A control method based on power factor for improving output voltage stability and efficiency of LLC converter in wide range of output voltage and load impedance

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A Control Method Based on Power Factor for Improving Output Voltage Stability and Efficiency of LLC Converter in Wide Range of Output Voltage and Load Impedance

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***Abstract*—The LLC converter is attractive for its small volume and comparatively high efficiency. However, the application of this converter is limited to the power supply to a static load with almost constant output voltage and current because the LLC design covering a wide range in the output voltage and the load variation can deteriorate the output voltage stability and efficiency. This paper addresses this issue by proposing a novel control method. Unlike the conventional control, which adjusts the operating frequency to stabilize the output voltage, the proposed control adjusts the power factor of the transformer primary-side to achieve the same excellent dynamic performance as the buck chopper. Simulation supported the effectiveness of the proposed control at a wide load variation range.**

Keywords—control, efficiency, LLC converter, output stability, power factor

I. Introduction

The LLC converter [1]–[4] has been emerging as a promising dc-dc converter topology with small volume and high efficiency, which are attractive for the power supply circuit in industry and commercial electronics. The key feature of the LLC converter lies in the utilization of the resonance between the inductor and the capacitor, unlike the other typical dc-dc converters like the buck chopper and the boost chopper. The resonance enables the soft-switching operation of the switching devices, which reduces the switching loss to improve the efficiency. Furthermore, the LLC converter can therefore operate at a high frequency, which leads to the size reduction of the passive components as the inductors and the capacitors.

Owing to these attractive features, the LLC converters are recently expected to replace the conventional buck chopper. For example, the LLC converter already found application in LED lamp driver [5][6] and battery charger [7][8]. These applications are characterized by little change in the output voltage and load current. Therefore, they do not need sophisticated control to stabilize the output voltage against the change in the load current. However, for various applications requiring stabilization of the output voltage against the sudden load current change under a wide range of the output voltage gain and the load current, the LLC converter has still difficulty in replacing the conventional buck chopper.

For example, the researchers are recently investigating the application of the LLC converter to the electric vehicles to generate the 12.5V dc bus voltage from the high voltage battery power [1][2], which tends to have a variation of the dc voltage due to the state of charge. In this application, the LLC converter is required to stabilize the output voltage against the sudden load current change under a wide range of the operating condition, i.e. the output voltage gain and the load current. However, the dynamic performance of the LLC converter is reported to be highly dependent on the output voltage gain, as well as the load current, and may exhibit large output impedance at certain output conditions [9]–[11].

Furthermore, the transformer in the LLC converter has been required to have small magnetizing inductance for improving the output voltage stability according to the conventional control method, particularly if the converter is designed to cover a wide range of the output voltage gain and the load current [12]. Therefore, the LLC converter can suffer from worse efficiency at the light load condition due to the large primary current.

As discussed in this paper, this drawback is mainly originated from the basic concept of the conventional LLC converter control, which is based on the adjustment of the operating frequency [13]–[15]. Therefore, to overcome this difficulty, this paper proposes an LLC converter control method based on the novel idea that adjusts the phase factor of the transformer primary winding rather than the operating frequency to regulate the output voltage. Circuit analysis revealed that the proposed LLC converter control can exhibit the same control characteristics as the buck chopper, which is commonly utilized for the highly stable fast response dc-dc converter as POL converters [16][17] and exhibits stable dynamic performance unaffected by the output voltage gain and the load current. Therefore, the LLC converter with the proposed control can have excellent output voltage stability regardless of the output voltage gain and the load current, thus

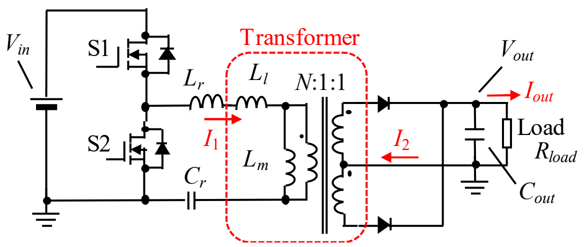


Fig. 1. Schematic diagram of LLC converter

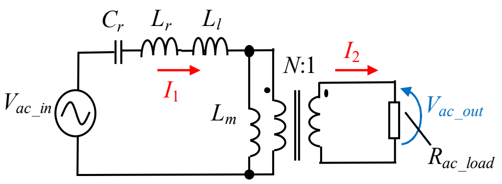


Fig. 2. Equivalent circuit of LLC converter

offering better output voltage stability than the conventional LLC converter control in a wide range of the operating conditions. Furthermore, the proposed control enables the LLC converter to have higher magnetizing inductance than the conventional control, which can reduce the primary current and therefore improve the efficiency.

The remainder of this paper comprises five sections. Section II briefly reviews the principles of conventional control and analyzes the reason for its drawbacks. Then, section III proposes a novel fundamental concept of the proposed LLC converter control and discusses how the proposed control can overcome these drawbacks. Then, section IV proposes a practical control circuit to implement the proposed control. Based on this control circuit, section V performs the simulation and reports the results of the dynamic characteristics, the output voltage stability, and the transformer primary current in comparison with the buck chopper and the LLC converter with the conventional control. Finally, section VI gives the conclusions.

