JPE 7-1-2

Design Consideration of Half-Bridge LLC Resonant Converter

Hang-Seok Choi†

† Fairchild Korea Semiconductor, Korea

ABSTRACT

LLC resonant converters display many advantages over the conventional LC series resonant converter such as narrow frequency variation over wide range of load and input variation and zero voltage switching even under no load conditions. This paper presents analysis and design consideration for the half bridge LLC resonant converter. Using the fundamental approximation, the gain equation is obtained, where the leakage inductance in the transformer secondary side is also considered. Based on the gain equation, the practical design procedure is investigated to optimize the resonant network for a given input/output specifications. The design procedure is verified through an experimental prototype of the 115W half-bridge LLC resonant converter.

Keywords: Half-bridge resonant converter, zero voltage switching (ZVS)

1. Introduction

The Conventional PWM technique processes power by controlling the duty cycle and interrupting the power flow. All the switching devices are hard-switched with abrupt changes of currents and voltages, which results in severe switching losses and noises. Meanwhile, the resonant technique process power in a sinusoidal form and the switching devices are softly commutated. Therefore, the switching losses and noises can be dramatically reduced. For this reason, resonant converters have drawn a lot of attentions in various applications [1-3]. Among many resonant converters, the half-bridge LLC-type resonant converter has been the most popular topology for many applications since this topology has many advantages over other topologies; it can regulate the output over wide line and load variations with a relatively small variation of switching frequency, it can achieve zero voltage switching (ZVS) over the entire operating range, and all essential parasitic elements, including junction capacitances of all semi-conductor devices and the leakage inductance of the transformer, are utilized to achieve soft-switching.

While much research has been done on the LLC resonant converter topology ever since its introduction in the 1990s [4], most of the research has been focused on the steady state analysis rather than practical design consideration. In [5], the low noise features were mainly investigated and no design procedure was studied. Reference [6] discussed the above resonance operation in buck mode only, where LLC topology was introduced as "LCL type series resonant converter." In [7], detailed analysis was done on the buck mode operation as well as on the boost mode operation, but no design consideration was given. References [8-12] investigated LLC topology in detail using fundamental approximation, but simplified the AC equivalent circuit by ignoring the leakage inductance

Manuscript received July. 18, 2006; revised Sep. 21, 2006.
† Corresponding Author: hschoi@fairchildsemi.co.kr
Tel: +82-32-680-1383, Fax: +82-32-680-1317
Power Conversion Team, Fairchild Korea Semiconductor

in the secondary side. In general, the magnetic components of the LLC resonant converter are implemented with one core by utilizing the leakage inductance as the resonant inductor. Consequently, the leakage inductance exists not only in the primary side but also in the secondary side. Since the leakage inductance in the secondary side affect the gain equation, ignoring the leakage inductance in the secondary side results in an incorrect design.

This paper presents design consideration for the half bridge LLC resonant converter. Using the fundamental approximation, the gain equation is obtained, where the leakage inductance in the transformer secondary side is also considered. Based on the gain equation, the practical design procedure is investigated to optimize the resonant network for given input/output specifications. The design procedure is verified through an experimental prototype converter of the 115W half-bridge LLC resonant converter.

2. Operation Principle and Fundamental Approximation

The primary side stage of LLC resonant converter can be built as a full-bridge or half-bridge type and the output stage can be implemented as a full-bridge or center tapped rectifier configuration with capacitive output filter. Fig. 1 shows the half-bridge implementation of the LLC resonant converter with full-bridge output rectifier, where L_m is the magnetizing inductance and L_{lkp} and L_{lks} are the leakage inductances in the primary and secondary, respectively. Fig. 2 shows the typical waveforms of the LLC resonant converter. 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. Thus the resonance between L_m+L_{lkp} and C_r affect the converter operation. Since the magnetizing inductor is relatively small, there exists considerable magnetizing current (I_m) as shown in Fig. 2.

The half-bridge totem pole composed of Q_1 and Q_2 applies a square wave voltage (V_d) to the resonant network. Since the resonant network has the effect of filtering the higher harmonic voltages, essentially, a sinusoidal current appears in the resonant network. 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. The current is lagging the voltage applied to the resonant network (that is, the fundamental component of the square wave applied by the half-bridge totem pole), which enables the MOSFETs to be turned on with zero voltage.

