Wide Voltage Resonant Converter Using a Variable Winding Turns Ratio

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Abstract: This paper presents a inductor–inductor–capacitor (LLC) resonant converter with variable winding turns to achieve wide voltage operation (100–400 V) and realize soft switching operation over the entire load range. Resonant converters have been developed for consumer power units in computers, power servers, medical equipment, and adaptors due to the advantages of less switching loss and better circuit efficiency. The main disadvantages of the LLC resonant converter are narrow voltage range operation owing to wide switching frequency variation and limited voltage gain. For computer power supplies with hold-up time function, electric vehicle battery chargers, and for power conversion in solar panels, wide input voltage or wide output voltage operation capability is normally demanded for powered electronics. To meet these requirements, the variable winding turns are used in the presented circuit to achieve high- or low-voltage gain when $V_{in}$ is at low- or high-voltage, respectively. Therefore, the wide voltage operation capability can be implemented in the presented resonant circuit. The variable winding turns are controlled by an alternating current (AC) power switch with two back-to-back metal-oxide-semiconductor field-effect transistors (MOSFETs). A 500-W prototype is implemented and test results are presented to confirm the converter performance.

Keywords: pulse frequency control; DC converters; wide voltage operation; zero voltage switching

1. Introduction

Renewable energy systems with electronic power techniques [1–4] have been widely researched to lessen the demands for fossil fuels and reduce the effects of global warming. Form the various renewable energy sources, solar and wind power are more attractive out of the clean energy sources. The main problem with solar and wind power is the non-constant output voltage from wind turbine generators and photovoltaic (PV) solar panels. The output voltage of a wind turbine generator is related to wind speed and the output voltage of a PV solar panel is related to solar intensity. High-efficiency direct current (DC) power converters [5–8] have been proposed for power units for computers, battery dischargers and chargers, telecommunication systems, and server systems. Out of the various soft switching techniques, the active clamp technique [9,10], asymmetric pulse width modulation (PWM) [11,12], and phase-shift PWM converters [13,14] are the most attractive circuit topologies with the lowest switching losses. The PWM scheme is widely used to control these circuit topologies with constant output voltage. However, the turn-on time of the PWM scheme is dependent on the input voltage. PWM converters have a low (high) duty cycle at high (low) voltage input. However, the low duty cycle results in a high root mean square current and high conduction loss. Thus, the PWM converters have low (high) circuit efficiency at high (low) input voltage conditions. This drawback limits PWM converters in terms of wide voltage operation. Pulse frequency modulation (PFM), or variable frequency control, has been adopted in resonant converters [15]. In many resonant circuit topologies, the inductor–inductor–capacitor (LLC) resonant converter is a more attractive circuit.
topology for use in high-input voltage inputs. The main advantages of LLC resonant converters [16–18] are low switching loss in power devices, high efficiency, and low electromagnetic interference (EMI). The main drawback of LLC resonant converters is the limited voltage gain. This means it is difficult to operate the resonant converter in wide voltage operation. To realize wide voltage operation, two-stage converters and series–parallel connection for full-bridge or half-bridge converters [19–22] have been proposed for battery charger, solar power, and wind power applications. The two-stage converters have the drawback of low efficiency, while the series- or parallel-connected converters have more circuit components, which increase the conduction loss and reduce the converter reliability. In [23,24], two full-bridge LLC converters with primary-parallel–secondary-series connection were proposed to achieve wide input or output voltage operation. However, eight power switches and eight diodes are used in this circuit topology. The conduction losses in power semiconductors and the cost are increased. An interleaved LLC resonant converter has been studied and implemented in [25] to realize wide input voltage operation and ripple-free input current. However, this circuit topology is a two-stage circuit structure, meaning that the circuit efficiency is low in full load conditions, when both the boost converter and LLC converter are operated. In [26], a half-bridge LLC resonant converter with two sets of center-tapped rectifiers was proposed to extend the hold-up time problem for personal computers. However, four winding turns are used on the secondary side in this circuit topology, and the copper losses on magnetic windings are increased.

