Implementation of a Wide Input Voltage Resonant Converter with Voltage Doubler Rectifier Topology

Bor-Ren Lin * and Yen-Chun Liu

Department of Electrical Engineering, National Yunlin University of Science and Technology, Yunlin 640, Taiwan; m10812048@yuntech.edu.tw
* Correspondence: linbr@yuntech.edu.tw; Tel.: +886-912-312-281

Received: 8 October 2020; Accepted: 16 November 2020; Published: 17 November 2020

Abstract: A new circuit structure of LLC converter is studied and implemented to achieve wide zero-voltage switching range and wide voltage operation such as consumer power units without power factor correction and long hold up time demand, battery chargers, photovoltaic converters and renewable power electronic converters. The dc converter with the different secondary winding turns is adopted and investigated to achieve the wide input voltage operation (50–400 V). To meet wide voltage operation, the full bridge and half bridge dc/dc converters with different secondary turns can be selected in the presented circuit to have three different voltage gains. According to input voltage range, the variable frequency scheme is employed to have the variable voltage gain to overcome the wide input voltage operation. Therefore, the wide soft switching load variation and wide voltage operation range are achieved in the presented resonant circuit. The prototype circuit is built and tested and the experiments are demonstrated to investigate the circuit performance.

Keywords: frequency modulation; dc/dc converters; wide voltage variation; soft switching operation

1. Introduction

Power electronics are more and more important for the energy conversion systems such as renewable power conversion, electric vehicle systems, traction vehicles, light rail vehicles, dc nano- or micro-grid systems, battery-based storage systems, personal computer power units and industry power units. Power semiconductors such as insulated gate bipolar transistors (IGBTs), Metal-Oxide-Semiconductor Field-Effect Transistor (MOSFET), Gallium Nitride (GaN) FET and Silicon Carbide (SiC) devices are widely used power switches in power electronic circuits. However, IGBT devices have low switching frequency problem. GaN and SiC devices have the advantages of low switching loss and high frequency operation. However, the cost of GaN and SiC is much more expensive than the IGBT and MOSFET devices. Compared to IGBT, GaN and SiC devices, MOSFET elements have low cost and medium frequency operation for modern power converters. Pulse-width modulation (PWM) power electronic circuits have been widely studied and developed in power electronics to lessen reduce global warming and air pollution problems. The dc-dc or dc-ac power converters are the most attractive interface circuits between the photovoltaic (PV) cells (or fuel cells) and the utility dc or ac voltage. However, the output voltage of PV cells, fuel cells and dc wind turbine power generators is not constant and related to solar intensity or wind speed. Conventional duty cycle control converters [1–4] and frequency control converters [5–7] cannot be operated well under wide voltage variation condition due to the limit voltage gain of power converters. Multi-stage dc converters with parallel or series structure [8–11] have been researched to overcome wide voltage variation problem and used for renewable energy conversion and battery charger applications. However, the disadvantages of multi stage dc-dc converters have low reliability and more active and passive power components. Two full bridge circuits structure was presented in [12] to overcome limit voltage range
The problem of conventional resonant converters. However, sixteen power semiconductors are used in this circuit structure. The conventional LLC converter with two sets of secondary windings was studied in [13] to have wide hold-up time operation for PC power units. The main drawbacks of this circuit structure are four secondary windings and large copper losses. The resonant converter with interleaved duty cycle control was achieved in [14] to overcome wide voltage variation problem and achieve low current ripple input. However, the circuit topology is still the multi stage converter.