II. Review of Conventional Control

Figure 1 shows the typical circuit topology of the LLC converter. The transformer is expressed by the T-shaped network of the leakage inductance L_l and the magnetizing inductance L_m. The primary winding is connected with the resonant capacitor C_r to form the LC resonator with the leakage inductance L_l and the additional resonant inductor L_r. This resonator is connected to the output of the half-bridge circuit of S1 and S2 to excite the resonance. Then, the secondary winding is connected with the rectifier to convert the ac power of the LC resonator to the dc output.

The overall operation characteristics of the LLC converter can be simply analyzed by extracting the fundamental wave element from the voltage and current waveforms in the circuit operation and constructing the equivalent circuit model based on the fundamental wave element. Figure 2 illustrates the resultant equivalent circuit model. The half-bridge circuit is represented by the sinusoidal ac voltage source, whose root-mean-square value V_{ac_in} is expressed as

$$V_{ac\_in} = \frac{\sqrt{2}}{\pi} V_{in}, \tag{1}$$

where V_{in} is the dc input voltage. Meanwhile, the rectifier and the load resistance R_{load} are represented by equivalent resistor R_{ac_load}. The root-mean-square value V_{ac_out} of the output ac voltage, i.e. the ac voltage of R_{ac_load}, is expressed as

$$V_{ac\_out} = \frac{2\sqrt{2}}{\pi} V_{out}, \tag{2}$$

where V_{out} is the dc output voltage. Therefore, considering that R_{ac_load} must consume the same power as the load resistance R_{load} of the real LLC converter, R_{ac_load} can be expressed as

$$R_{ac\_load} = \frac{8}{\pi^2} R_{load}. \tag{3}$$

According to the circuit theory, V_{ac_out} can be calculated as

$$V_{ac\_out} = \frac{1}{N} \frac{g(\omega)}{j\omega(L_r + L_l) + \dfrac{1}{j\omega C_r} + g(\omega)} V_{ac\_in}, \tag{4}$$

where ω is the angular frequency and g(ω) is the function defined as

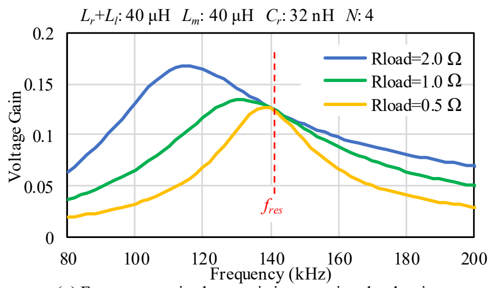
$$g(\omega) = \frac{j\omega L_m N^2 R_{ac\_load}}{j\omega L_m + N^2 R_{ac\_load}}. \tag{5}$$

Therefore, the voltage gain, i.e. the ratio V_{in}/V_{out}, can be obtained using (1), (2), and (4) as

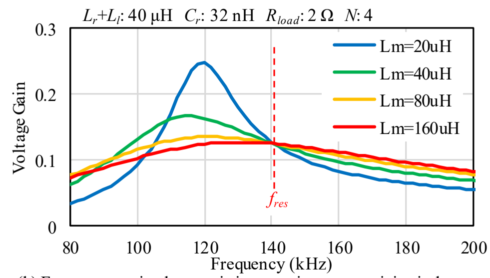
$$\frac{V_{out}}{V_{in}} = \frac{1}{2N} \frac{g(\omega)}{j\omega(L_r + L_l) + \dfrac{1}{j\omega C_r} + g(\omega)}. \tag{6}$$

As can be seen in (6), the prominent feature of the voltage gain is that the gain equals $1/2N$, which is independent of the load resistance R_{load}, when the operating frequency is set at the resonant frequency of the LC resonator of C_r and L_l, L_r. Figure 3(a) plots the theoretically calculated voltage gain as functions of the operating frequency at various load resistance. As can be seen in the figure, the voltage gain characteristics have a unique common point at the frequency identical to the resonant frequency f_{res} of the LC resonator of C_r and L_l, L_r.

Conventionally, the output voltage of the LLC converter is controlled by adjusting the operating frequency. In other words, the conventional control is designed to observe the small deviation of the output voltage from the output voltage command value and adjust the operating frequency to compensate for the deviation. Therefore, if the voltage gain at the operating condition varies according to the change of the load resistance, this change of the load current must cause the small fluctuation in the output voltage and this motivates the controller to adjust the operating frequency to compensate for the output voltage deviation. This implies that the less dependent on the load resistance the voltage gain characteristic is, the smaller output impedance the LLC converter will have. Consequently, the LLC converter is designed to be operated at f_{res} in the steady operation because this frequency is the unique condition in which the output



(a) Frequency-gain characteristics at various load resistance



(b) Frequency-gain characteristics at various magnetizing inductance

Fig. 3. Frequency-gain characteristics of LLC converter

voltage is not affected by the load resistance and therefore the operation at f_{res} exhibits the smallest output impedance.

This design requirement of the LLC converter with the conventional control however suggests that the deviation of the operating frequency from f_{res} deteriorates the output impedance, damaging the output voltage stability. This drawback is inconvenient for designing the LLC converter to cover a wide range of the output voltage and the load current.

