A Resonant Converter with Wide Input Range

Je-Hyung Cho, Ki-Bum Park, Gun-Woo Moon, Myung-Joong Youn

Division of Electrical Engineering School of Electrical Engineering & Computer Science, KAIST
335 Gwahangno (373-1 Guseong-dong), Yuseong-gu, Daejeon 305-701, Republic of Korea

Abstract

A resonant converter suitable for applications of wide input variation is proposed. A half-bridge converter and a phase-shift converter are combined in primary and the secondaries of each transformer are connected in series which makes four voltage levels. Furthermore, by adopting resonant converter scheme, a resonant capacitor is connected with the secondaries of the transformers in series and an output inductor is removed. Zero voltage switching is achieved from full load to no load. The analysis of the resonant converter and experimental results of 200W prototype is presented to verify the features of the converter.

1. Introduction

Nowadays, the switching frequency of switched mode power supplies(SMPS) are increasing to achieve high power density. As the switching frequency increases, the switching losses of switches become more important issue on design of SMPS. Furthermore, some applications require high efficiency under the wide load range and the wide input variation. Soft-switching converters which achieve zero voltage switching (ZVS) so that the switching losses can be reduced have been proposed.

Among them, the phase-shift full-bridge converter[1] is one of the most popular topologies from middle to high power applications. However the phase-shift full-bridge has a narrow ZVS range on lagging leg switches and high circulating energy.

To achieve ZVS along wide load range, resonant converters, which use resonant components including parasitic components to reduce switching losses, have been used. LLC resonant converter has a wide ZVS range from nearly zero to full load. However, when the resonant converter deals with the wide input variation, the range of the switching range is getting too wide. That makes the optimization of the magnetizing components of the converter difficult.

Several methods to manage the wide input variation have been presented. The hybrid converter[2], which is combination of the half-bridge converter and the phase-shift full-bridge converter and uses fixed frequency control, can be used to operate under the wide input variation.

A resonant converter with a wide input range is proposed in this paper. The proposed converter has two control schemes, which are phase-shift and frequency control. ZVS can be achieved along a wide load range and under a wide input range. Moreover, an output inductor is removed. The 200W experimental prototype is presented to verify the features of the converter.

2. Operational Principles

The proposed converter is hybrid combination of a half bridge converter and a phase-shift full-bridge converter and they share the one leg of switches. All the switches are operated in one-half duty,

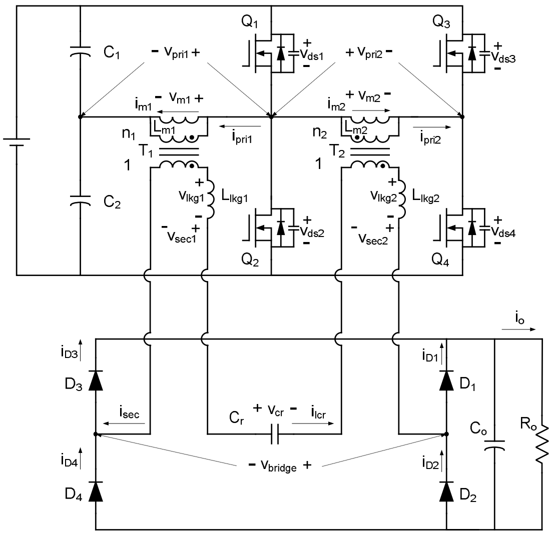

Fig. 1 Proposed Converter with wide input Range

and the voltage applied to the full-bridge converter is controlled by using phase-shift modulation which controls the effective duty of the converter.

The secondaries of each transformer are connected in series and the resonant capacitor connecting with them in series at the secondary side forms a resonant tank with the leakage inductances. The leakage inductance of the transformer is reflected to the secondary as a matter of convenience of analysis.

Mode 1 (t_0~t_1) : In this mode, the power is transferred from primary to secondary. In the half-bridge section, since $V_s/2$ is applied to the primary of transformer, the magnetic current of T_1, i_{m1} is increased by $(V_s/2)/L_{m1}$. In the full-bridge section, the magnetic current of T_2, i_{m2} is increased by V_s/L_{m2}. C_r, L_{lkg1} and L_{lkg2} resonate each other with input voltage of v_{in} of fig. 3, i_{lkg1} and i_{lkg2} have resonant current shapes. The sum of the secondary voltages of two transformers, v_{bridge} is $(V_s/2)/n_1 + V_s./n_2$.