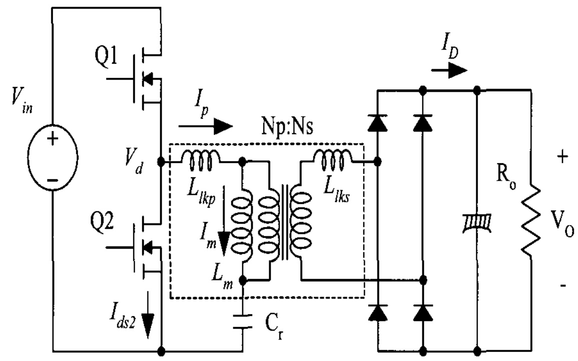


Fig. 1 A schematic of half-bridge LLC resonant converter

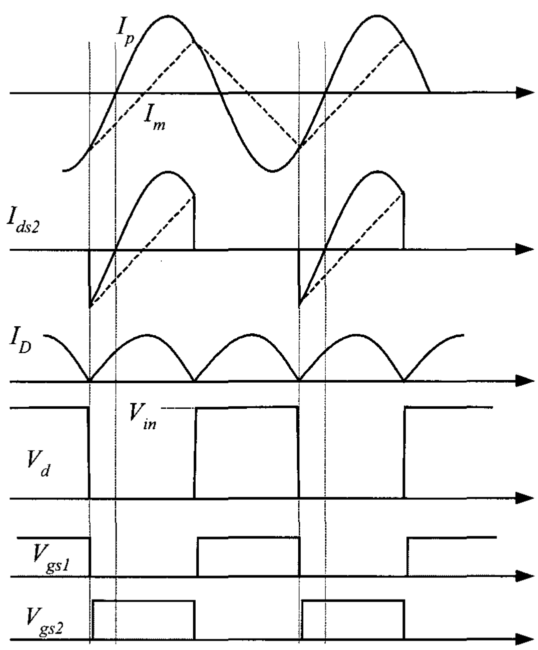


Fig. 2 Typical waveforms of half-bridge LLC resonant converter

It is important to note that the equivalent load resistance shown in the primary side is different from actual load resistance. Fig. 3 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. Considering the transformer turns the ration ($n=N_p/N_s$), the equivalent load resistance shown in the primary is obtained as

$$R_{ac} = \frac{8n^2}{\pi^2} R_o \tag{1}$$

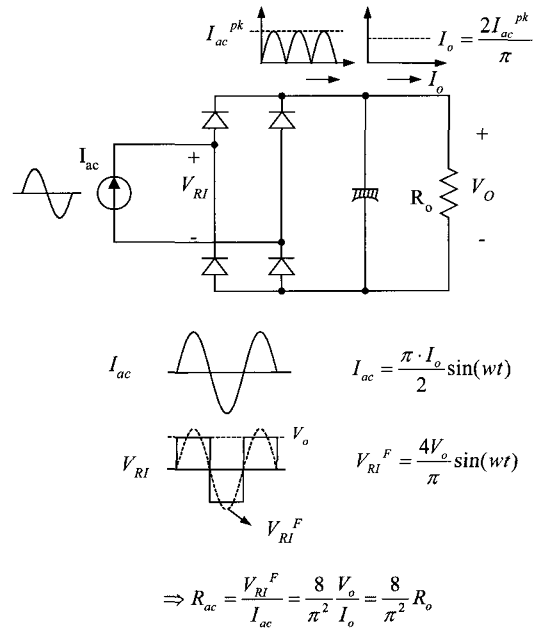


Fig. 3 Equivalent Load resistance R_{ac}

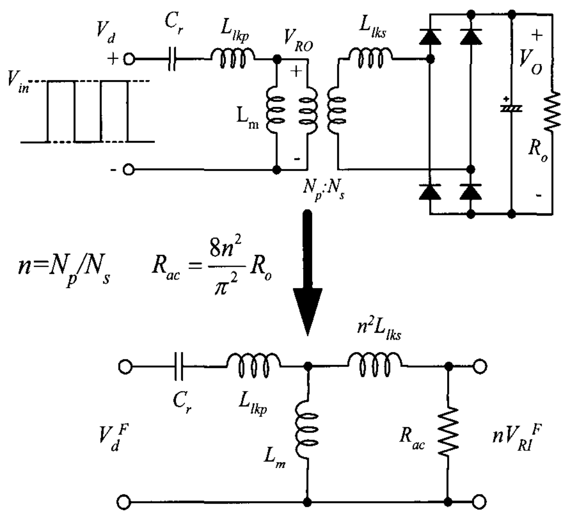


Fig. 4 AC equivalent circuit for the LLC resonant converter

With the equivalent load resistance obtained in (1), the characteristics of the LLC resonant converter can be derived. Using the AC equivalent circuit of Fig. 4, the voltage gain is obtained as