Another drawback of this design requirement is that the transformer needs to have small magnetizing inductance L_m. Because the operating frequency at the steady-state is designed close to f_{res}, the voltage gain of the LLC converter needs to cover a wide range of voltage gain including $1/2N$, which is the voltage gain at f_{res}, for implementing a wide range of the output voltage or sufficiently fast output regulation. However, this does not accept the reactance of L_m, i.e. $j\omega L_m$, to have a far greater value than the equivalent output impedance $N^2R_{ac\_load}$. The reason is that $g(\omega)$ equals approximately to $N^2R_{ac\_load}$ if $j\omega L_m >> N^2R_{ac\_load}$ and therefore the voltage gain can scarcely take a value greater than $1/2N$. In fact, as seen in Fig. 3(b), large L_m narrows the possible range of the voltage gain. Therefore, the magnetizing inductance L_m should be designed to have the reactance of at most similar order as the smallest possible value of $N^2R_{ac\_load}$.

Accordingly, L_m is designed to have extremely small inductance, particularly if the output voltage or the output current has a large variation. However, this design damages the efficiency due to the large primary current at the light load condition. This drawback is also inconvenient for applying the LLC converter to the power supply that needs to cover a wide range of the output voltage gain and the load current.

As reviewed above, the aforementioned drawbacks are originated from the frequency characteristic of the output voltage. However, this may indicate that these drawbacks are caused by the conventional control approach, in which the operating frequency was adjusted to compensate for the deviation of the output voltage from the command value. Hence, this paper proposes to control the output voltage by adjusting another parameter for avoiding these drawbacks.

III. Proposed Control

The proposed control rather adjusts the power factor of the ac power supplied to the primary winding to regulate the output voltage instead of the operating frequency. As elucidated in the following discussion, this control approach enables the LLC converter to achieve the same excellent dynamic performance as the buck chopper, in which the duty cycle is used to regulate the output voltage. In the buck chopper operation, the output voltage characteristic on the duty cycle, i.e. the control parameter, is independent of the load resistance at any duty cycle and therefore the small output impedance can be obtained regardless of the operating condition.

To analyze the performance of the proposed control, this section firstly formulates the state-space model of the LLC converter controlled by adjusting the power factor based on the equivalent circuit shown in Fig. 2. Therefore, the voltage and current of this LLC converter are approximated as the sinusoidal waves for simplifying the analysis. The LLC converter with the proposed control should be designed to have large L_m, because the proposed control has small output impedance with large L_m regardless of the operating condition, as shown later. Therefore, in the following discussion, the L_m is approximated to be infinitely large for simplifying the analysis. Furthermore, the characteristic impedance of the LC resonator of C_r, L_r, and L_l is assumed to be far greater than $N^2R_{ac\_load}$ so that the operating frequency of the LLC converter is close to the resonance frequency of the resonator f_{res}. In other words, the Q factor Q_F of Fig. 2 is assumed to be far greater than 1, i.e.

$$Q_F = \frac{\sqrt{(L_r + L_l)/C_r}}{N^2 R_{ac\_load}} >> 1. \tag{7}$$

Let θ be the power factor angle between the ac voltage source of Fig. 2, which represents the output voltage of the half-bridge circuit of S1 and S2, and the transformer primary current. The symbols V_{ac_in} and I_1 denote the root-mean-square values of the ac voltage source and the primary current, respectively. Then, the effective power P_{in} supplied by the ac voltage source per unit time can be expressed as

$$P_{in} = V_{ac\_in} I_1 \cos\theta. \tag{8}$$

Similarly, the effective power P_{cap} output from the secondary winding per unit time can be expressed as

$$P_{cap} = V_{ac\_out} I_2, \tag{9}$$

where V_{ac_out} and I_2 are the root-mean-square values of the output ac voltage and the transformer secondary current.

Because L_m is assumed to take a large value, the energy stored in L_m is negligible compared to the energy stored in the LC resonator of C_r, L_r, and L_l, which is denoted by E_{res}. According to the assumption of (7), the LC resonator can be approximated to be driven at its resonance frequency. Noting that E_{res} is almost constant during the period of the ac voltage when driven at the LC resonant frequency and that the instantaneous ac voltage of C_r is zero at the moment of the peak current in L_r and L_l, E_{res} can be formulated as

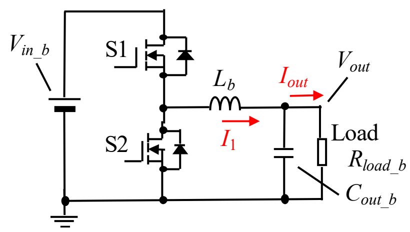


Fig. 4. Schematic diagram of buck chopper

$$E_{res} = \frac{1}{2}(L_r + L_l)\left(\sqrt{2}I_1\right)^2 = (L_r + L_l)I_1^2. \tag{10}$$

The difference between P_{in} and P_{cap}, i.e. P_{in}–P_{cap}, denotes the energy charged in the LC resonator per unit time. Therefore, from (8)–(10), an equation of I_1 can be obtained as

$$\begin{aligned}
&\frac{dE_{res}}{dt} = P_{in} - P_{cap} \\
&\therefore 2(L_r + L_l)\frac{dI_1}{dt} = V_{ac_in}\cos\theta - V_{ac_out}\frac{I_2}{I_1} \\
&\qquad\qquad = V_{ac_in}\cos\theta - NV_{ac_out}.
\end{aligned} \tag{11}$$