Mode 2 (t_1~t_2) : When Q_4 is turned off, C_{oss3} and C_{oss4} are discharged and charged by i_{pri1} and i_{pri2}, respectively. After the primary voltage of transformer T_2, v_{pri2}, decreases to zero, D_1 and D_4 are still conducting. Voltage across L_{lkg1} and L_{lkg2}, V_o - v_{cr}, is divided on L_{lkg1} and L_{lkg2} due to impedance ratio of L_{lkg1} and L_{lkg2}, and that makes i_{pri1} and i_{pri2} decreases. The sum of the secondary voltages of two transformers keeps up $(V_s/2)/n_1 + V_s./n_2$ until this mode ends. This mode ends when the primary current of each transformer meets the magnetizing current of each transformer.

Mode 3 (t_2~t_3) : In the half-bridge section, since $V_s/2$ is still applied to the primary of transformer, T_1, the primary current of T_1, i_{pri1}, is increased by $(V_s/2)/L_{m1}$. In the full-bridge section, the primary current of T_2, i_{pri2}, equals to I_{m1}, keep constant current

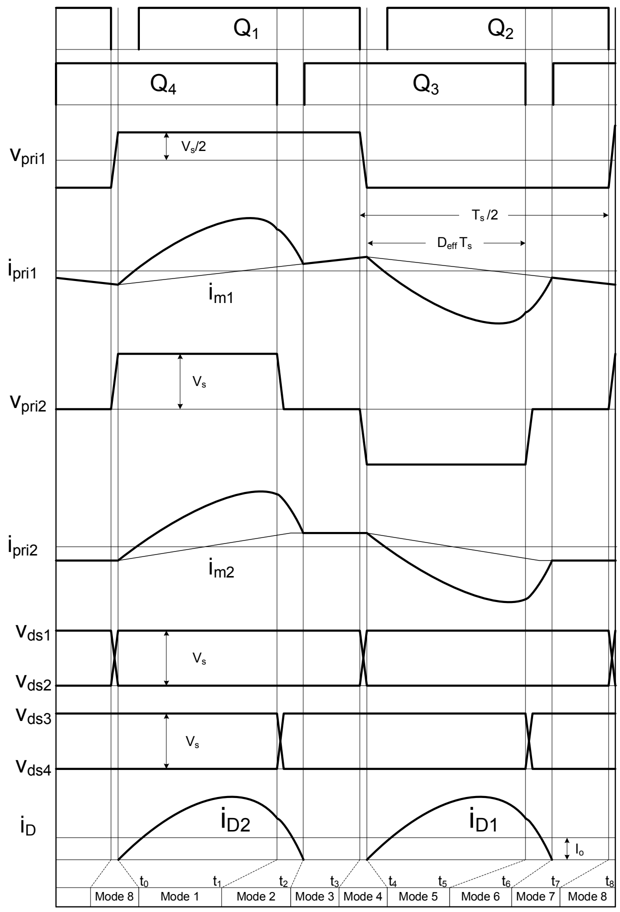

Fig. 2 Key Waveforms of Proposed Converter

$i_{m2}(t_2)$. The sum of the secondary voltages of two transformers, becomes $(V_s/2)/n_1$.

Mode 4 (t_3~t_4) : When Q_1 is turned off, C_{oss1} and C_{oss2} are charged and discharged by i_{pri1} and i_{pri2}, respectively. Therefore, v_{pri1} decreases from $V_s/2$ to $-V_s/2$ and v_{pri2} decreases from 0 to $-V_s$, v_{bridge} decreases to $-(V_s/2)/n_1 - V_s/n_2$.

The operations from Mode 5 to Mode 8 are similar to those from Mode 1 to Mode4, respectively.

3. Analysis and Design of Proposed Converter

3.1 DC Conversion Ratio

The assumption for getting DC conversion ratio is that the durations of the switch voltage transition time in Mode 2 and Mode 4 is negligible. So the diode current of those time duration does not contribute to the output current. The capacitance of output capacitor C_o is large enough to be regarded as voltage source V_o.