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

where

$$R_{ac} = \frac{8n^2}{\pi^2} R_o \ , \ L_p = L_m + L_{lkp} \ , \ L_r = L_{lkp} + L_m // (n^2 L_{lks})$$

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

As can be seen in (2), 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.

Important feature that should be observed in (2) is that the gain is fixed at resonant frequency (w_o) regardless of the load variation, which is given as

$$M_{@\omega=\omega_o} = \frac{L_m}{L_p - L_r} = \frac{L_m + n^2 L_{lks}}{L_m} \tag{3}$$

Without considering the leakage inductance in the transformer secondary side, the gain in (3) becomes unity. In the previous research, the leakage inductance in the transformer secondary side was ignored to simplify the gain equation [7-12]. However, as observed, there exists considerable error when ignoring the leakage inductance in the transformer secondary side, which generally results in non-optimized design.

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

$$M = \frac{2n \cdot V_O}{V_{in}} = \left| \frac{(\frac{\omega^2}{\omega_p^2})\frac{k}{k+1}}{j(\frac{\omega}{\omega_o}) \cdot (1 - \frac{\omega^2}{\omega_o^2}) \cdot Q \frac{(k+1)^2}{2k+1} + (1 - \frac{\omega^2}{\omega_p^2})} \right| \tag{4}$$

where

$$k = \frac{L_m}{L_{lkp}} \tag{5}$$

$$Q=\frac{\sqrt{L_r/C_r}}{R_{ac}} \tag{6}$$

The gain at the resonant frequency (w_o) is also simplified as

$$M_{@\omega=\omega_o}=\frac{L_m+n^2L_{lks}}{L_m}=\sqrt{\frac{L_p}{L_p-L_r}}=\frac{k+1}{k} \tag{7}$$

By using the gain at the resonant frequency of (7) as a virtual additional gain of the transformer, the AC equivalent circuit of LLC resonant converter of Fig. 4 can be simplified in terms of L_p and L_r as shown in Fig. 5.

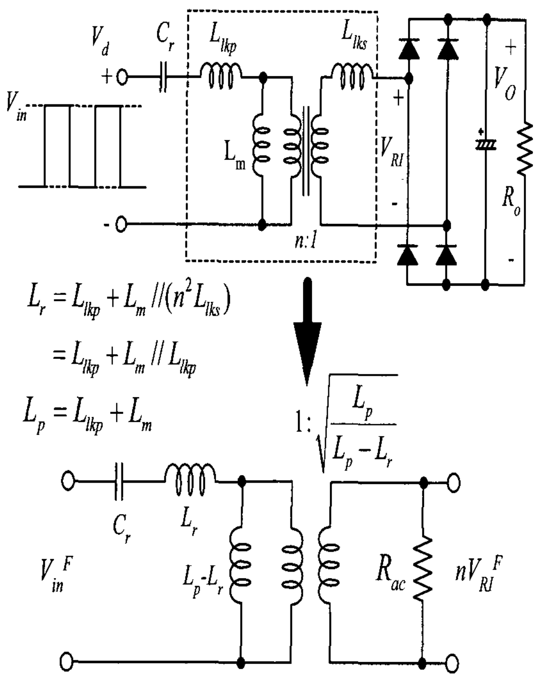


Fig. 5 Simplified AC equivalent circuit for LLC resonant converter

The equation (4) is plotted in Fig. 6 for different Q values with k=5 and f_o=100kHz. As observed in Fig. 6, the LLC resonant converter shows nearly load independent characteristics when the switching frequency is around the resonant frequency. This is a distinctive advantage of the LLC-type resonant converter over conventional series-resonant converters. Therefore, it is natural to operate the converter around the resonant frequency to minimize the switching frequency variation at light load conditions.