For deriving the rightmost equality, the relation I_2=NI_1 is used because large L_m is assumed. By utilizing (1) and (2), (11) can be rewritten as

$$2(L_r + L_l)\frac{dI_1}{dt} = \frac{\sqrt{2}}{\pi}V_{in}\cos\theta - \frac{2\sqrt{2}}{\pi}NV_{out}. \tag{12}$$

Next, V_{ac_out} is analyzed to derive an equation of V_{out}. The dc output power per unit time of the LLC converter, which is denoted as P_{out}, is expressed as

$$P_{out} = \frac{V_{out}^2}{R_{load}}. \tag{13}$$

Noting that P_{cap} indicates the ac power per unit time supplied to the rectifier, P_{cap}–P_{out} can be interpreted as the energy charged in the output smoothing capacitor per unit time. If E_{cap} denotes the total energy stored in the output smoothing capacitor, E_{cap} can be expressed as

$$E_{cap} = \frac{1}{2}C_{out}V_{out}^2. \tag{14}$$

Therefore, an equation of V_{out} can be obtained from (2), (9), (13), and (14) as

$$\begin{aligned}
&\frac{dE_{cap}}{dt} = P_{cap} - P_{out} \\
&\therefore C_{out}\frac{dV_{out}}{dt} = \frac{V_{ac_out}}{V_{out}}I_2 - \frac{V_{out}}{R_{load}} \\
&\qquad\qquad = \frac{2\sqrt{2}}{\pi}NI_1 - \frac{V_{out}}{R_{load}}.
\end{aligned} \tag{15}$$

Consequently, from (12) and (15), the state-space model of the LLC converter operated with the proposed control is constructed as

$$\begin{aligned}
\frac{d}{dt}\begin{pmatrix} I_1 \\ V_{out} \end{pmatrix} &= \begin{pmatrix} 0 & -\dfrac{\sqrt{2}N}{\pi(L_r + L_l)} \\ \dfrac{2\sqrt{2}N}{\pi C_{out}} & -\dfrac{1}{C_{out}R_{load}} \end{pmatrix}\begin{pmatrix} I_1 \\ V_{out} \end{pmatrix} \\
&\quad + \begin{pmatrix} \sqrt{2}N/\pi(L_r + L_l) \\ 0 \end{pmatrix}\frac{V_{in}}{2N}\cos\theta.
\end{aligned} \tag{16}$$

This resultant state-space model is fundamentally the same as that of the buck chopper. In fact, the state-space model of the buck chopper, depicted in Fig. 4, can be formulated as

$$\begin{aligned}
\frac{d}{dt}\begin{pmatrix} I_1 \\ V_{out} \end{pmatrix} &= \begin{pmatrix} 0 & -\dfrac{1}{L_b} \\ \dfrac{1}{C_{out_b}} & -\dfrac{1}{C_{out_b}R_{load_b}} \end{pmatrix}\begin{pmatrix} I_1 \\ V_{out} \end{pmatrix} \\
&\quad + \begin{pmatrix} 1/L_b \\ 0 \end{pmatrix}V_{in_b}D,
\end{aligned} \tag{17}$$

where D is the duty cycle of the switch S1. Therefore, the two models, i.e. (16) and (17), exhibit accurately the same performance if the circuit parameters are set according to

$$\begin{aligned}
&C_{out_b} = \frac{\pi}{2\sqrt{2}N}C_{out}, \quad R_{load_b} = \frac{2\sqrt{2}N}{\pi}R_{load} \\
&L_b = \frac{\pi(L_r + L_l)}{\sqrt{2}N}, \quad V_{in_b} = \frac{V_{in}}{2N}, \quad D = \cos\theta.
\end{aligned} \tag{18}$$

It is worth noticing that the power factor of the LLC converter corresponds to the duty cycle of the buck chopper. The proposed control adjusts the power factor to regulate the output power in the same fashion as the buck chopper. Therefore, the LLC converter with the proposed control has the same dynamic performance as the buck chopper. The proposed control can hence utilize the existing design know-hows for the buck choppers. For example, sophisticated compensation circuits of the buck chopper, such as the 3-pole-2-zero compensator [18]–[20], can be also applied and designed for the proposed control in the similar fashion as the buck chopper.

As is similar to the buck chopper, the proposed control has attractive merit that the output voltage gain is dependent only on the power factor and independent of the load resistance at any output voltage. Furthermore, the LLC converter for the proposed control is designed to have large magnetizing inductance L_M, which can avoid the excessive primary current under light load conditions. Consequently, the proposed control can be expected to mitigate the aforementioned problems of the conventional control, thus possibly improving the output voltage stability and increasing the efficiency.

