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Comparision of Isolated and Non Isolated Resonant Boost Converter

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ABSTRACT

This paper presents comparison of Non-Isolated and Isolated Resonant Boost converter The isolated boost resonant converter size is large more component is presented in the circuit, core saturation problem, more switching losses, high efficiency as compared to the non isolated resonant dc-dc converter. The designed Integrated Boost Resonant Converter employs a unique modulation method for extending the input range of Pulse-Width Modulation, low component count, galvanic isolation, simple control and high efficiency across a wide input and load range. The conventional non-isolated Boost Converter generates switching loss at turn on and off thus the whole system's efficiency is reduced. The proposed isolated converter utilizes soft switching method using a resonant circuit. The implementation of the isolated boost converter that exhibit no parasitic voltage ringing across all semiconductor devices on the primary and secondary sides of the transformer is introduced.

Keywords: Resonant Boost converter, Hybrid frequency modulation technique, Zero Voltage Switching.

1. INTRODUCTION

Boost topology has very high EMI due to reverse recovery of boost diode and very high switching losses due to hard switching of boost switch. Many variations of original boost topology have been suggested to overcome these problems. Increase in the frequency of operation also increases the switching losses and decrease in the system efficiency. So, here comes resonant switch which uses the resonances of circuit capacitances and inductances to shape the waveform of either the current or the voltage across the switching element such that when switching takes place, there is no current through or voltage across it, and hence no power dissipation. In this project zero voltage switching is used. It is a soft switching technique which is used to maintain higher efficiency at higher frequency . The conventional non-isolated boost converter topology has been extensively used in various ac/dc and dc/dc applications. In fact, the front end of today's ac/dc power supplies with power-factor correction (PFC) is almost exclusively implemented with the boost topology. The boost topology is also used in numerous battery-powered applications to generate a high output voltage from a relatively low battery voltage. However in some applications, it may be advantageous to use a boost converter with a galvanic isolated input and output. For example, fault tolerant power systems that use a dual ac-input architecture can be implemented with isolated boost converters, [1], [2]. In fact, the isolated-boost-converter implementation offers a reduced number of components compared to the implementations with non-isolated boost converters in applications which require dual ac input [2]. Also, in applications where a power supply with both ac and dc inputs is required, the isolated boost converter can be applied to provide safety-required isolation between the inputs. So far, a number of boost topologies utilizing an isolation transformer have been proposed,[3] Generally, these circuits exhibit increased voltage stresses on the switches and/or diodes due to the parasitic ringing of the leakage inductance of the transformer with the output capacitances of the switching devices. To control the parasitic ringing voltage, these converters rely on various snubbers, which have detrimental effect on their efficiency and also limit their switching frequency.

One such method of enhancing converter operation is the selection of an appropriate modulation scheme. In this paper, a unique dc–dc converter modulation scheme is proposed for a class of converters that integrate PWM stages into unregulated resonant converters. The resonant stage provides galvanic isolation with high efficiency, while the PWM stage provides the necessary regulation. Though the efficiency is good with a narrow input range and fixed-frequency PWM, it is still possible to extend the operating range while maintaining high efficiency. This new method, a hybrid between constant-on, constant-off and fixed-frequency modulation, optimizes the converter efficiency at the nominal line input while allowing an extended input range. In order to maintain high efficiency under low-power conditions, it is necessary to minimize the amount of circulating energy in the system. One popular option for the dc–dc conversion stage is a simple continuous-conduction-mode flyback converter, It has the benefit of simple construction and low circulating energy. However, the switching loss for both the primary switch and the diode can be quite large, and the overall system efficiency is typically low.(<90%) [2]. Improvements in flyback efficiency can be made using variants such as Zero voltage transition or active clamp, both of which use the transformer leakage inductance as resonant element to achieve Zero-voltage switching (ZVS) across the main device. However this effectively trades switching loss for circulating energy, reducing efficiency at high line or low power [5].