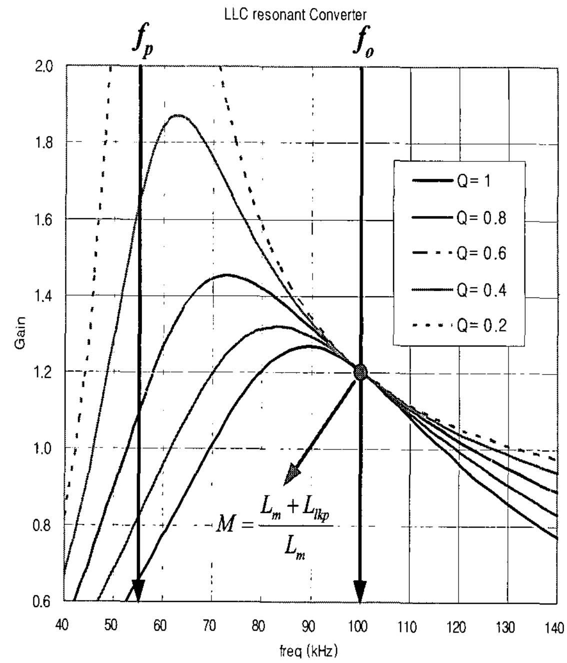


Fig. 6 Typical gain curves of the LLC resonant converter

The operation range of the LLC resonant converter is determined by the available peak voltage gain. The frequency where the peak gain is obtained exists between f_p and f_o as shown in Fig. 6. As Q decreases (as load decreases), the peak gain frequency moves to f_p and higher peak gain is obtained. Meanwhile, the peak gain frequency moves to f_o and the peak gain drops as Q increases (as load increases). 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 (5). Even though the peak gain at a given condition can be obtained by using the gain in (4), it is difficult to express the peak gain in explicit form. Moreover, the gain obtained from (4) 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. 7, which shows how the 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), decreasing k or Q 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.

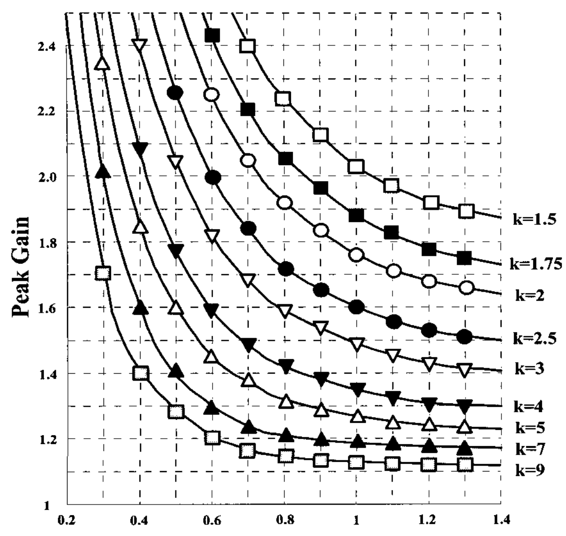


Fig. 7 Peak gain versus Q for different k values

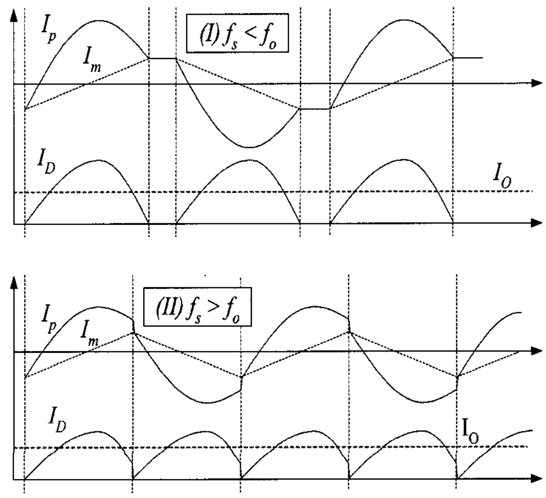


Fig. 8 Waveforms of current in the transformer primary side and secondary sides for difference operation modes

3. Design Procedure

Based on the previous analysis, the practical design procedure is presented in this section. It discusses optimizing the resonant network for given input/output specifications.