IV. Circuit Implementation of Proposed Control

The proposed control adjusts the power factor of the ac power supplied to the transformer primary winding according

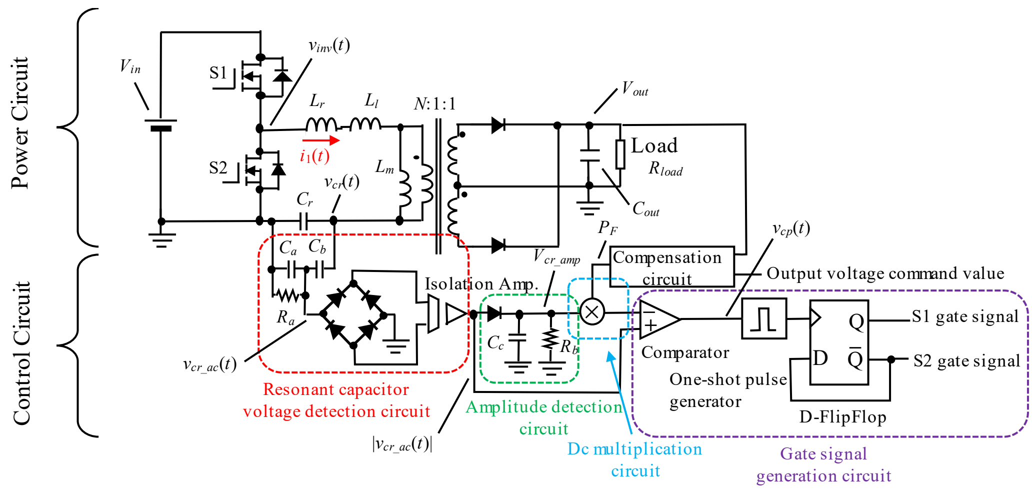


Fig. 5. LLC converter with proposed control circuit

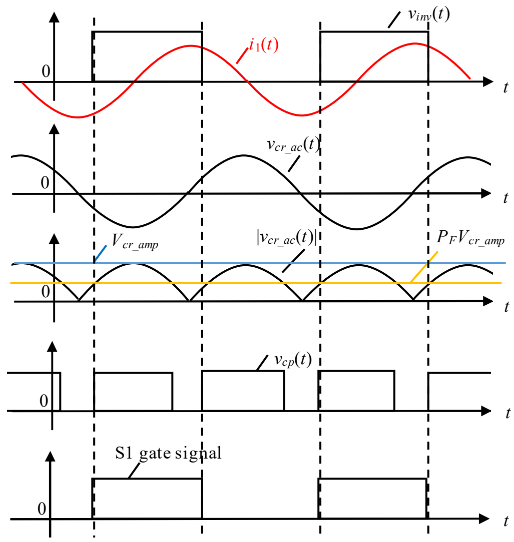


Fig. 6. Operating waveforms of LLC converter with proposed control circuit

to the output voltage deviation from the command value. Therefore, practical implementation of the proposed control needs a control circuit that can set the power factor at the required value. However, to the best of the authors' knowledge, such a circuit has never been proposed. Therefore, this section proposes a control circuit of the power factor.

Figure 5 illustrates the schematic diagram of the LLC converter with the proposed control circuit of the power factor. Figure 6 shows the operating waveforms. The operating principles of the proposed control circuit is described hereafter.

A. Switching Timing for Achieve Required Power Factor

The overall operation of the LLC converter is commonly analyzed based on the equivalent circuit of Fig. 2, in which the voltage and current waveforms are approximated to be sinusoidal. However, the voltage and current waveforms in the actual LLC converter are known to be significantly deformed from the sinusoidal waves. Therefore, the power factor angle cannot be simply determined from the time difference of the zero-crossing timing between the voltage and the current of the primary winding. For this reason, the proposed control circuit does not estimate the power factor P_F by calculating the power factor angle θ from the zero-crossing timing and determine P_F according to P_F=cosθ.

Instead of the zero-crossing timing, the proposed control circuit estimates the power factor by the instantaneous voltage of the resonant capacitor C_r. Let $v_{inv}(t)$ and $i_1(t)$ be the instantaneous voltage outputted from the half-bridge circuit and the instantaneous transformer primary current. The effective power supplied to the transformer primary winding per unit time, which is denoted as P_{in}, can be expressed as

$$P_{in} = \frac{1}{T}\int_0^T v_{inv}(t) i_1(t) dt, \tag{19}$$

where T is the switching period of the half-bridge circuit. As is natural for the half-bridge circuit of the LLC converter, $v_{inv}(t)$ has the following waveform:

$$v_{inv}(t) = \begin{cases} V_{in} & (0 \le t < T/2), \\ 0 & (T/2 \le t < T). \end{cases} \tag{20}$$

Therefore, substituting (20) into (19) yields

$$P_{in} = \frac{V_{in}}{T}\int_0^{T/2} i_1(t) dt = \frac{V_{in}}{T} C_r \{v_{cr}(T/2) - v_{cr}(0)\}, \tag{21}$$

where $v_{cr}(t)$ is the instantaneous voltage of the resonant capacitor. The resonant capacitor voltage $v_{cr}(t)$ is the sum of the dc component, which is identical to $V_{in}/2$, and the ac component, which is denoted as $v_{cr\_ac}(t)$. Because of the symmetry of the ac component of $v_{inv}(t)$ between the former half switching period and the latter half period, $v_{cr\_ac}(t)$ bas the following relation:

$$v_{cr\_ac}(T/2) = -v_{cr\_ac}(0). \tag{22}$$

Therefore, substituting (22) into (21) yields

$$\begin{aligned} P_{in} &= \frac{V_{in}}{T} C_r \{v_{cr\_ac}(T/2) - v_{cr\_ac}(0)\} \\ &= -2V_{in} C_r \frac{v_{cr\_ac}(0)}{T}. \end{aligned} \tag{23}$$