An LLC resonant converter with variable winding turns is proposed and investigated to achieve soft switching operation on active switches over the entire load range and for wide voltage range operation ($V_{in} = 100–400$ V). The variable winding turns controlled by an alternating current (AC) switch are adopted to achieve different voltage gains based on low- or high-input voltage ranges. During the low-input voltage range (from $V_{in} = 100$ V to 200 V), the converter with more secondary turns is selected to achieve high-voltage gain for the LLC converter. During the input voltage range ($V_{in} = 200–400$ V), the converter with fewer secondary turns is operated to achieve low-voltage gain for the LLC converter. Therefore, the proposed converter can be operated under wide input voltage variation, such as for solar power or switching mode power supplies without power factor correction, under the universal AC input voltage. The LLC resonant circuit with pulse-frequency modulation is used to control active devices. Due to the inductive impedance load of the LLC resonant tank, the converter can realize soft switching operation for both active switches and rectifier diodes. The voltage double rectifier topology is adopted on the output side to limit the voltage rating of the secondary diodes at $V_o$ instead of $2V_o$ in center-tapped rectifiers. The rectifier diodes with a low-voltage rating have low-voltage drop. Comparing the center-tapped rectifier structure, the voltage double rectifier structure has less conduction losses for rectifier diodes and secondary windings. Comparing the conventional two-stage power converters, parallel-series connected converters, and the other resonant converters in [19–26], the presented LLC resonant converter has a simple circuit structure, a simple control algorithm, and requires fewer power components to achieve wide voltage operation. The description of the presented circuit schematic and operation principle is presented in Section 2. The circuit analysis and design procedures of the presented resonant converter are provided in Section 3. Experiments are demonstrated and discussed to confirm the effectiveness of the presented converter in Section 4. Finally, the conclusions are discussed in Section 5.

2. Proposed Converter and Operation Principle

Figure 1a shows the conventional LLC converter. The half-bridge leg and center-tapped rectifier are used on the primary and secondary sides, respectively, for high-voltage input and low-voltage output applications. According to the center-tapped rectifier topology, the secondary diodes have $2V_o$ voltage stress. Figure 1b shows the circuit topology with the voltage double rectifier for medium- or high-voltage output applications. The advantages of this circuit topology are the low-voltage rating of the rectifier diodes ($V_{D1,rating} = V_{D1,rating} = V_o$) and fewer secondary winding turns ($n_s$ instead of $2n_s$)
in the center-tapped rectifier structure. Therefore, the voltage double rectifier structure has less copper loss on the secondary windings.

![Diagram](a)

![Diagram](b)

**Figure 1.** Conventional inductor–inductor–capacitor (LLC) converter based on (a) a center-tapped rectifier structure and (b) a voltage double rectifier structure.