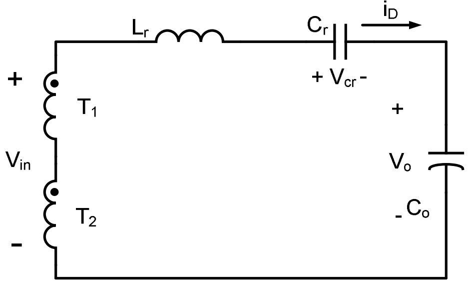

Fig. 3 Simplified Resonant Circuit Secondary of Proposed Converter

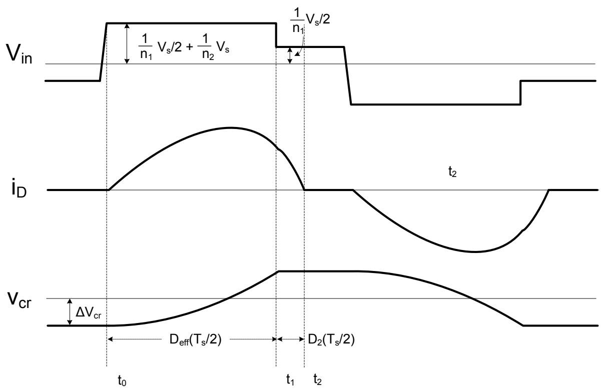

Fig. 4 Simplified Waveforms of I_D, and V_{cr}

Fig. 3 is the simplified resonant circuit of the secondary of the proposed converter. A resonant inductor L_r is the sum of the leakage inductances of transformers, L_{lkg1} and L_{lkg2}. Since the components composing the secondary are L_r, and C_r, which make resonant tank, the secondary current $i_D(t)$, that is the sum of $i_{D1}(t)$ and $i_{D2}(t)$, has sinusoidal waveform during Mode 1 and Mode 2. Since when the mode is changed the input voltage of the equivalent resonant circuit is changed at the same time, the resonant current i_D should be derived in each mode. The DC conversion ratio can be calculated by using that the averaged resonant current i_D during one switching period is equal to the output current I_o.

First, start with calculate the resonant current of Mode 1. The initial condition for this mode is $i_D(t_0) = 0$, and $V_{cr} = -\Delta V_{cr}$ where ΔV_{cr} is a variation of the resonant capacitor voltage. The secondary current $i_D(t)$ during Mode 1 is expressed as following formula of LC resonant circuit.

$$i_D(t-t_0) = I_p(t_o)\sin(\omega_r(t-t_0)) \tag{1}$$

where $I_p(t_o) = \dfrac{V_{in}(t_0) - V_o - V_{cr}(t_0)}{Z_r}$, $\omega_r = \dfrac{1}{\sqrt{L_r C_r}}$, $Z_r = \sqrt{\dfrac{L_r}{C_r}}$,

$$L_r = L_{lkg1} + L_{lkg2}, \quad V_{in}(t_0) = \frac{1}{n_1}V_s/2 + \frac{1}{n_2}V_s$$

The resonant capacitor voltage V_{cr} can be calculated as follows.

$$\begin{aligned} V_r(t-t_0) &= V_{cr}(t_o) + \frac{1}{C_r}\int_{t_0}^{t} i_D\,dt \\ &= V_{cr}(t_o) + \frac{1}{C_r}\frac{I_p(t_o)}{\omega_r}\left(1-\cos\left(\omega_r\left(t-t_0\right)\right)\right) \end{aligned} \tag{2}$$

The resonant capacitor voltage V_{cr} is biased at zero and the variation of the resonant capacitor voltage ΔV_{cr} can be calculated by integrating the resonant current flowing into capacitor during the resonant capacitor is charged. Since the average of I_D is equal to the output current I_o, integrated resonant current during one half switching period is equal to the output current I_o times the half of period $T_s/2$. The voltage variation $2\Delta V_{cr}$ is

$$2\Delta V_{cr} = \frac{1}{C_r}\int_{t_0}^{t_0+T_s/2} i_D\,dt = \frac{1}{C_r}\frac{T_s}{2}I_o \tag{3}$$