Now, we consider to imaginary turn on and off the switch S1 at the time when $v_{cr\_ac}(t)$ takes the minimum and maximum values, respectively. Noting that the time from the minimum $v_{cr\_ac}(t)$ to the maximum $v_{cr\_ac}(t)$ equals $T/2$, this is a possible operation of the LLC converter and will maximize P_{in}. Therefore, this maximum P_{in} can be regarded as the apparent power S_{in}. Hence, S_{in} is obtained as

$$S_{in} = 2V_{in}C_r \frac{V_{cr\_amp}}{T}, \tag{24}$$

where V_{cr_amp} is the peak value of $v_{cr\_ac}(t)$. Because the power factor P_F is the ratio of P_{in} to S_{in}, P_F can be calculated as

$$P_F = \cos\theta = \frac{P_{in}}{S_{in}} = \frac{-v_{cr\_ac}(0)}{V_{cr\_amp}}. \tag{25}$$

The above discussion describes the estimation method of the power factor for the given switching pattern of S1 and S2. Contrarily, this discussion can also be used to determine the switching timing of S1 and S2 for achieving the required power factor. According to the above discussion, $v_{cr\_ac}(0)$ and $v_{cr\_ac}(T/2)$ are identical to $-P_F V_{cr\_amp}$ and $P_F V_{cr\_amp}$. Therefore, to achieve required P_F, S1 should be turned on and off when $v_{cr\_ac}(t)$ coincides with $-P_F V_{cr\_amp}$ and $P_F V_{cr\_amp}$, respectively.

B. Operating Principles

The proposed control circuit generates the switching gate signals to achieve the given power factor P_F. The proposed circuit comprises four subcircuits: 1. resonant capacitor voltage detection circuit, 2. amplitude detection circuit, 3. dc multiplication circuit, and 4. gate signal generation circuit.

The resonant capacitor voltage detection circuit eliminates the dc component and attenuates the resonant capacitor voltage by dividing the voltage by two small capacitors C_a and C_b. The output of this voltage divider is further conducted to the rectifying circuit. Consequently, the output of this circuit corresponds to $|v_{cr\_ac}(t)|$.

The signal of $|v_{cr\_ac}(t)|$ is conducted to the amplitude detection circuit. In Figure 5, the amplitude detection circuit is made as a simple rectifying circuit with a low-pass filter. The output of this circuit corresponds to V_{cr_amp}.

The output signal of the amplitude detection circuit is conducted to the multiplication circuit. In this circuit, the signal is multiplied by the dc voltage representing the power factor command value, which is supplied by the compensation circuit. The output of this circuit corresponds to $P_F V_{cr\_amp}$.

The compensation circuit generates the power factor command value according to the deviation of the output voltage from the output voltage command value. The same compensation circuits as for the buck choppers can be utilized for the proposed control, as discussed in the previous section.

Finally, the output signals of the resonant capacitor voltage detection circuit and the multiplication circuit are conducted to the gate signal generation circuit. The gate signal generation circuit firstly compares the two signals using a comparator to detect the timing at which $v_{cr\_ac}(t)$ coincides with $-P_F V_{cr\_amp}$ or $P_F V_{cr\_amp}$. The output of the comparator is conducted to a D-FF finally to generate the switching gate signals of S1 and S2.

TABLE I. SIMULATION PARAMETERS FOR SUBSECTION V.A

LLC Converter with proposed control		Buck chopper	
V_{in}	100 V	V_{in_b}	25 V
R_{load}	0.25Ω	R_{load_b}	0.45Ω
C_{out}	200µF	C_{out_b}	111µF
L_r+L_l	10µH	L_b	11.1µH
L_m	100µH		
C_r	5nF		
C_a	1µF		
C_b	5nF		
R_a	1kΩ		
C_c	0.1µF		
R_b	1kΩ		

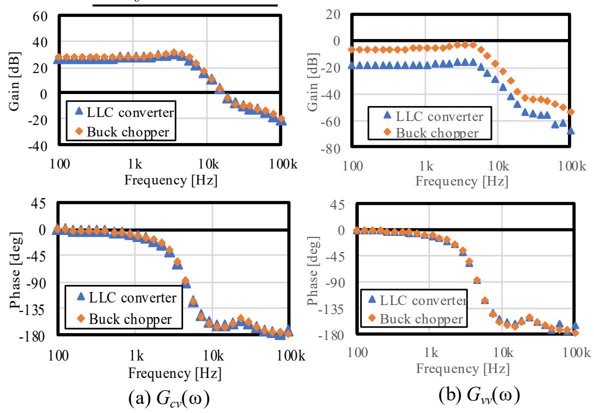


Fig. 7. Simulation results of dynamic characteristics of LLC converter with proposed control circuit and buck chopper. (a) Command value fluctuation vs. output voltage characteristic $G_{cv}(\omega)$. (b) Input voltage fluctuation vs. output voltage characteristic $G_{vv}(\omega)$.