In order to improve the drawback of the limit voltage range in conventional LLC converters, Figure 2a shows the circuit configuration of the presented LLC converter with variable winding turns to achieve wide input voltage operation (100–400 V). The input voltage of the proposed LLC converter can be from the output of the solar panels, wind turbine DC generators, or battery banks. The output voltages of battery banks and large solar panels vary widely. A half-bridge LLC resonant tank is used on the primary side for soft switching operation in active devices $S_1$ and $S_2$ over the entire load range. The voltage rating of the active devices equals the maximum input voltage. Two sets of voltage double rectifiers and one AC switch $S_{ac}$ implemented with two metal-oxide-semiconductor field-effect transistors (MOSFETs) back-to-back connection are used on the output-side. If the proposed converter is operated under low-voltage input $V_{in,L}$ (100–200 V), then $S_{ac}$ is turned on (Figure 2b). Due to $S_{ac}$ is conducting, the $2n_p$ secondary winding turns are connected to load voltage. One observes that $D_3$ and $D_4$ are off. The transformer turns-ratio under low-input voltage operation is $n_p/(2n_p)$. The proposed converter has DC voltage gain $V_{dc}/V_{in,L} = G(f)(2n_p)/n_p$, where $V_{in,L}$ and $G(f)$ are low-input voltage range and the voltage transfer function of LLC converter. When the circuit is worked under high-voltage range $V_{in,H}$ (200–400 V), $S_{ac}$ is controlled at off state (Figure 2c). The $n_p$ winding turns are connected to load voltage and the diodes $D_1$ and $D_2$ are off. Therefore, the transformer turns-ratio for high-voltage operation is equal to $n_p/n_p$. The direct current voltage gain $V_{dc}/V_{in,H}$ of the LLC converter is $G(f)n_p/n_p$, where $V_{in,H}$ is high-input voltage range. From the previous discussion, one can observe that $V_{dc}/V_{in,L} = G(f)(2n_p)/n_p = 2V_{dc}/V_{in,H}$. It is concluded that $V_{in,H} = 2V_{in,L}$, such as $V_{in,H} = 200–400$ V and $V_{in,L} = 100–200$ V. Therefore, the presented resonant converter has high (low) DC voltage gain operated at low (high) input voltage range with low (high) transformer turns-ratio. The PWM signals of $S_1$ and $S_2$ are controlled with pulse-frequency modulation scheme shown in Figures 3a and 4a and the AC switch $S_{ac}$ is controlled by the Schmitt voltage comparator. The reference voltage of the Schmitt voltage comparator is 200 V and the hysteresis voltage band is 10 V.
According to the switching states of (series resonant frequency) \( f \) is less than the frequency \( \text{the resonant frequency is} \), the circuit goes to mode 2 at time \( t \). The magnetizing voltage \( v \) is increased in this mode. In mode 1, the series resonant frequency is \( \text{less than the frequency} \). It is clear that the resonant frequency \( \text{in mode 2} \), \( f_{\text{sw}} \) (switching frequency).

**Figure 2.** Circuit structure of (a) the presented converter (b) under low-voltage operation (\( S_{ac} \) on) and (c) under high-voltage operation (\( S_{sc} \) off).

### 2.1. Low-Voltage Operation \( V_{in,L} = V_{in,min} - 2V_{in,min} \) (\( S_{ac} \): on, \( D_3, D_4 \): off)

If the converter is operated at the low-voltage input from \( V_{in,min} \) to \( 2V_{in,min} \), then \( S_{ac} \) is turned on and \( D_3 \) and \( D_4 \) are turned off. The proposed converter has a DC voltage gain of \( V_o/V_{in,L} = G(f)(2n_s)/n_p \). According to the converter states of \( S_1, S_2, D_1, \) and \( D_2 \), the converter has six operating modes for each switching cycle. The PWM waveforms are demonstrated in Figure 3a. Figure 3b–g gives the equivalent circuits of six operating modes under \( f_{r1} \) (series resonant frequency) > \( f_{sw} \) (switching frequency).

Mode 1 \([t_0 \sim t_1]\): This mode begins at \( t_0 \) if \( v_{CS1} = 0 \) V and \( v_{CS2} = V_{in} \). Owing to \( i_{DS}(t_0) < 0 \) and \( i_{Lr} > i_{Lm} \), the body diode \( D_{S1} \) conducts and \( D_1 \) is forward-biased. Switch \( S_1 \) turns on after \( t_0 \) to achieve zero voltage switching. The magnetizing voltage \( v_{Lm} = n_p V_{Co1}/(2n_s) = n_p V_o/(4n_s) \) and the magnetizing current \( i_{Lm} \) increase in this mode. In mode 1, the series resonant frequency is \( f_{r1} = 1/2\pi \sqrt{L_r C_r} \). If \( f_{r1} \) (series resonant frequency) < \( f_{sw} \) (switching frequency), then the circuit goes to mode 3. Otherwise, the circuit goes to mode 2 at time \( t_1 \).

