minimum primary side turns of the transformer is given as

$$N_p^{\min} = \frac{n(V_o + 2V_F)\times 10^6}{2f_s^{\min}\Delta B \cdot A_e}$$

$$= \frac{8.6\times 25.2}{2\cdot 66\times 10^3 \cdot 0.3\cdot 107\times 10^{-6}} = 51.1 \textit{ turns}$$

$$\therefore N_p = n\cdot N_s = 8.6\times 6 = 51.6 > N_p^{\min}$$

Choosing N_s as 6 turns, N_p is given as

$$N_p = n\cdot N_s = 8.6\times 6 = 51.6 \Rightarrow 52 > N_p^{\min}$$

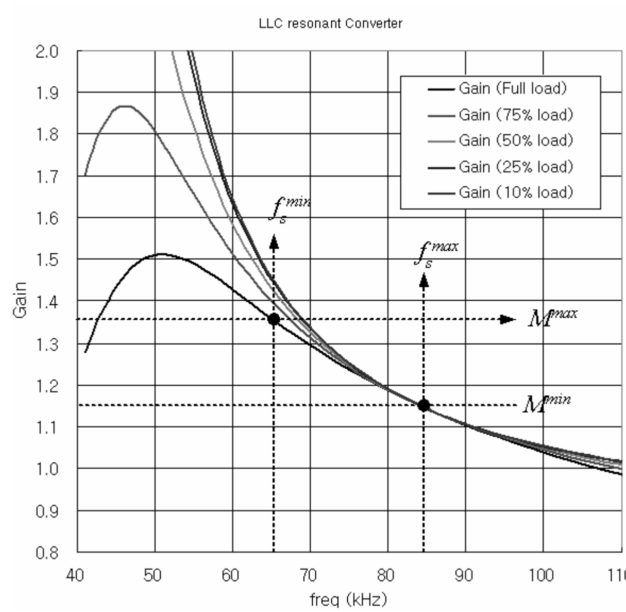

Fig. 13 Gain curve

[STEP-7] Transformer Construction

Parameters L_p and L_r of the transformer were determined in STEP-5. L_p and L_r can be measured in the primary side with the secondary side winding open circuited and short circuited, respectively. Since LLC converter design requires a relatively large L_r, a sectional bobbin is typically used as shown in Figure 14 to obtain the desired L_r value. For a sectional bobbin, the number of turns and winding configuration are the major factors determining the value of L_r, while the gap length of the core does not affect L_r much. Whereas, L_p can be easily controlled by adjusting the gap length. Table 1 shows measured L_p and L_r values with different gap lengths. With a gap length of 0.15mm, the desired L_p and L_r values are obtained.

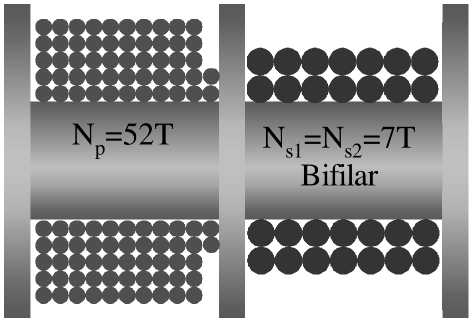

Fig. 14 Sectional bobbin

Table. 1 Measured Lp and Lr with different gap lengths

Gap length	L_p	L_r
0.0 mm	5,669 μH	237 μH
0.05 mm	2,105 μH	235 μH
0.10 mm	1,401 μH	233 μH
0.15 mm	**1,065 μH**	**230 μH**
0.20 mm	890 μH	225 μH
0.25 mm	788 μH	224 μH
0.30 mm	665 μH	223 μH
0.35 mm	623 μH	222 μH

Even though the integrated transformer approach in LLC resonant converter design can implement the magnetic components in a single core and save one magnetic component, the value of L_r is not easy to control in real transformer design. Thus, the resonant network design sometimes requires iteration with an actual L_r value after the transformer is actually built. Or, an additional resonant inductor can be added in series with the resonance capacitor to obtain the desired L_r value.

[STEP-8] Select the resonant capacitor

When choosing the resonant capacitor, the current rating should be considered since a considerable amount of current flows through the capacitor. The RMS current through the resonant capacitor is given as

$$I_{C_r}^{RMS} \cong \sqrt{[\frac{\pi I_o}{2\sqrt{2}n}]^2 + [\frac{n(V_o + 2\cdot V_F)}{4\sqrt{2} f_o L_m}]^2} \tag{28}$$

Then, the maximum voltage of the resonant capacitor in normal operation is given as

$$V_{C_r}^{\max} \cong \frac{V_{in}^{\max}}{2} + \frac{\sqrt{2}\cdot I_{Cr}^{RMS}}{2\cdot\pi\cdot f_o \cdot C_r} \tag{29}$$

(Design Example)

$$I_{C_r}^{RMS} \cong \sqrt{[\frac{\pi I_o}{2\sqrt{2}n}]^2 + [\frac{n(V_o + 2\cdot V_F)}{4\sqrt{2} f_o L_m}]^2}$$
$$= \sqrt{[\frac{\pi\cdot 5}{2\sqrt{2}\cdot 8.6}]^2 + [\frac{8.6\cdot(24+1.2)}{4\sqrt{2}\cdot 873\times10^{-6}\cdot 85\times10^{3}}]^2}$$
$$= 0.87A$$

$$V_{C_r}^{\max} \cong \frac{V_{in}^{\max}}{2} + \frac{\sqrt{2}\cdot I_{Cr}^{RMS}}{2\cdot\pi\cdot f_o \cdot C_r}$$
$$= \frac{380}{2} + \frac{\sqrt{2}\cdot 0.916}{2\cdot\pi\cdot 85\times10^{3}\cdot 15\times10^{-9}} = 343V$$

IV. CONCLUSION

This paper has presented the design of an LLC resonant converter utilizing the leakage inductance and magnetizing inductance of transformer as resonant components. The leakage inductance in the transformer secondary side was also considered in the gain equation.

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