Another option is the series-resonant converter, and more recently the LLC resonant converter, both of which operates on a similar principle and, typically, use a variable frequency control to adjust the output voltage. When the series resonant, or LLC converter, is operated near the resonant frequency of the tank circuit, the converter achieves nearly ZVS and zero-current switching (ZCS) with very low circulating energy, giving it a high peak efficiency. However, as the operating frequency diverges from the resonant frequency, the amount of circulating energy increases.

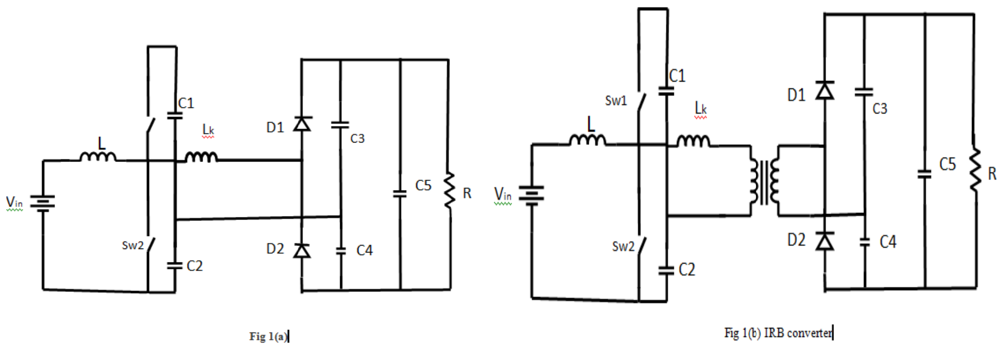

Fig.1(a), 1(b). Isolated and Non Isolated Converter

Several authors have proposed methods to extend the line and load range of the LLC, once again complicating the circuit topology and control [3]. The method proposed in this paper isolated resonant boost converter element into the DC voltage source with only the addition of a single inductor. The overall design is straightforward and may be controlled Hybrid frequency PWM with only the need to observe limitations on the maximum and minimum duty cycle[1]. This circuit satisfies the need for galvanic isolation, low switching loss, minimal circulating energy, as well as simple gate drive and control.

2. CONVERTER PRINCIPLE AND ITS OPERATION

The proposed modulation scheme is developed primarily for circuits that employ this integration of PWM and resonant conversion. One such circuit is the non isolated resonant Boost converter, which is shown in Fig.1. The isolated resonant boost(IRB) rectifier capacitors, C1–C4, are sized appropriately so that they resonate fully with the transformer leakage inductance during each half of the switching cycle. This resonant action occurs simultaneously with a synchronous boost circuit formed by the input inductor and the two switches sw1 and sw2.

Since Bipolar Transistors require a substantial drive current, they are not generally used in resonant converters, unless the base drive is provided by the resonant circuit itself. Power MOSFETs and IGBTs, with their effectively capacitive inputs and low drive energy requirements, are the most frequently used types. MOSFET is used as a switch in this project. The two switches are switched complementary to one another in the proceeding analysis, with the duty cycle D defined for the lower switch sw2. Thus, the boost action is said to be integrated into the resonant converter. Allowing this resonant action to complete fully adds four primary benefits,

1. the output diodes D1 and D2 achieve zero current switching (ZCS);
2. switching loss in the primary-side switches is equal to a normal synchronous boost;
3. the transformer has zero circulating energy;
4. the resonant stage gain is fixed and equal to the transformer turns ratio (1:n).