V. SIMULATION

The simulation was carried out to verify the performance of the LLC converter with the proposed control. This section firstly compares the simulated bode plots of the various dynamic characteristics between the LLC converter with the proposed control and the buck chopper to confirm that these two power converters have the same dynamic performance, as predicted by the theory. Then, this section compares the output impedance of the LLC converter, as well as the root-mean-square value of the transformer primary current, between the proposed control and the conventional frequency control to confirm the effectiveness of the proposed control.

For this purpose, the simulation models of the LLC converters with the proposed control and the conventional control, as well as the buck chopper, were constructed in the model space of the circuit simulator. The software used in this simulation was PSIM (Powersim Inc.)

A. Comparison with Buck Chopper

The dynamics characteristics were compared between the simulation circuits of the LLC converter with the proposed control and the buck chopper. The circuit shown in Fig. 5 was utilized for the LLC converter with the proposed control. (The compensator was eliminated. Instead, a fixed power factor value was input to the multiplier circuit.) The circuit shown in Fig. 4 was utilized for the buck chopper. The power circuits of these converters were designed to theoretically have the same dynamic characteristic according to (18). Consequently, the circuit parameters were set as listed in Table I.

In this simulation, the power factor or the duty cycle was set constant at 0.5. Then, the bodes plots of the command

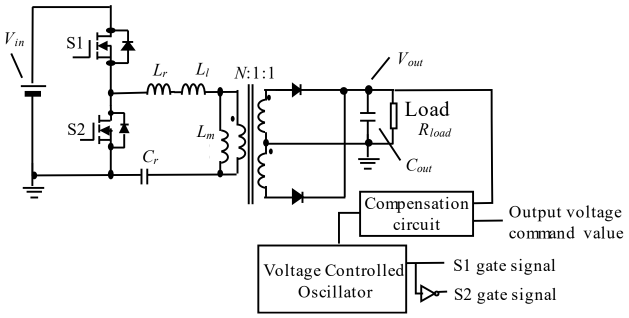


Fig. 8. Similation circuit of LLC converter with conventional control

TABLE II. SIMUATION PARAMETERS FOR SUBSECTION V.B

Proposed control				Conventional control			
V_{in}	100 V	C_c	0.1μF	V_{in}	100 V	N	4
C_{out}	200μF	R_b	1kΩ	C_{out}	200μF	A_1	2.50×10^{-8}
L_r+L_l	10μH	N	2	L_r+L_l	40μH	A_2	8.00×10^{-4}
L_m	100μH	A_1	9.00×10^{-7}	L_m	2.1μH	A_3	4.50
C_r	5nF	A_2	0.0278	C_r	2.5nF	A_4	1.60×10^{-13}
C_a	1μF	A_3	220	VCO $Gain$	1MHz/V	A_5	7.00×10^{-6}
C_b	5nF	A_4	5.89×10^{-11}	VCO $Range$	400kHz-700kHz		
R_a	1kΩ	A_5	3.86×10^{-5}				

value, i.e. power factor or duty cycle, fluctuation vs. the output voltage characteristic $G_{cv}(\omega)$ and the input voltage fluctuation vs. the output voltage characteristic $G_{vv}(\omega)$ were calculated.

Figure 7 shows the simulation results. The LLC converter with the proposed control exhibited almost the same characteristics in $G_{cv}(\omega)$ and $G_{vv}(\omega)$. Certainly, the constant gap was found between the two models in $G_{vv}(\omega)$. However, this gap is originated from the natural difference of the voltage gain: the LLC converter with the proposed control has the voltage gain 1/4 time as small as the buck converter. Therefore, the simulation results supported the theoretically predicted dynamical equivalence between the LLC converter with the proposed control and the buck converter.

B. Comparison with Conventional Control

Next, the output impedance, as well as the root-mean-square value of the primary current, was compared between the LLC converters with the proposed control and the conventional control. The circuit shown in Fig. 5 and Fig. 8 were utilized for the LLC converter with the proposed and conventional control including the compensation circuit. The circuit parameters are listed in Table II.

Because the LLC converter with the proposed control is equivalent to the buck chopper, the nominal power factor was set close to 0.5, which results in the output voltage close to $0.5V_{inv}/2N$. On the other hand, the conventional control is commonly designed to be operated at the operating frequency close to f_{res}, which results in the output voltage close to $V_{inv}/2N$. Hence, the turn ratio N of the proposed control was set at the half value of that of the conventional control. As a result of this difference, we designed L_r+L_l and C_r of the proposed control to be 1/4 time and 4 time, respectively, as large as those of the conventional control because the load resistance seen from the primary side is 1/4 times as large in the proposed control as that in the conventional control and therefore the characteristic impedance of the LC resonator in the proposed control should be set 1/4 of that of the conventional control to have the same quality factor of the resonator.