The new resonant converter with voltage doubler configuration and two sets of secondary windings is presented in this paper to accomplish soft switching operation and overcome wide voltage input variation ($V_{in} = 50–400$ V). According to the input voltage range, the present LLC converter can be operated under full bridge resonant structure or half bridge resonant structure with different secondary turns. Therefore, the proposed converter has much wide range of input voltage operation. When $V_{in} = 50–100$ V (low voltage region), the full bridge type LLC circuit with more secondary turns is operated to have high voltage gain. If $V_{in} = 100–200$ V (medium voltage input region), the full bridge type LLC circuit with fewer secondary turns is adopted to have less dc voltage gain. If $V_{in} = 200–400$ V (high voltage input region), the half bridge structure is used to obtain the lowest voltage gain on the presented circuit. With the proposed control strategy, the wide voltage operation ($V_{in} = 50–400$ V) is accomplished in the presented circuit structure. Since the proposed resonant circuit is controlled under the variable switching frequency with input inductive impedance operation, the active switches have soft switching turn-on characteristic. The voltage doubler rectification structure can reduce the voltage rating of diodes compared to the center-tapped rectifier topology. Thus, the diode conduction loss can be reduced. The proposed LLC converter has less circuit components compared to the wide voltage resonant converter presented in [8–14]. Finally, the experimental verifications are provided to demonstrate the performance of the studied LLC resonant circuit.

2. Circuit Configuration and Principles of Operation

The conventional resonant converters with half bridge or full bridge structure are provided in Figure 1. The fundamental root-mean-square (rms) input voltage of full bridge structure is two times of the fundamental rms input voltage of half bridge structure. Therefore, the full bridge resonant converter has more power capability than the half bridge resonant converter. Owing to the center-tapped rectification structure, the diodes $D_1$ and $D_2$ has at least $2V_o$ voltage stress. The other disadvantage of the center-tapped rectifier is two winding sets are needed on the secondary side of transformer $T$. For high load current applications, there are more copper losses on the secondary windings.

![Figure 1. Cont.](image-url)
where $G$ is the voltage gain. Under this condition, the rectifier diodes $D_{ab}$ are always in the on-state. Figure 2b gives the circuit diagram for low input voltage operation. The voltage gain in Figure 2c is $V_{o}/V_{in,M} = G_{ac}(f_{sw})(2n_{s})/n_{p}$. If $200 \, V < V_{in} < 400 \, V$ (high voltage region $V_{in,M}$), the active switches $S_{3}$ and $S_{ac}$ are turned off and $S_{4}$ is always in the on-state. Figure 2d illustrates the circuit diagram operated at high voltage region. The leg voltage $v_{ab}$ has voltage level $V_{in}$ or $0$. Only $n_{s}$ winding turns and diodes $D_{3}$ and $D_{4}$ are connected to load. The voltage gain in Figure 2d is $V_{o}/V_{in,H} = G_{ac}(f_{sw})n_{s}/n_{p}$. From the three voltage gains in previous discussion, it can observe that $V_{o}/V_{in,L} = G_{ac}(f_{sw})(4n_{s})/n_{p} = 2V_{o}/V_{in,M} = 4V_{o}/V_{in,H}$. Thus, one can conclude that $V_{in,H} = 2V_{in,M} = 4V_{in,L}$. Therefore, the proposed LLC converter has the highest voltage gain at low voltage region $V_{in,L}$ and the lowest voltage gain at high voltage region $V_{in,H}$.

2.1. Low Voltage Region ($V_{in,L}$: 50–100 V)

When $S_{ac}$ is in the on-state, the LLC converter has less transformer turns-ratio $n_{p}/(2n_{s})$ and large voltage gain. Under this condition, the rectifier diodes $D_{3}$ and $D_{4}$ are always off. The voltage gain of the converter is obtained as $V_{o}/V_{in,L} = G_{ac}(f_{sw})(4n_{s})/n_{p}$. From the gating signals of $S_{1}$–$S_{4}$ and the on/off state of $D_{1}$ and $D_{2}$, six equivalent operating states per switching periods are provided in Figure 3.
Figure 2. Proposed converter: (a) circuit diagram; (b) low input voltage region; (c) medium input voltage region; (d) high input voltage region.