Mode 2 \([t_1 \sim t_2]\): At \( t_1 \), \( i_{D1} = 0 \) and \( D_1 \) turns off without the reverse recovery current. In mode 2, the resonant frequency is \( f_{r2} = 1/2\pi \sqrt{(L_r + L_m) C_r} \). It is clear that the resonant frequency \( f_{r2} \) in mode 2 is less than the frequency \( f_{r1} \) in mode 1. The peak-to-peak current of \( L_m \) is approximately obtained as

\[
\Delta i_{Lm} = \frac{n_p V_o}{8n_s L_m f_{sw}} \tag{1}
\]
Thus, the magnetizing current at time $t_2$ is given as

$$i_{Lm}(t_2) = \frac{\Delta i_{Lm}}{2} \approx \frac{n_P V_o}{16n_iL_m f_{sw}}$$  \hspace{1cm} (2)

Mode 3 [$t_2 \sim t_3$]: This mode starts at $t_2$ when switch S1 is turned off. Owing to $i_{Lr}(t_2) > 0$ and $i_{Lr} < i_{Lm}$, $C_{S2}$ ($C_{S1}$) is discharged (charged) from $V_{in}$ (0 V) and D2 is forward-biased. If the inductor energy $E_{Lr+Lm}$ at time $t_2$ is larger than the capacitor energy $E_{C_{S1}+C_{S2}}$, then $v_{C2}$ will be decreased to zero voltage at time $t_3$. The zero voltage switching condition of $S_2$ is expressed as

$$i_{Lm}(t_2) \geq V_{in} \sqrt{2/C_{oss} / (L_m + L_r)}$$  \hspace{1cm} (3)

where $C_{oss} = C_{S1} = C_{S2}$. The dead time ($t_d$) between $S_1$ and $S_2$ must be greater than the discharge time of $C_{S2}$ to ensure the zero voltage turn-on of $S_2$. Thus, the maximum value of $L_m$ can be calculated as

$$L_m \leq \frac{n_P V_o f_{d}}{32 V_{in} f_s n_s C_{oss}}$$  \hspace{1cm} (4)

Mode 4 [$t_3 \sim t_4$]: At time $t_3$, $v_{CS2} = 0$ V and $v_{CS1} = V_{in}$. Owing to $i_{Lr}(t_3) > 0$ and $i_{Lr} < i_{Lm}$, the body diode $D_{S2}$ conducts and $D_2$ is forward-biased. $S_2$ can turn on after $t_3$ with soft switching operation. In mode 4, $v_{Lm} = -n_P V_{C2}/(2n_i) = -n_P V_o/(4n_i)$ and $i_{Lm}$ decreases. The resonant frequency in mode 4 is $f_{r_1}$. If $f_{sw} < f_{r_1}$, then $t_{D2} = 0$ at time $t_4$.

Mode 5 [$t_4 \sim t_5$]: At $t_4$, $i_{D2} = 0$ and $D_2$ turns off without the reverse recovery current. $L_r$, $I_{m}$, and $C_r$ are naturally resonant with the frequency $f_{r_2}$.

Mode 6 [$t_5 \sim T_{sw} + t_0$]: At $t_5$, $S_2$ is turned off. Owing to $i_{Lr}(t_5) < 0$ and $i_{Lr} > i_{Lm}$, $C_{S1}$ is discharged and $D_1$ is forward-biased. The soft switching turn-on condition of $S_1$ is expressed as

$$|i_{Lm}(t_5)| = i_{Lm}(t_2) \geq V_{in} \sqrt{2/C_{oss} / (L_m + L_r)}$$  \hspace{1cm} (5)

At time $t_0 + T_{sw}$, $v_{CS1} = 0$ V.

Figure 3. Cont.
Figure 3. Cont.
The resonant frequency in mode 5 equals \( f \). The resonant frequency in mode 2 is \( f \).

If the converter is worked at the high-voltage input from \( V_{in,H} \) to \( 4V_{in,min} \), then \( S_{ac} \) turns off, and \( D_1 \) and \( D_2 \) are also off. The DC voltage gain becomes \( V_o/V_{in,H} = G(n_s/n_p) \). According to the on-off states of \( S_1, S_2, D_3 \), and \( D_4 \), the LLC converter has six operating modes for each switching period. Figure 4 gives the PWM waveforms and equivalent circuits under high-input voltage operation.