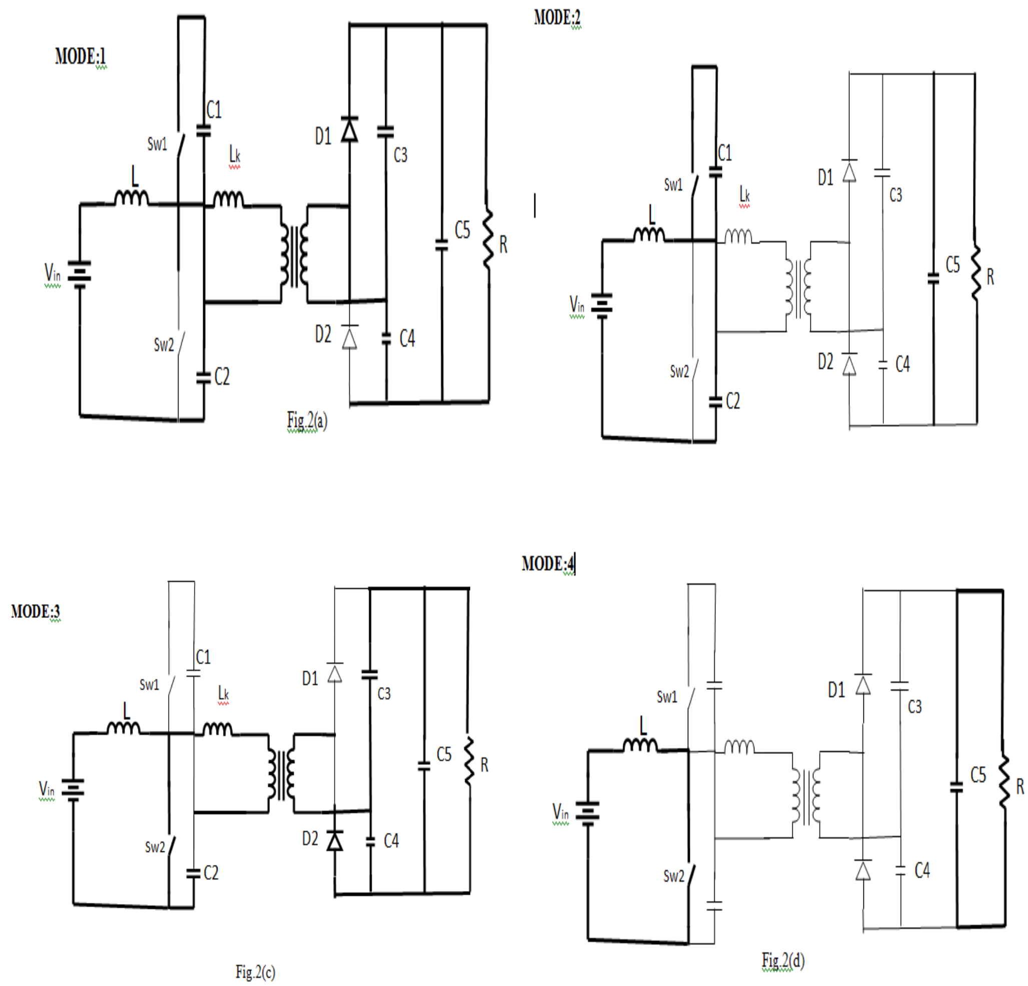

Fig.2(a)-2(d). Modes of operation

3. MODES OF OPERATION

A. MODE 1 (t0 < t < t1); as in Fig. 2(a):

Beginning with the turnoff of sw2 prior to t0 , the current in the input inductor L flows into the body diode of sw1 , discharging its parasitic capacitance. This allows sw1 to be turned ON under ZVS conditions at t0. At this time, the upper input-side capacitor C1 begins resonating with the transformer leakage inductance Lk and the output-side capacitors, C3 and C4, through D1 . Simultaneously, the input current begins charging the series combination of C1 and C2. During this phase, sw1 carries the difference between the transformer current, flowing from C1 through the positive terminal of the transformer and the input current. Once the transformer current resonates back to zero, D1 prevents the continued resonating in the reverse direction, ending mode 1.

B. MODE 2 (t1 < t < t2) as in Fig. 2(b):

sw1 is still active, yet it is only conducting the input inductor current, which is still decreasing, a pathway which is shown in Fig. 4(b). The resonant elements all conduct zero current during this interval. Only C5 continues discharging into the load at this time. Mode 2 ends with the turn-off of sw1 and the subsequent turn-on of sw2.