State 1 \([t_0-t_1]\): When \(t=t_0\), \(v_{CS1} = v_{CS4} = 0\) V. Power devices \(S_4\) and \(S_1\) turn on at this moment to realize soft switching operation due to \(i_{Lr}(t_0) < 0\). Since \(i_{Lr} > i_{Lm}\), \(D_1\) conducts in this state. In state 1, \(v_{ab} = V_{in,L}, v_{Lm} = (V_{C1}np)/(2ns) \approx (V_{onp})(4ns)\) and \(i_{Lm}\) increases. The resonant frequency of the LLC converter in this state is \(f_{r,1} = 1/(2\pi \sqrt{L_rC_r})\). If \(f_{r,1} > f_{sw}\), then the circuit operation will go to state 2 or 3.

State 2 \([t_1-t_2]\): When \(i_{D1} = 0\) at \(t_1\), \(D_1\) is turned off with zero-current switching. The resonant frequency in state 2 is \(f_{r,2} = 1/(2\pi \sqrt{C_r(L_r + L_m)})\). It is obvious that \(f_{r,1} > f_{r,2}\). The current variation on \(L_m\) approximates \(\Delta i_{Lm} \approx (npV_o)/(8t_3i_{Lm}f_{sw})\) and the magnetizing current \(i_{Lm}\) at \(t_2\) is \(i_{Lm}(t_2) = \Delta i_{Lm}/2 \approx (npV_o)/(16t_3i_{Lm}f_{sw})\).

State 3 \([t_2-t_3]\): Power devices \(S_1\) and \(S_4\) are off at time \(t_2\). Since \(i_{Lr}(t_2)\) is positive, \(v_{CS3}\) and \(v_{CS2}\) are decreased. Due to \(i_{Lm}(t_2) > i_{Lr}(t_2)\), \(D_2\) is conducting. To ensure the soft switching turn-on
of S₃ and S₂, the current \( i_{Lm}(t₂) \) must be greater than \( V_{in,L} \sqrt{C_{oss}/(L_m + L_r)} \) where \( C_{oss} = C_{S1} = \ldots = C_{S4} \). The other necessary condition for zero-voltage switching is that the dead time \( t_d \) between \( S_2 \) and \( S_1 \) is greater than time interval in state 3. To accomplish this condition, one can obtain \( L_{m,max} = (t_d V_o n_p)/(32 C_{oss} n_s f_{sw} V_{in,L}) \).

State 4 [\( t_3-t_4 \)]: At time \( t_3 \), \( v_{CS2} = v_{CS3} = 0 \) V. At this moment, power devices \( S_3 \) and \( S_2 \) are turned on under zero voltage. Owing to \( i_{Lm}(t_3) > i_{Lr}(t_3), D_2 \) conducts. In state 2, \( v_{ab} = -V_{in,L} v_{Lm} = -(V_{Co2} n_p)/(2 n_s) \approx -(V_o n_p)/(4 n_s), i_{Lm} \) decreases and \( C_{o1} (C_{o2}) \) is discharged (charged). If \( f_{r,1} > \) or < \( f_{sw} \), then the circuit operation will go to state 4 or 6.

State 5 [\( t_4-t_5 \)]: If \( f_{r,1} > f_{sw} \), one obtains \( i_{D2} \) equals 0 at \( t_4 \). Then, \( D_2 \) turns off. The resonant frequency in state 4 is \( f_{r,2} = 1/[2 π \sqrt{C_r (L_r + L_m)}] \). The magnetizing current \( i_{Lm} \) at \( t_5 \) approximates \( i_{Lm}(t_5) \approx -(n_p V_o)/(16 n_s L_m f_{sw}) \).

---

**Figure 3. Cont.**
The converter has low transformer turns-ratio.

Figure 4 shows the key PWM waveforms and state circuits for medium voltage operation (100–200 V).