Mode 1 \([t_0 \sim t_1]\): At time \( t_0 \), \( v_{CS1} = 0 \) V. Since \( i_{Lr} > i_{LM} \) and \( i_{Lr}(t_0) < 0 \), \( D_1 \) is forward-biased and \( D_3 \) conducts. Thus, the zero voltage turn-on of \( S_1 \) can be realized after \( t_0 \). Due to \( D_3 \) being forward-biased, one can obtain \( \Delta v_{LM} = n_p V_{Co1}/n_s = n_p V_o/(2n_s) \) and \( i_{LM} \) increases. The resonant frequency in mode 1 is \( f_1 = 1/2\pi \sqrt{L_r C_r} \).

Mode 2 \([t_1 \sim t_2]\): At time \( t_1 \), \( i_D = 0 \) and \( D_2 \) turns off at \( t_1 \) without the reverse recovery current. The resonant frequency in mode 2 is \( f_2 = 1/2\pi \sqrt{(L_r + L_m)}/C_r \). The peak-to-peak ripple current \( \Delta L_m \) and the magnetizing current at time \( t_2 \) are approximately calculated as

\[
\Delta i_{LM} = \frac{n_p V_o}{4 n_s L_m f_{sw}}
\]

(6)

\[
i_{LM}(t_2) = \frac{\Delta i_{LM}}{2} = \frac{n_p V_o}{8 n_s L_m f_{sw}}
\]

(7)

Mode 3 \([t_2 \sim t_3]\): At \( t_2 \), \( S_1 \) turns off. Due to \( i_{Lr} < i_{LM} \) and \( i_{Lr}(t_2) > 0 \), \( D_4 \) conducts and \( C_{S2} \) is discharged from \( V_m \). If the inductor energy on \( L_r \) and \( L_m \) is larger than the capacitor energy on \( C_{S1} \) and \( C_{S2} \), then \( v_{CS2} \) is decreased to zero voltage at \( t_3 \). The zero voltage turn-on condition of \( S_2 \) is expressed as

\[
i_{LM}(t_2) = \frac{n_p V_o}{8 n_s L_m f_{sw}} \geq \frac{V_m}{\sqrt{2C_{oss}/(L_m + L_r)}}
\]

(8)

If the dead time \((t_d)\) between \( S_1 \) and \( S_2 \) is given, the maximum value of \( L_m \) under the high-input voltage range is obtained as

\[
L_m \leq \frac{n_p V_{o,d}}{16 V_{in} f_{sw} n_s C_{oss}}
\]

(9)

Mode 4 \([t_3 \sim t_4]\): At \( t_3 \), \( v_{CS2} = 0 \). Owing to \( i_{LM} > i_{Lr} \) and \( i_{Lr}(t_3) > 0 \), \( D_4 \) conducts and \( D_{S2} \) is forward-biased. Therefore, the zero voltage turn-on of \( S_2 \) can be achieved. Due to \( D_4 \) conducting, one can obtain \( v_{LM} = n_p V_{Co2}/n_s = n_p V_o/(2n_s) \) and \( i_{LM} \) decreases in this mode. The resonant frequency in this mode is \( f_3 \).

Mode 5 \([t_4 \sim t_5]\): At time \( t_4 \), \( i_{D4} = 0 \) and \( D_4 \) turns off without the reverse recovery current. The resonant frequency in mode 5 equals \( f_2 \).

Mode 6 \([t_5 \sim T_{sw}+t_0]\): At time \( t_5 \), \( S_2 \) turns off. Since \( i_{LM} < i_{Lr} \) and \( i_{Lr}(t_5) < 0 \), \( D_3 \) is forward-biased and \( C_{S1} \) is discharged. The zero voltage turn-on condition of \( S_1 \) is expressed as
\[ |i_{Lm}(t_5)| = i_{Lm}(t_2) = \frac{n_p V_o}{8R_s L_m f_{SW}} \geq V_{in} \sqrt{2C_{oss}/(L_m + L_r)} \]  (10)

At \( t_0 + T_{SW} \), \( v_{CS1} = 0 \).