3.1 Operation mode

While the conventional LC series resonant converter always operates at a frequency above the resonant frequency, the LLC resonant converter can operate at frequency below or above the resonance frequency. Fig. 8 shows the waveforms of the currents in the transformer primary side and secondary side. As can be observed, operation below the resonant frequency (case I) allows the soft commutation of the rectifier diodes in the secondary side while the circulating current is relatively large. Meanwhile, operation above the resonant frequency (case II) allows the circulating current to be minimized, but the rectifier diodes are not softly commutated. Thus, below resonance operation is preferred for high output voltage application where reverse recovery loss in the rectifier diode is severe. On the other hand, above resonance operation can show better efficiency for application where the output voltage is low and schottky diodes are available for the secondary side rectifiers since the conduction loss is minimize

3.2 Maximum Gain

The maximum gain should be determined considering the input voltage variation. Since the peak gain takes place when the converter operates at the boundary of zero voltage switching (ZVS) and zero current switching (ZCS) mode, ZVS condition is lost at the maximum gain condition. Therefore, some margin is required when determining the maximum gain. Based on the maximum gain, the proper Q and k values can be obtained from Fig. 7. While higher gain is obtained with small k, too small k value results in poor coupling of the transformer. It is typical to set k to be 5~10, which results in a gain of 1.1~1.2 at the resonant frequency.

The value of k affects the losses of the converter. The major portion of the conduction loss is caused by the magnetizing current whose peak value is given by:

$$I_m^{pk} = \frac{nV_o}{L_m}\frac{T_s}{4} \tag{8}$$

The value of k also affects the switching loss. Since the turn-on switching loss is removed by zero voltage switching operation, the turn-off switching loss is dominant. The turn off switching loss is proportional to the turn-off current, which is the same as the peak magnetizing current of (8). Therefore, magnetizing current should be minimized for high efficiency.

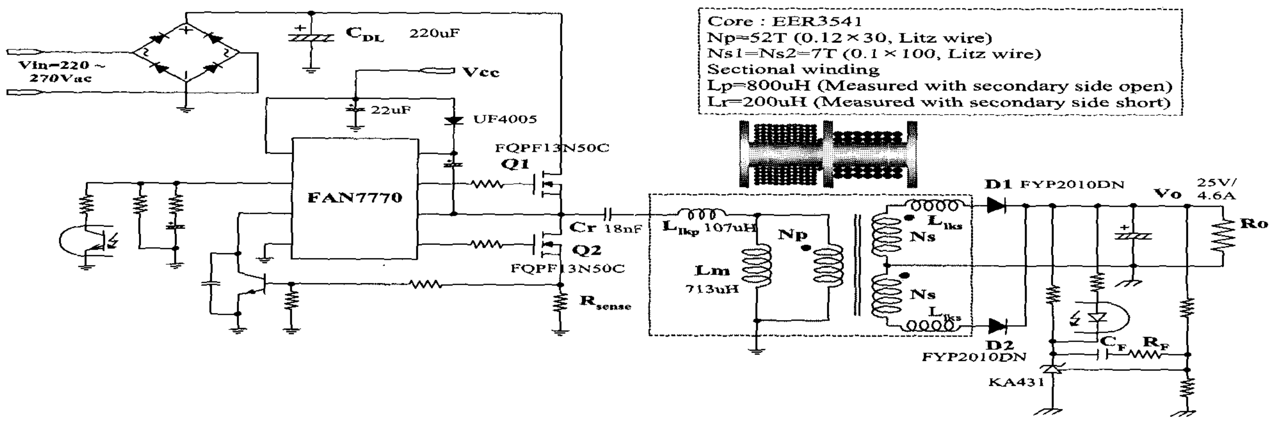


Fig. 9 Schematic of the prototype converter

4. Experimental Verification

In order to show validity of the previous analysis and design consideration, an experimental prototype converter of the 115W half-bridge LLC resonant converter has been built and tested. The schematic of the converter and circuit components are shown in Fig. 9. The input voltage is 220Vac~270Vac and the output is 25V/4.6A. In terms of DC voltage, the input voltage is 260~380V considering holdup time.

The ratio (k) between L_m and L_{lkp} is determined as 6.5, which results in the gain at the resonant frequency as

$$M_{@\,\omega=\omega_o} = \frac{k+1}{k} = 1.15 \tag{9}$$

As observed in the previous analysis, there is a trade-off between the available gain range and conduction loss.