If $V_{in}$ is in the medium voltage region between 100 V and 200 V, the switch $S_{ac}$ is turned off. The converter has low transformer turns-ratio $n_p/n_n$ and less voltage gain. The circuit diagram is shown in Figure 2c. The voltage gain at medium voltage operation is equal to $V_{os}/V_{in,M} = G_{ac}(f_{sw})(2n_n)/n_p$.

Figure 4 shows the key PWM waveforms and state circuits for medium voltage operation (100–200 V).
State 1 \([t_0 - t_1]\): The drain-to-source voltages of \(S_1\) and \(S_4\) are decreased and equal to zero at \(t_0\). Due to \(i_{LR}(t_0) < 0\), the primary-side current \(i_{LR}\) flows through the anti-parallel diodes \(D_{S4}\) and \(D_{S3}\). Power devices \(S_4\) and \(S_1\) turn on at this moment to accomplish soft switching turn-on. Since \(i_{LR} > i_{LM}\), \(D_3\) conducts. From Figure 4b, it can be obtained that \(v_{ab} = V_{in}, v_{Lm} \approx (V_o n_p)/(2 n_i), i_{LM}\) increases and the resonant frequency \(f_{r1} = 1/[2 \pi \sqrt{L_r C_r}]\). In this state, \(C_{o1}\) is charged and \(C_{o2}\) is discharged.

State 2 \([t_1 - t_2]\): If \(f_{r1} > f_{sw}\), then \(i_{D3}\) is decreased to zero ampere at time \(t_1\) and \(D_3\) turns off at zero-current switching. Then, the resonant frequency in state 2 is \(f_{r2} = 1/[2 \pi \sqrt{C_r (L_r + L_m)}]\). The magnetizing current \(i_{LM}\) at the end of this state is \(i_{LM}(t_2) \approx (n_p V_o)/(8 n_i L_{m, sw})\).

State 3 \([t_2 - t_3]\): Power devices \(S_4\) and \(S_1\) are turned off in state 3. The positive primary current \(i_{LR}(t_2)\) will discharge \(C_{S2}\) and \(C_{S3}\). Owing to \(i_{LM}(t_2) > i_{LR}(t_2), D_4\) on the output-side is conducting. The soft switching turn-on condition of \(S_3\) and \(S_2\) is \(i_{LM}(t_2) \geq V_{in,M} \sqrt{C_{oss} / (L_m + L_r)}\). To ensure \(v_{CS2} = v_{CS3} = 0\) at \(t_3\), the maximum magnetizing inductance \(L_{m, max}\) is obtained as \(L_{m, max} = (L_r V_o n_p)/(16 C_{oss} n_s f_{sw} V_{in,M})\).

State 4 \([t_3 - t_4]\): The \(v_{CS3}\) and \(v_{CS2}\) are decreased to zero at \(t_3\). The primary-side current \(i_{LR}(t_3)\) is positive and flows through the body diodes \(D_{S2}\) and \(D_{S3}\). At this moment, power devices \(S_3\) and \(S_2\) turn on to realize zero-voltage switching. Since \(i_{LM}(t_3) > i_{LR}(t_3), D_4\) conducts. From Figure 4c, it can be obtained that \(v_{ab} = -V_{in}, v_{Lm} \approx -(V_o n_p)/(2 n_i)\) and \(i_{LM}\) decreases. The output capacitors \(C_{o1}\) and \(C_{o2}\) are discharged and charged.

State 5 \([t_4 - t_5]\): \(i_{D4}\) is decreased and equal to zero at \(t_4\). Then, \(D_4\) turns off. The resonant frequency in state 5 is \(f_{r2} = 1/[2 \pi \sqrt{C_r (L_r + L_m)}]\). The magnetizing current \(i_{LM}\) at \(t_5\) approximates \(i_{LM}(t_5) \approx -(n_p V_o)/(8 n_i L_{m, sw})\).