Figure 4. Cont.
The presented LLC converter is operated with pulse frequency modulation (PFM). Power switches have a constant duty cycle \( d = 0.5 \) with variable frequency. The switching frequency is dependent on \( V_{in} \) and \( P_o \). The fundamental harmonic frequency approach is used to analyze the circuit characteristics. The variable winding turns controlled by the AC switch \( S_{ac} \) are used on the secondary side to adjust the voltage gain under low- or high-voltage input. Based on the fundamental harmonic frequency analysis, the AC equivalent circuit of the resonant tank is given in Figure 5. \( R_e \) is the equivalent primary side resistor of transformer. Since the PWM signals of \( S_1 \) and \( S_2 \) are square waves with duty cycle \( d = 0.5 \), the input voltage of the LLC circuit becomes a square voltage wave measuring 0 V (if \( S_2 \) is on) or \( V_{in} \) (if \( S_1 \) is on). The fundamental root mean square (rms) input voltage is calculated as \( \sqrt{2}V_{in}/\pi \). Since \( S_{ac} \) is turned on at low-voltage inputs and off at high-voltage inputs, the secondary winding turns \( 2n_s \) \( (n_s) \) are connected to the output load under low (high) input voltage ranges. The primary side magnetizing voltage is a square wave with a voltage value of \( v_{lm} = \pm n_p V_o/(4n_s) \) if \( S_{ac} \) is conducting or \( v_{lm} = \pm n_p V_o/(2n_s) \) if \( S_{ac} \) is off. The fundamental rms voltage \( V_{Lm,rms} \) is calculated as \( n_p V_o/ (\sqrt{2}\pi n_s) \) for low-input voltage ranges or \( \sqrt{2}n_p V_o/ (\pi n_s) \) for high-input voltage operation. The DC load resistor \( R_o \) is reflected to the primary side of the transformer and the AC equivalent resistor \( R_e \) can be obtained.

3. Circuit Analysis and Design Procedure

Figure 4. LLC circuits operated at high-input voltage ranges: (a) PWM waveforms; (b) mode 1; (c) mode 2; (d) mode 3; (e) mode 4; (f) mode 5; (g) mode 6.
The voltage transfer function in Figure 5 is calculated as in Equations (15) and (16) as $V_{in} = \frac{n_p}{n_s} V_{o}$ for either high- or low-input voltage ranges. For high-voltage range operation, the minimum voltage has the same resonant tank under high and low-voltage ranges, the prototype circuit can be designed when operated over wide voltage ranges. Therefore, the minimum primary winding turns are calculated as

$$n_p,_{\text{min}} \geq \frac{(n_p/n_s) V_o}{4 f_{s,\text{min}} \Delta B A_e} = \frac{(8.333) \times 48}{4 \times 50000 \times 0.4 \times 354 \times 10^{-6}} \approx 14.123$$

In the laboratory prototype, we selected $n_p = 16$ and $n_s = 2$. Thus, the actual DC gains are obtained in Equations (15) and (16) as

$$G_{\text{max},H} = \frac{V_o n_p}{V_{in,\text{min}} n_s} = \frac{48 \times 16}{200 \times 2} \approx 1.92$$
Similarly, the DC voltage gains of the presented resonant converter for low-voltage ranges are obtained as

\[
G_{\text{min},L} = \frac{V_o n_p}{2 V_{\text{in,min}} n_s} = \frac{48 \times 16}{2 \times 100 \times 2} \approx 0.96 \quad (16)
\]

\[
G_{\text{max},L} = \frac{V_o n_p}{2 V_{\text{in,max}} n_s} = \frac{48 \times 16}{2 \times 200 \times 2} \approx 1.92 \quad (17)
\]

\[
G_{\text{min},H} = \frac{V_o n_p}{V_{\text{in, max}}} = \frac{48 \times 16}{400 \times 2} \approx 0.96 \quad (18)
\]

From the calculated winding turns \( n_p \) and \( n_s \), the equivalent resistance \( R_e \) at high-voltage ranges and the rated power are calculated as

\[
R_e = \frac{2 (n_p / n_s)^2 R_o}{\pi^2} = \frac{2 \times (16 / 2)^2 \times (48^2 / 500)}{3.14159^2} \approx 60 \Omega \quad (19)
\]