C. Mode 3 (t2 < t < t3); Fig. 2(c):

After the turn-off of sw1 , but prior the turn-on of sw2 , the inductor current is still shunted into charging the series combination of C1 and C2, this time through the body diode of sw1 , and still decreasing almost linearly. When sw2 is turned ON, the body diode of sw1 is hard commutated, causing some switching loss. At t2, C2 begins to resonate with Lk and the parallel combination of C3 and C4 , through the diode D2 . Simultaneously, the inductor current also flows through sw2, increasing linearly. During this interval,sw2 carries the sum of the transformer current and the inductor current. Thus, the rms current through sw2 is significantly larger than that of sw1, which carries the difference of the two currents. Once the transformer current resonates back to zero, D2 blocks the continued oscillation, marking the end of mode 3.

D. Mode 4 (t3 < t < t4); Fig. 2(d):
The inductor current continues to flow through the lower device, increasing until sw2 is turned OFF and the circuit returns to mode 1. Also, during both modes 3 and 4, sw1 effectively isolates the upper capacitor from charging or discharging.

4. DESIGN PROCEDURE

A. Determining Duty Cycle Limits From Input Requirements:

The most critical element of this design procedure is the identification of the input voltage requirements, so that the duty cycle range is fully utilized. With this converter, there is a direct tradeoff between increased input range and lower rms currents in the circuit. The most basic method involves setting the maximum and minimum duty ratios such that the middle of the input range results in a 50% duty cycle at the converter.

$$D_{\max} = \frac{V_{in,\max}}{V_{in,\min} + V_{in,\max}} \tag{1}$$

$$D_{\min} = 1 - D_{\max} \tag{2}$$

With the nominal input assigned to have a 50% duty cycle, the bus voltage Vbus which is measured across C1 and C2 can be calculated by

$$V_{bus} = V_{in,\max} + V_{in,\min} \tag{3}$$

B. Determining Maximum Resonant Period Lengths:

With the maximum and minimum duty ratios known, the limits for the resonant periods Tres1 and Tres2 may be calculated based on the desired switching period Tsw utilizing the following equations:

$$T_{res_1,\max} = (1 - D_{\max})T_{sw} \tag{4}$$

$$T_{res_2,\max} = D_{\min} T_{sw} \tag{5}$$

C. *Design Transformer*

Based on the calculated Vbus and the desired output voltage Vout , the following equation can be used to calculate the necessary transformer turns ratio, n:

$$n = \frac{V_{out}}{V_{bus}} \tag{6}$$

The transformer design process can be carried out under a number of different procedures; however, the peak V–S product, provided in the following equation, is often a useful quantity when determining the transformer flux density, core size and number of primary turns:

$$VS_{Peak} = \frac{V_{bus}T_{SW}}{4} \tag{7}$$

D. *Design Input Inductor Based on Allowable Current Ripple*

Multiple criteria may be used for designing the input inductor. The following equation specifies the input inductance based on the maximum allowable current ripple (which occurs at D = 0.5)

$$L = \frac{V_{bus}T_{SW}}{4I_{L,avg\%ripple,\max}} \tag{8}$$

E. Resonant Capacitor Design

In order to reduce the rms currents in the circuit, the resonant period needs to be as close to the calculated maximum as possible. Because the leakage inductance of the transformer, Lk , is involved in the resonant circuit and is a consequence of the transformer design in step C, it is left as a constant here. The full equation for calculating Tres1 is given in (4). If the duty cycle was not adjusted as in the second half of step 1, then Tres2 is not necessary to design.

$$T_{res1} = \Pi \sqrt{L_k \left(\frac{2(C_1 + C_2)(C_3 + C_4)}{C_1 + C_2 + n^2(C_3 + C_4)} \right)} \tag{9}$$

$$T_{res2} = \Pi \sqrt{\frac{L_k C_2 n^2 (C_3 + C_4)}{C_2 + n^2 (C_3 + C_4)}} \tag{10}$$