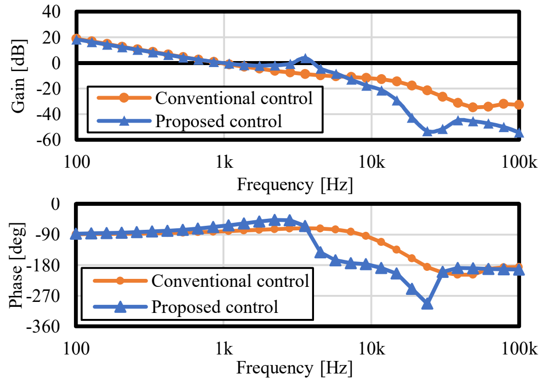


Fig. 9. Loop transfer function of LLC conveters with proposed and conventional controls at V_{out}=12.5 V and R_{load}=1 Ω

The transformer of the LLC converter with the proposed control was designed to have a sufficiently large magnetizing inductance than the leakage inductance. Meanwhile, the transformer of the LLC converter with the conventional control was designed to cover the same voltage gain range as the LLC converter with the proposed control, i.e. V_{in}/V_{out}=0~1/4, at the nominal load resistance 1 Ω.

Both of these two simulation circuits adopted the 3-pole-2-zero compensation circuit, which has the transfer function $T(s)$ expressed as

$$T(s) = \frac{A_1 s^2 + A_2 s + A_3}{A_4 s^3 + A_5 s^2 + s}, \qquad (26)$$

where s is the complex variable, A_1–A_5 are the constants characterizing the transfer function. However, the two simulation circuits adopted different values for A_1–A_5 so that the loop transfer function $G_{ov}(\omega)$ of these LLC converters have the same cross-over frequency of 1 kHz, the same gain margin of 20 dB, and the same gain of 20 dB at 100 Hz when the nominal output voltage is 12.5 V and the nominal load resistance is 1 Ω. The values of A_1–A_5 are listed in Table II. The resultant $G_{ov}(\omega)$ at V_{out}=12.5 V and R_{load}=1 Ω is presented in Fig. 9.

Figure 10 shows the simulation result of the output impedance at V_{out}=12.5 V, 9 V, and 6 V with R_{load}=1 Ω, respectively. The results revealed that the proposed control exhibited the smallest output impedance than the conventional control at the low-frequency region regardless of the output voltage. As the low-frequency fluctuation of the load current tends to have large amplitude and therefore be the main contributor of the output voltage fluctuation in common applications, these results indicate that the proposed control can offer excellent output voltage stabilization compared to the conventional control in a wide range of the output voltage.

The proposed control showed far better output voltage stabilization also at V_{out}=12.5 V, although the output voltage gain is not dependent on the output current in both of the controls at this output voltage. Therefore, the improvement of the output voltage by the proposed control cannot be entirely attributed to the difference in the output voltage gain vs. the output current characteristics, which was anticipated by this paper. Hence, this paper cannot give a complete explanation for the mechanism of this improvement, although future research will investigate the reason for this improvement.

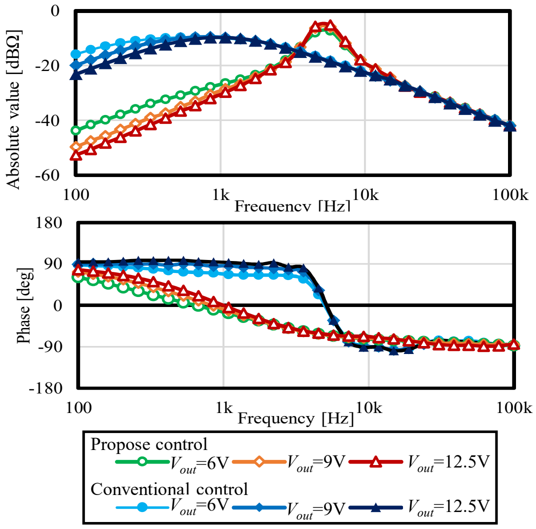


Fig. 10. Output impedance of LLC conveters with proposed and conventional controls with R_{load}=1 Ω at V_{out}=12.5 V, 9 V, and 6 V.

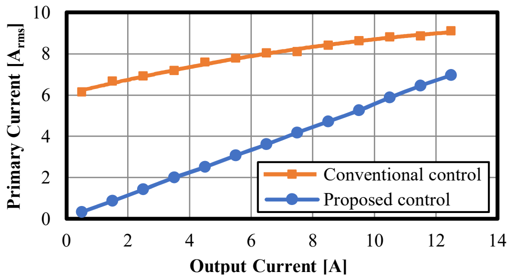


Fig. 11. Simulation result of the root-mean-square values of the transformer primary current of LLC conveters operated at V_{out}=12.5 V.

Figure 11 shows the simulation result of the transformer primary current at V_{out}=12.5 V. The result indicated that the proposed control can effectively reduce the primary current particularly under light load condition, which implies a possible improvement in the efficiency at the light load. Consequently, the simulation supported the effectiveness of the proposed control.

VI. Conclusions

The LLC design covering a wide range in the output voltage gain and the load current can deteriorate the output voltage stability and efficiency. For mitigating this problem, this paper proposed a novel control method that adjusts the power factor rather than the operating frequency. This paper elucidated that the dynamic performance of the LLC converter with the proposed control is equivalent to the buck chopper. Therefore, the proposed control can offer excellent output voltage regulation regardless of the output voltage gain and the load current, thus possibly improving the output voltage stability than the conventional control. Furthermore, this paper proposed a practical control circuit to implement the proposed control. The simulation verified the improvement of the output voltage stability and reduction in the primary current, which is the cause of the efficiency deterioration, suggesting the promise of the proposed control.

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