Since $V_{cr}(t_0)$ is $-\Delta V_{cr}$, $V_r(t)$ is expressed as follows.

$$V_{cr}(t-t_0) = -\frac{1}{C_r}\frac{T_s}{4}I_o + \frac{1}{C_r}\frac{I_p(t_o)}{\omega_r}\left(1-\cos\left(\omega_r\left(t-t_0\right)\right)\right) \tag{4}$$

Second, in Mode 2, the initial condition of the resonant current is $i_D(t_1)$, and that of the resonant capacitor voltage is $V_{cr}(t_1)$. Since t_1 is $D_{eff}\,T_s/2 + t_0$, the initial conditions can be calculated as follows.

$$i_D(t_1) = I_p(t_o)\sin\left(\omega_r\left(D_{eff}\ T_s/2\right)\right) = I_p(t_o)\sin\left(\pi\left(\frac{f_r}{f_s}\right)D_{eff}\right) \quad (5)$$

$$V_{cr}(t_1) = -\frac{1}{C_r}\frac{T_s}{4}I_o + \frac{1}{C_r}\frac{I_p(t_o)}{\omega_r}\left(1-\cos\left(\pi\left(\frac{f_r}{f_s}\right)D_{eff}\right)\right) \quad (6)$$

The resonant current i_D has resonant form in Mode 2 with above initial conditions.

$$i_D(t-t_1) = i_D(t_1) + I_p(t_1)\sin(\omega_r(t-t_1)) \quad (7)$$

where $I_p(t_1) = \dfrac{V_{in}(t_1) - V_o - V_{cr}(t_1)}{Z_r}$, $V_{in}(t_1) = \dfrac{1}{n_1}V_s/2$

At the end of this mode, t_2, i_D decreases from $i_D(t_1)$ to zero. The end time of this mode, t2, can be obtained from above eq. (4) using $i_D(t_2\text{-}t_1) = 0$. D_2 is defined as $t_2 - t_1 = D_2(T_s/2)$.

The DC conversion ratio can be derived from the relationship that the average of $i_D(t)$ is equal to output current I_o.

$$I_o = \frac{V_o}{R_o} = \frac{1}{T_s/2}\int_{t_o}^{t_o+T_s/2} i_D\ dt$$

$$= \frac{1}{T_s/2}\left[\int_{t_o}^{t_1} i_D(t-t_0)\ dt + \int_{t_1}^{t_2} i_D(t-t_1)\ dt\right] \quad (8)$$

Where $\displaystyle\int_{t_o}^{t_0+D_{eff}T} i_D(t-t_0)\ dt = \frac{I_p(t_0)}{\omega_r}\left(1-\cos\left(\pi\left(\frac{f_r}{f_s}\right)D_{eff}\right)\right)$,

$$\int_{t_1}^{t_2} i_D(t-t_1)\ dt = i_D(t_1)\left(D_2\frac{T_s}{2}\right) + \frac{I_p(t_1)}{\omega_r}\left(1-\cos\left(\pi\left(\frac{f_r}{f_s}\right)D_2\right)\right)$$

By solving equation above equations, the DC conversion ratio is derived. However, it is hard to achieve closed form DC conversion ratio, since to solve for t_2 is not explicit. It can be seen using numerical and graphical methods.

In above analysis for DC conversion ratio, it is shown that DC conversion ratio is a function of an effective duty (D_{eff}) and a ratio of a switching frequency and a resonant frequency (f_s / f_r). That means various control methods for this converter can be derived. 1) Phase-shift modulation controls the effective duty, 2) frequency modulation controls ratio of resonant frequency and switching frequency, and 3) mentioned two modulations can be combined for dealing with wide input voltage using phase shift feed-forward modulation and for regulating output voltage regulation using frequency modulation.

When the control method is determined, one of considerations is what frequency range will be used among above resonance and below resonance. The characteristic of above and below resonance will be analyzed to make design guide.

4. Experimental Results

A 200W prototype of the proposed converter was built to verify the analysis. Table. 1 shows the components and parameters used for the experiment. Experimental results show the proposed converter achieves ZVS on 10% load condition, and the experimental waveforms are similar to the analyzed key waveforms.