F. Average Diode Current

The output diodes D1 and D2 need to block the full output voltage. The average current they need to carry is given as

$$i_{avg,diode} = \frac{V_{out}}{R_{load}} \tag{11}$$

In order to overcome the drawbacks of traditional PWM, other modulation methods have been proposed, three of the most popular being constant-on, constant-off, and hysteretic control. For the IBR, constant-on modulation provides a selectable minimum on-time that can ensure ZCS during at least one-half cycle. There is no controllable off-time, however; therefore, ZCS is only guaranteed for the output diode for duty cycles greater than 50%. Also, constant-on control requires an extremely wide frequency range, with the maximum frequency occurring only at the minimum input voltage. Similarly, constant-off control provides only a selectable off-time, guaranteed ZCS for only duty cycles less than 50%, with the maximum frequency occurring at the maximum input voltage. With hysteretic control, there is no minimum on- or off-time and no guarantee of ZCS. In order to improve the utilization and efficiency of the IRB converter under a wide load range, the designed modulation scheme needs to have a selectable minimum on- and off-time, guaranteed ZCS across the operating range, narrow operating frequency band, and maximum frequency occurring at 50% duty cycle so as to minimize the transformer core loss. The input voltage reference is generated by the maximum-power-point tracking (MPPT) loop, passing a reference to the input voltage control loop. The normal output of the digital compensator is the only required input to the hybrid frequency modulator, and the output works naturally with a PWM comparator that requires both a switching period length and a value for the main switch on-time.

5. SIMULATION PARAMETERS

Table-1. POWER STAGE ELEMENT VALUES FOR 250-W PROTOTYPE

ELEMENT	VALUE	ELEMENT	VALUE
Tres1	4.61µs	C1,C2	10µF
Tres2	4.03µs	C3,C4	100nF
Tsw	14.3µs	C5	2µF
Fsw	70KHz	Vin,min	20V
Dmin	0.33	Vout,max	40V
Dmax	0.67	Vout	400V
L	100µH	Vbus	60V

There is a significant difference in the circuit behavior between the two resonant modes (1 and 3). During mode 3, C1 is effectively isolated from the rest of the circuit due to the presence of sw1. However, during mode 1, C2 has an ac discharge path through the DC source and the input inductor, allowing the two input capacitors (C1 and C2) to appear in parallel, though the resonant current is not shared evenly between them. Thus, the length of mode 1 can be significantly longer than mode 3 as in Fig. 4 and Fig. 5, depending on the relative sizing of C1–C4 and the transformer turns ratio.

According to the fig1(a) the simulation results of the Isolated boost converter are as follows:

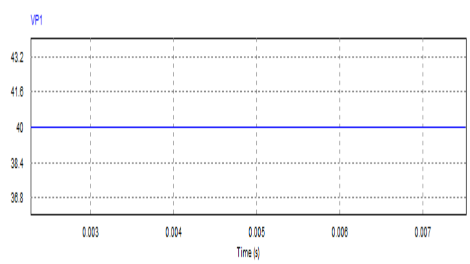

Fig. 3. Input Voltage

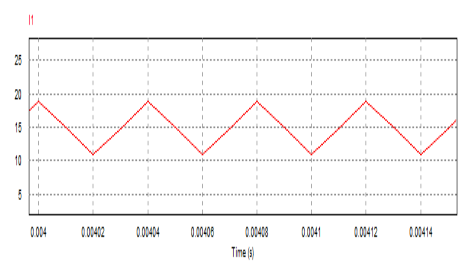

Fig. 4. Input Current

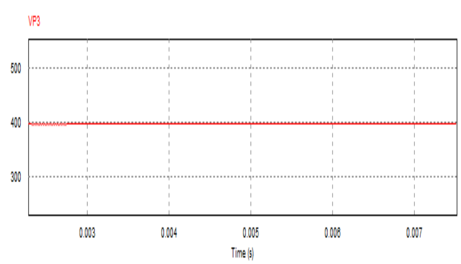

Fig. 5. Output Voltage

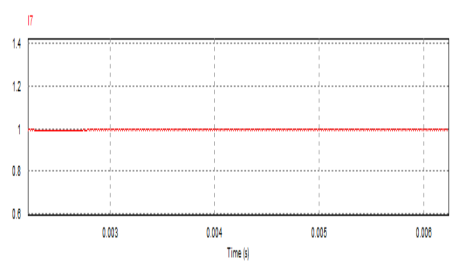

Fig. 6. Output Current

According to fig1(b) the simulation results of non isolated boost converter are:

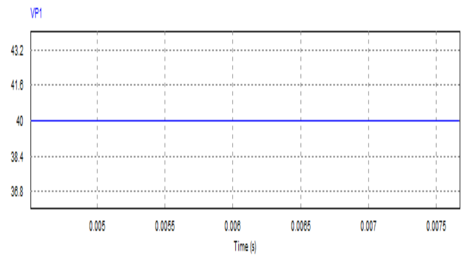

Fig. 7. Input Voltage

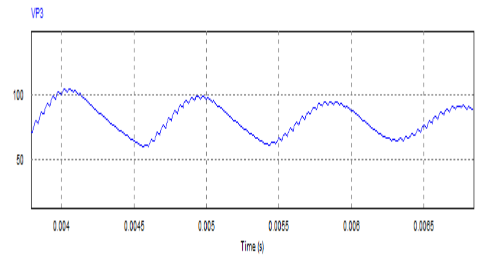

Fig. 8. Input Current

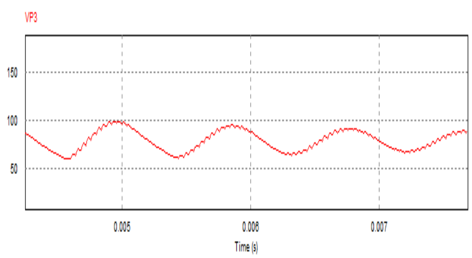

Fig. 9. Output Voltage

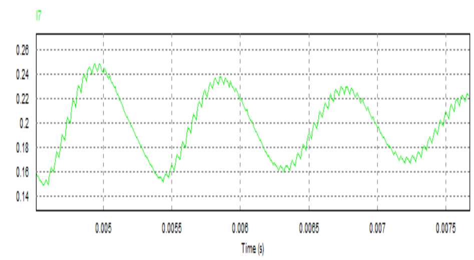

Fig. 10. Output Current

6. COMPARISION OF ISOLATED AND NON ISOLATED BOOST RESONANT CONVERTER RESULTS:

The isolated and non isolated boost resonant converter are simulated and the results are compared and tabulated as follows:

Table 1: For isolated boost resonant converter

Input Voltage	Input Current	Input Power	Output Voltage	Output Current	Output Power	Efficiency
30	14	420	300	0.75	348.6	83
40	20	800	400	1	672	84
50	24	1200	500	1.2	1032	86
60	29	1740	600	1.5	1531	88

The simulated results are compared and the graphs are plotted between input voltage and efficiency for isolated and non isolated boost resonant converter as follows:

Table 2: For non isolated boost resonant converter:

Input Voltage	Input Current	Input Power	Output Voltage	Output Current	Output Power	Efficiency
30	15	450	70	0.17	360	80
40	18	720	100	0.25	590.4	82
50	20	1000	122	0.3	830	83
60	30	1800	150	0.38	1530	85