The quality factor \( X \) is selected as 0.2. The resonant inductor \( L_r \) is calculated as

\[
L_r = \frac{x R_e}{2 \pi f_r} = \frac{0.2 \times 60}{2 \pi \times 100000} \approx 19 \mu\text{H} \quad (20)
\]

In the proposed converter, the actual resonant inductance \( L_r \) is 20 \( \mu\text{H} \). Then, the resonant capacitor is calculated as

\[
C_r = \frac{1}{4 \pi^2 L_r f_r^2} = \frac{1}{4 \pi^2 \times 20 \times 10^{-6} \times (100000)^2} \approx 127 \text{nF} \quad (21)
\]

Since the selected \( L_m = L_m / L_r = 7 \), the magnetizing inductance \( L_m \) is obtained as

\[
L_m = L_m L_r = 7 \times 20 = 140 \mu\text{H} \quad (22)
\]

MBR40100PT diodes with a 100 V/40 A rating are used for \( D_1-D_4 \). The selected output capacitance \( C_{o1} = C_{o2} = 540 \mu\text{F}/100 \text{V} \). The Schmitt voltage comparator with a reference voltage of 200 V is used to control switch \( S_{ac} \). Switch \( S_{ac} \) is implemented by two IRF1405ZPBF metal-oxide-semiconductor field-effect transistors (MOSFETs) with a 55 V/118 A rating. The voltage stress of \( S_1 \) and \( S_2 \) is \( V_{\text{in,max}} = 400 \text{V} \). In the prototype, STW48N60M2 MOSFETs with a 600 V/26 A rating are used for \( S_1 \) and \( S_2 \). The UCC25600 PWM integrated circuit is adopted to realize pulse frequency modulation. The voltage TL431 regulator and the PC817 optocoupler are adopted on the low-voltage side to achieve load voltage regulation and signal isolation, respectively. Figure 6 provides the circuit diagram of the prototype with the adopted control scheme.
4. Experimental Results

The measured waveforms of the prototype circuit are given to confirm the effectiveness of the presented LLC resonant converter with wide voltage operation. The experimental waveforms of $v_{S1_{ac}}$, $v_{S2_{ac}}$, $v_{Cr}$, and $i_{Lr}$ for different input voltages 100 V, 190 V, 210 V, and 400 V are given in Figure 7. For $V_{in} = 100$ V and 190 V, the presented converter is operated at low-voltage ranges. The $S_{ac}$ conducts and $2n_{s}$ winding turns are connected to the output load. The converter operated at $V_{in} = 100$ V has high-voltage gain compared to the converter operated at $V_{in} = 190$ V. One can observe that the switching frequency of the converter at $V_{in} = 190$ V is greater than the switching frequency at $V_{in} = 100$ V. For $V_{in} = 210$ V and 400 V conditions, the presented LLC converter is operated at high-voltage ranges. The $S_{ac}$ is turned off and $n_{s}$ instead of $2n_{s}$ winding turns are connected to the output load. Similarly, the proposed converter operated at $V_{in} = 210$ V has a lower switching frequency compared to the converter operated at $V_{in} = 400$ V. The experimental waveforms of $i_{D1}–i_{D4}$ for different input voltage conditions and the full power are given in Figure 8. For $V_{in} = 100$ V and 190 V conditions, the switch $S_{ac}$ is turned on and $D_{3}$ and $D_{4}$ are reverse-biased. Therefore, $i_{D3} = i_{D4} = 0$, as shown in Figure 8a,b. On the other hand, $S_{ac}$ is turned off, while $D_{1}$ and $D_{2}$ are also off for $V_{in} = 210$ V and 400 V conditions. Therefore, $i_{D1} = i_{D2} = 0$, as shown in Figure 8c,d. Diodes $D_{1}$ and $D_{2}$ turn off without the reverse recovery current at $V_{in} = 100$ V (Figure 8a) because $f_{sw}$ at $V_{in} = 100$ V is about 45 kHz $< f_{r1} = 100$ kHz. Similarly, $D_{3}$ and $D_{4}$ also turn off without the reverse recovery current at $V_{in} = 210$ V (Figure 8c) because $f_{sw}$ is about 50 kHz $< f_{r1} = 100$ kHz. The experimental waveforms of switch $S_{1}$ at 20% power and different input voltage conditions are given in Figure 9. The experimental waveforms of $S_{1}$ at 100% power and different input voltage conditions are illustrated in Figure 10. From the experimental waveforms in Figure 9; Figure 10, it is clear that $S_{1}$ is turned when zero voltage switches from 20% power to 100% power and over the entire voltage range.
Figure 7. Measured results of the secondary side diode currents at rated power values of: (a) $V_{in} = 100$ V; (b) $V_{in} = 190$ V; (c) $V_{in} = 210$ V; (d) $V_{in} = 400$ V.