Since the input voltage varies over wide range, if the converter is designed to operate only below resonance frequencies, the excessive circulating current can deteriorate the efficiency. Thus, the converter is designed to operate above resonance at high input voltage conditions and below resonance at low input voltage to minimize the conduction loss caused by circulating current. The minimum gain at full load is determined as 1.0. With the minimum gain, the transformer turn ratio is obtained as

$$n=\frac{N_p}{N_s}=\frac{M_{min}\cdot V_{in}^{max}/2}{(V_o+V_F)}=\frac{1\cdot 380/2}{(25+0.7)}=7.4 \tag{10}$$

where V_F is the diode forward voltage drop.

The maximum gain to cover the input voltage variation is 380/260=1.46. With 10% margin, maximum gain of 1.6 is required. From the gain curves in Fig. 10, Q is obtained as 0.4. By selecting the resonant frequency as 85kHz, the resonant network is determined as L_m=713uH, L_{lkp}=107uH (L_p=800uH, L_r=200uH) and C_r=18nF. The transformer is implemented with a sectional winding method to increase the leakage inductance as depicted in Fig. 9.

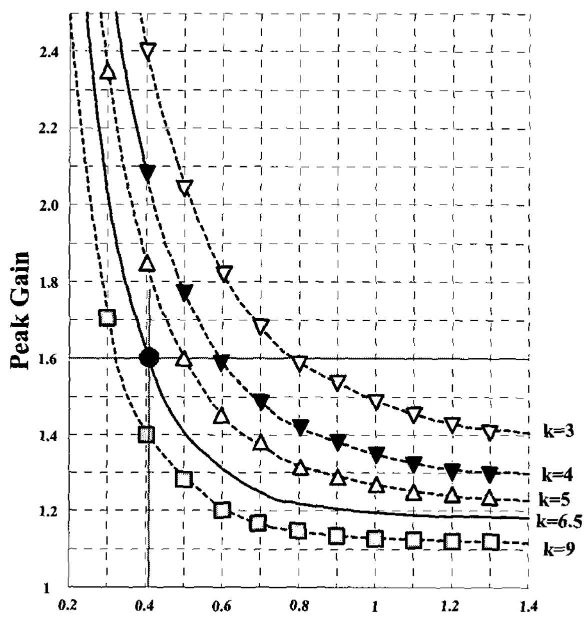


Fig. 10 Design gain curves

Fig. 11 and 12 show the operation waveforms for full load condition with input voltage of 220Vac and 270Vac, respectively. Fig. 13 and 14 show the operation waveforms for no load condition with input voltage of 220Vac and 270Vac, respectively. As can be seen, ZVS is achieved for entire input/output range.

Fig. 15 shows the measured efficiency. As expected, efficiency decreases for low input voltage condition due to the increased circulating current.

Fig. 16 shows the measured switching frequency with load and input voltage variation. Fig. 17 shows the gain curves for different load conditions. As can be seen, the switching frequency variation is well matched with the gain curves of Fig. 17 when the switching frequency is close to the resonant frequency.

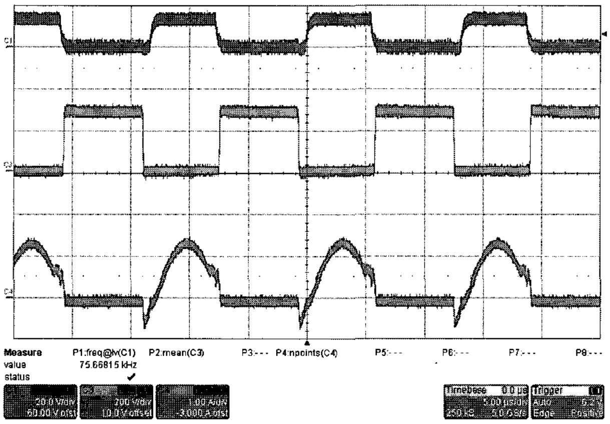


Fig. 11 Operation waveforms : 220Vac input and full load
C1: gate drive signal (20V/div), C3: Drain voltage (200V/div)
C4: Drain current (1A/div)

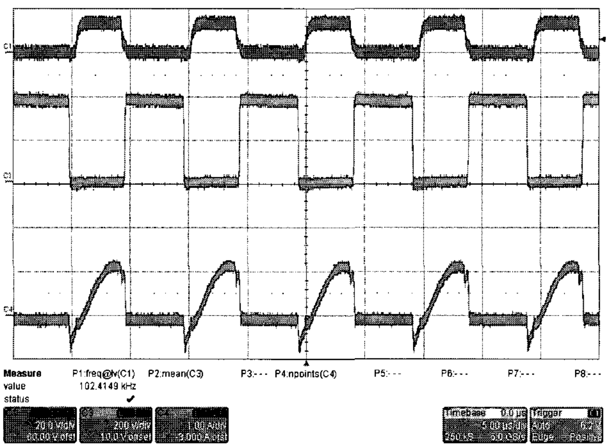


Fig. 12 Operation waveforms : 270Vac input and full load
C1: gate drive signal (20V/div), C3: Drain voltage (200V/div)
C4: Drain current (1A/div)