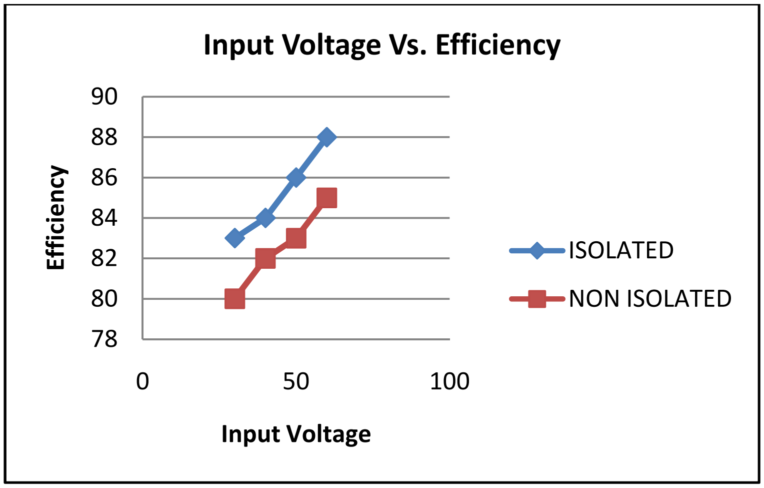

Fig. 11. Input Voltage Vs. Efficiency

7. CONCLUSION

As a solution, the comparison of isolated and non isolated resonant boost converter has been performed. The isolated resonant dc-dc converter is quite good and more sophosticated compared to the non-isolated resonant dc-dc converter. In this paper, the isolated boost dc-dc converter is performed with resonance to produce more efficiency in the output side, reduce the switching losses, reduce the conduction losses, reduce the harmonic with out the transformer are presented.

The principle advantages of utilizing this (IBR) were as follows:

1. high weighted efficiency because of low circulating energy and reduced switching loss with resonant energy transfer and output diode ZCS;
2. low potential cost due to minimal number of active devices and a small overall component count;
3. galvanic isolation allows for the use of high efficiency inverter stages without additional concern over ground leakage current;

4. reduced control complexity provides lower auxiliary power loss and simpler controller IC configurations. Finally, simulation verification of proposed modulation method was presented.

REFERENCES

[1] W.P. Turner IV and K.G. Brill, “Industry standard tier classifications define site infrastructure performance,” White Paper Uptime Institute. http://www.upsite.com/TUIpages/whitepapers/tuitiers.html

[2] M.M. Jovanović, “Dual AC-Input Power System Architectures,” IEEE Applied Power Electronics (APEC) Conf. Proc., pp. 584-589, Mar. 2002.

[3] P. W. Clarke, “Converter Regulation by Controlled Conduction Overlap,” U.S. Patent 3,938,024, Feb. 10, 1976.

[4] C.K. Tse and M.H.L. Chow, “New Single-stage PFC Regulator Using Sheppard-Taylor Topology,” IEEE Trans. Power Electronics, vol. 13, pp. 842-851, Sep. 1998.

[5] Ben York, Wensong Yu and jih-Sheng Lai(Feb 2013) ‘Hybrid Frequency Modulation for PWM-Integrated Resonant Converter’ IEEE Trans on power electronics, vol.28,no.2.

[6] Chung B.-G., Yoon K.-H, Phum S, Kim E.-S, and Won J.-S(2011) ‘A novel LLC resonant converter for wide input voltage and load range’, in Proc.Int. Conf. Power Electron./ECCE Asia , pp. 2825–2830.

[7] Hsieh Y.-C, Chen M.-R, and Cheng H.-L.(Jan.2011) ‘An interleaved flyback converter featured with zero-voltage transition’, IEEE Trans. Power Electron., vol. 26,no. 1,pp. 79-84.

[8] Kim C.-E, Moon G.-W, and Han S.-K (Nov.2007) ‘Voltage doubler rectified boost integrated half bridge (VDRBHB) converter for digital car audio amplifiers’, IEEE Trans. Power Electron., vol. 22, no. 6, pp. 2321–2330.

[9] Lazar J. F and MartinelliR(2001) ‘Steady-state analysis of the LLC series resonant converter’, in Proc. Appl. Power Electron. Conf., 2001, vol. 2,pp. 728–735.