State 6 \([t_5 - T_{sw} + t_0]\): Power devices \(S_3\) and \(S_2\) turn off in this state. \(i_{LR}(t_5)\) is negative and discharges \(C_{S1}\) and \(C_{S4}\). Since \(i_{LR}(t_5) > i_{LM}(t_5), D_3\) is conducting. The soft switching turn-on condition of \(S_4\) and \(S_1\) is obtained as \(i_{LM}(t_5) \geq V_{in,M} \sqrt{C_{oss} / (L_m + L_r)}\).

Figure 4. Cont.
2.3. High Voltage Region ($V_{in,H}$: 200–400 V)

When $V_{in}$ is increased from 200 V to 400 V, the presented circuit is controlled under high input voltage region. Power devices $S_3$ and $S_{ac}$ turn off and $S_4$ turns on. Only power semiconductors $S_2$ and
S1 are controlled with frequency modulation so that the half bridge LLC resonant circuit \((S1, S2, S4, Lr, C_r, \text{ and } T)\) is operated on the input-side. The voltage gain of the converter at high voltage region is \(V_d/V_{in,H} = G_{ac}(f_{sw})n_p/n_p\). From the on/off state of \(D3, D4, S1\) and \(S2\), six equivalent operating states per switching periods can be observed in Figure 5 for high input voltage range (200–400 V).

State 1 \([t_0-t_1]\): The capacitor voltage \(v_{CS1} = 0\) at time \(t_0\). The primary-side current \(i_{Lr}(t_0)\) is negative so that the body diode \(D_{S1}\) conducts and the leg voltage \(v_{ab} = V_{in}\). Due to \(i_{Lr} > i_{Lm}, D3\) is forward biased. The inductor voltage \(v_{Lm} \approx (V_0 n_p)/(2n_p)\). The resonant frequency \(f_{r1} = 1/[2\pi \sqrt{C_r(L_r + L_m)}]\) and \(C_{oi}\) is charged in this state.

State 2 \([t_1-t_2]\): If \(f_{r1} > f_{sw}, i_{D2} = 0\) at time \(t_1\). Then \(D3\) is turned off at zero-current switching. The resonant frequency in state 2 is expressed as \(f_{r2} = 1/[2\pi \sqrt{C_r(L_r + L_m)}]\). At time \(t_2, i_{Lm}(t_2) = (n_p V_o)/(8n_p L_m f_{sw})\) and \(S1\) turns off.

State 3 \([t_2-t_3]\): Power device \(S1\) is turned off in this state. The primary current \(i_{Lr}(t_2)\) is positive and discharges \(C_{S2}\). To ensure the soft switching turn-on of \(S2\), the inductor current \(i_{Lm}(t_2)\) must greater than \(V_{in,H} \sqrt{2C_{oss}/(L_m + L_r)}\). The other necessary condition for zero-voltage switching is that the dead time \(t_d\) between \(S2\) and \(S1\) is greater than time interval in state 3. To achieve this condition, the magnetizing inductance can obtain \(L_{m,max} = (L_d V_o n_p)/(16C_{oss} n_p f_{sw} V_{in,H})\).

State 4 \([t_3-t_4]\): The drain-to-source voltage \(v_{CS2} = 0\) at \(t_3\). The primary current \(i_{Lr}(t_3) > 0\) and \(i_{Lr}\) flows through the body diode \(D_{S2}\). At this moment, \(S2\) turns on to have zero-voltage switching operation. Since \(i_{Lr}(t_3) < i_{Lm}(t_3), D4\) is conducting. From Figure 5e, it is clear that \(v_{ab} = 0, v_{Lm} = -(V_0 n_p)/(2n_p)\) and \(i_{Lm}\) decreases.

State 5 \([t_4-t_5]\): \(i_{D4}\) is decreased to zero at time \(t_4\) and \(D4\) turns off at zero-current switching. The resonant frequency in state 5 is \(f_{r2} = 1/[2\pi \sqrt{C_r(L_r + L_m)}]\). \(S2\) is turned off at the end of this state and \(i_{Lm}(t_5)\) is expressed as \(i_{Lm}(t_5) = -(n_p V_o)/(8n_p L_m f_{sw})\).