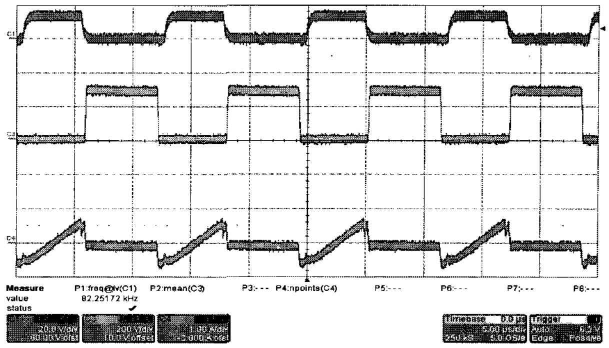


Fig. 13 Operation waveforms : 220Vac input and no load
C1: gate drive signal (20V/div), C3: Drain voltage (200V/div)
C4: Drain current (1A/div)

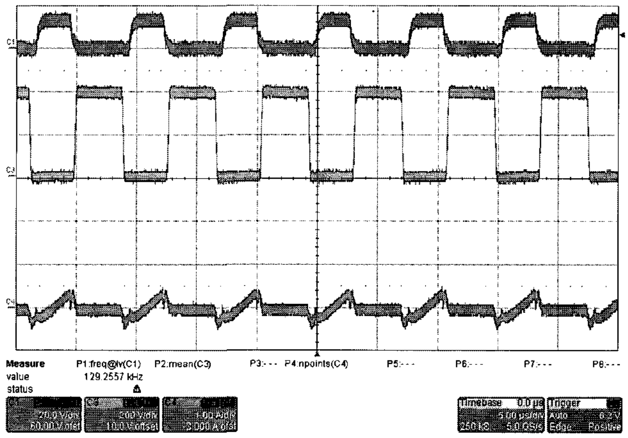


Fig. 14 Operation waveforms : 270Vac input and no load
C1: gate drive signal (20V/div), C3: Drain voltage (200V/div)
C4: Drain current (1A/div)