Figure 8. Measured results of the secondary side diode currents at rated power values of: (a) $V_{in} = 100$ V; (b) $V_{in} = 190$ V; (c) $V_{in} = 210$ V; (d) $V_{in} = 400$ V.
The measured input voltage and the PWM signal of the AC switch $S_{ac}$ are provided in Figure 11. The hysteresis voltage band and reference voltage of the voltage comparator are 10 V and 200 V, respectively. First, the input voltage is increased from 100 V to 400 V. When $V_{in}$ is greater than 205 V, the PWM signal of $S_{ac}$ is changed from the “on” state (high-voltage) to “off” state (low-voltage). Second, the input voltage is decreased from 400 V. When $V_{in}$ is less than 195 V, $S_{ac}$ is turned on.
(high-voltage). Figure 12 gives the experimental switching frequency of the proposed resonant converter. For low-voltage ranges and the same load condition, the voltage gain of the resonant converter under $V_{in} = 100$ V is higher than $V_{in} = 190$ V. One can see that the measured switching frequency at 100 V input voltage is lower than 190 V input voltage. The experimental circuit efficiencies of the proposed converter at 100% load are 87.3% (100 V input), 91.2% (190 V input), 89.5% (210 V input), and 92.8% (400 V input). Since the turns ratio $n_{p}/n_{s}$ is used under high-voltage ranges ($V_{in} = 200–400$ V) instead of $n_{p}/(2n_{s})$ under low-voltage ranges ($V_{in} = 100–200$ V), the converter has a lower magnetizing current under high-voltage ranges compared to the low-voltage range operation. The converter has lower conduction loss on power switches at high-input voltage ranges. Therefore, the experimental circuit efficiency at 400 V input is better than the 190 V input, and the efficiency with the 210 V input is better than the 100 V input.

![Figure 11. The experimental waveforms of the input voltage $V_{in}$ and PWM signal of $S_{ac}$.](image)

![Figure 12. Measured switching frequency under various input voltages and load conditions.](image)

5. Conclusions

The half-bridge LLC resonant converter with variable winding turns is studied and implemented in this paper to accomplish two objectives: (1) wide, soft switching ranges over the entire input voltage and output road; and (2) wide voltage operation. The general half-bridge resonant converter is adopted on the high-voltage side to achieve series resonant circuit and realize the soft switching operation for powered semiconductors. Two sets of winding turns are adopted on the low-voltage side to accomplish wide voltage operation capability. Based on the circuit analysis and test results, the proposed LLC converter can achieve zero voltage switching with wide input voltage (100 V to 400 V) and wide load...
(20% to 100% rated power) ranges. Compared to the conventional LLC resonant converters with wide voltage operation, the proposed converter has lower component counts, a simple control scheme, and less voltage stress on rectifier diodes. The studied LLC converter can be used for power units in computers and power servers with large hold-up times, PV power converters, fuel cell converters, and battery chargers and dischargers with wide voltage operation. Experimental waveforms from a 500 W prototype are given and presented to confirm the usefulness of the presented circuit topology. The future works will investigate the new wide voltage converters with low cost and high power density advantages for industry power applications and renewable energy power conversion.

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