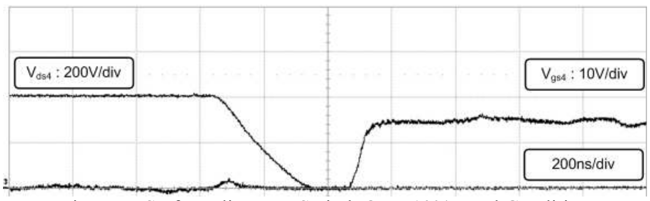

Fig. 5 ZVS of Leading Leg Switch Q_4 at 10% Load Condition

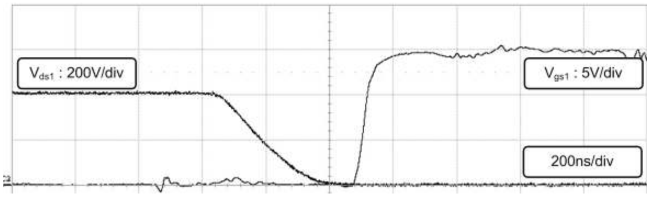

Fig. 6 ZVS of Lagging Leg Switch Q_1 at 10% Load Condition

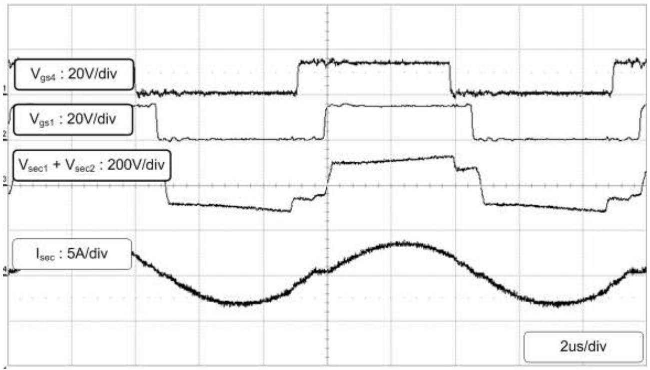

Fig. 7 Operational Waveforms at V_{in} = 200V, Full-load

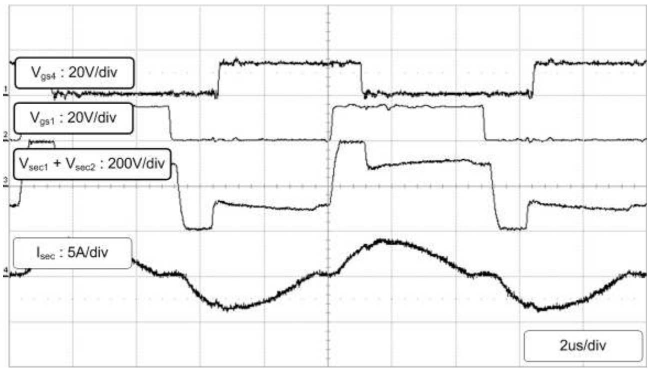

Fig. 8 Operational Waveforms at V_{in} = 400V, Full-load

5. Conclusion

A resonant converter with wide input range is proposed and verified by experiments of 200W prototype. The structure of the proposed converter is hybrid combination of a half bridge converter and a phase-shift full-bridge converter and a large output inductor is removed. A resonant tank comprising leakage inductances of transformers and a resonant capacitor connected in series to them is formed in secondary of the converter. It achieves ZVS along wide load range easily without increasing a circulating current, and operates under wide input voltage variation by using phase-shift control. The proposed converter is suitable for from middle to high power applications with mentioned features.

Reference

[1] J.A. Sabate, V Vlatkovic, R.B. Ridely, F.C. Lee, and B.H. Cho, “Design considerations for high-voltage high-power full-bridge zero-voltage-switched PWM converter”, APEC ’90, pp. 275-284, 1990, March.

[2] R. Ayyanar, and N. Mohan, “A novel full-bridge DC-DC converter for battery charging using secondary-side control combines soft switching over the full load range and low magnetics requirement”, IEEE Transactions on Industry Applications, Vol. 37, pp. 559-565, 2001, March/April.