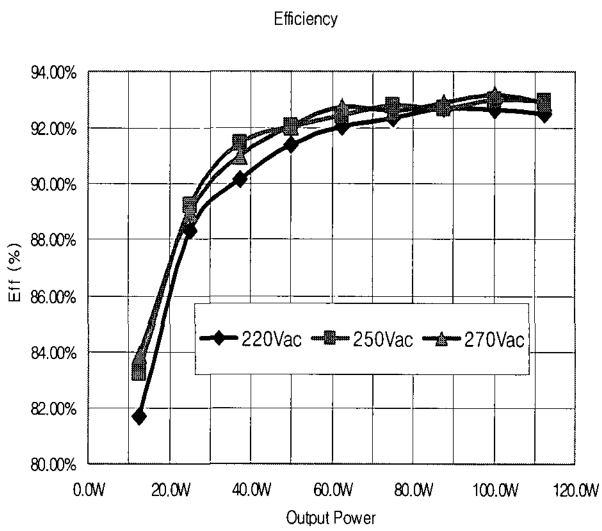


Fig. 15 Measured efficiency

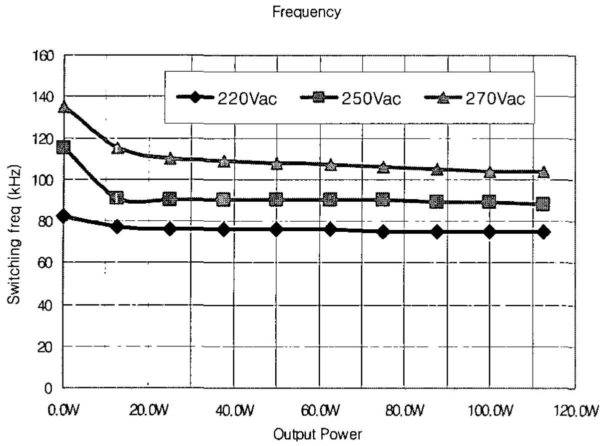


Fig. 16 Measured switching frequency variation

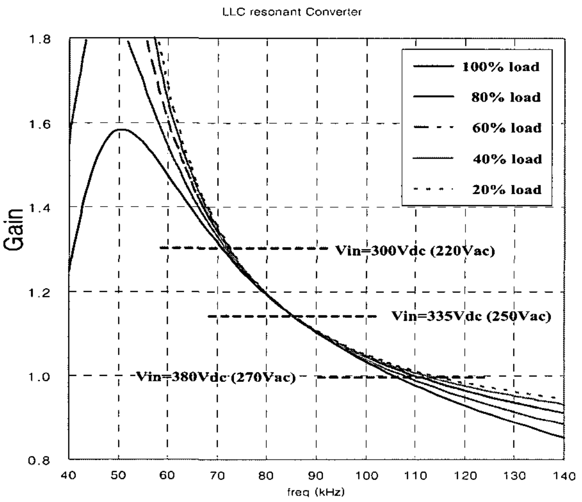


Fig. 17 Gain curves and operating range

5. Conclusions

This paper has presented design consideration for the LLC resonant converter utilizing the leakage inductance and magnetizing inductance of the transformer as resonant components. The leakage inductance in the transformer secondary side was also considered in the gain equation. The design procedure was verified through experimental results.

References

[1] Robert L. Steigerwald, "A Comparison of Half-bridge resonant converter topologies," IEEE Transactions on Power Electronics, Vol. 3, No. 2, April 1988.

[2] A. F. Witulski and R. W. Erickson, "Design of the series resonant converter for minimum stress," IEEE Transactions on Aerosp. Electron. Syst., Vol. AES-22, pp. 356-363, July 1986

[3] Y. G. Kang, A. K. Upadhyay, D. L. Stephens, "Analysis and design of a half-bridge parallel resonant converter operating above resonance," IEEE Transactions on Industry Applications Vol. 27, March-April 1991 pp. 386 - 395

[4] Yasuhito Furukawa, Kouichi Morita, Taketoshi Yoshikawa "A High Efficiency 150W DC/DC Converter," Intelec 1994, pp.148-153

[5] Koichi Morita "Novel Ultra Low-noise Soft switch-mode Power Supply," Intelec 1998, pp.115-122

[6] Ashoka K. S. Bhat, "Analysis and Design of LCL-Type Series Resonant Converter," IEEE Transactions on Industrial Electronics, Vol. 41, No. 1, Feb. 1994.

[7] J. F. Lazar and R. Martineli "Steady-state analysis of the LLC series resonant converter," APEC 2001, 728-735

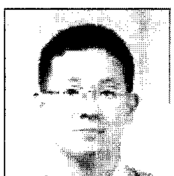
[8] Yan Liang, Wenduo Liu, Bing Lu, van Wyk, J.D, " Design of integrated passive component for a 1 MHz 1 kW half-bridge LLC resonant converter," IAS 2005, pp. 2223-2228

[9] B. Yang, F.C. Lee, M. Concannon,"Over current protection methods for LLC resonant converter," APEC 2003, pp. 605 - 609

[10] Yilei Gu, Zhengyu Lu, Lijun Hang, Zhaoming Qian, Guisong Huang, "Three-level LLC series resonant DC/DC converter," IEEE Transactions on Power Electronics Vol.20, July 2005, pp.781 - 789

[11] Bo Yang, Lee, F.C, A.J Zhang, Guisong Huang, "LLC resonant converter for front end DC/DC conversion," APEC 2002. pp.1108 – 1112

[12] Bing Lu, Wenduo Liu, Yan Liang, Fred C. Lee, Jacobus D. Van Wyk, "Optimal design methology for LLC Resonant Converter," APEC 2006. pp.533-538



Hang-Seok Choi Hang-seok Choi received the B.S., M.S. and Ph.D degrees in electrical engineering from Seoul National University, in 1996, 1999 and 2002, respectively. He is currently working for Fairchild Semiconductor in Bucheon, Korea as a system and application engineer. His research interests include high frequency power conversion, and modeling and control of converters. He has published 14 papers in IEEE conferences and transactions and 10 application notes in Fairchild semiconductor.