State 6 \([t_5-T_{sw} + t_0]\): Owing to \(i_{Lr}(t_5) < 0, i_{Lr}\) discharge \(C_{S1}\). The soft switching turn-on of \(S1\) is \(i_{Lm}(t_5) \geq V_{in,H} \sqrt{2C_{oss}/(L_m + L_r)}\). The capacitor \(C_{S1}\) is discharged to zero voltage at time \(t_0 + T_{sw}\).

Figure 5. Cont.
Figure 5. Cont.
3. Circuit Characteristics and Design Example

The proposed LLC resonant converter is operated by variable frequency control. The square voltage waveform is generated on the leg voltage \( v_{ab} \) with voltage values \( \pm V_{in} \) in Figure 2b, c or \( V_{in} \) and 0 in Figure 2d. The circuit characteristics of the converter are based on the fundamental frequency analysis. Figure 6 gives the equivalent resonant circuit on the input-side. \( R_{e,ac} \) is the primary-side resistance of transformer \( T \) and \( v_{ab,\text{rms}} \) is the input fundamental voltage. From the leg voltage \( v_{ab} \) in Figure 2, the fundamental root mean square (rms) voltage \( v_{ab,\text{rms}} \) can be expressed in Equation (1).

\[
v_{ab,\text{rms}} = \begin{cases} 
2 \sqrt{\frac{V_{in}}{\pi}}, & \text{in low and medium input voltage regions} \\
\sqrt{\frac{2V_{in}}{\pi}}, & \text{in high input voltage region}
\end{cases}
\]  

In low input voltage operation, \( S_{ac} \) is always in the on-state. The equivalent turns-ratio of transformer \( T \) in Figure 2b is \( n_p/(2n_s) \) and the \( \text{rms} \) value of the magnetizing voltage \( v_{lm,\text{rms}} = V_o n_p / (\sqrt{2\pi n_s}) \). In medium and high input voltage ranges (Figure 2c,d), \( S_{ac} \) is always in the off-state. The equivalent turns-ratio of transformer \( T \) is \( n_p/n_s \). Thus, the \( \text{rms} \) value of the magnetizing voltage \( v_{lm,\text{rms}} = \sqrt{2}V_o n_p / (\pi n_s) \). The primary-side equivalent resistance \( R_{e,ac} \) is obtained as \( R_{e,ac} = \left( \frac{n_p}{n_s} \right)^2 R_o / (2\pi^2) \) in low input voltage operation or \( R_{e,ac} = 2 \left( \frac{n_p}{n_s} \right)^2 R_o / \pi^2 \) in medium and high input voltage operation. From the resonant circuit in Figure 6, the output/input transfer function is expressed in Equation (2).

\[
\left| G_{ac}(f_{sw}) \right| = 1 / \sqrt{ \left( \frac{f_o}{f_{sw}} \right)^2 - 1 } + \left[ 1 + \left( \frac{f_o}{f_{sw}} \right)^2 - 1 \right] = \begin{cases} 
\frac{V_o n_p}{V_{o,lm,\text{rms}}}, & \text{in low voltage region} \\
\frac{V_o n_p}{V_{o,lm,\text{rms}}}, & \text{in medium voltage region} \\
\frac{V_o n_p}{V_{o,lm,\text{rms}}}, & \text{in high voltage region}
\end{cases}
\]
The output voltage $V_o$ can be expressed in Equations (3)–(5) for low, medium and high input voltage regions.

$$V_o = 4V_{in,L}n_s/(n_p)$$

$$V_o = 2V_{in,M}n_s/(n_p)$$

$$V_o = V_{in,H}n_s/(n_p)$$

In the proposed circuit, the electric specifications are $V_{in} = 50–400$ V, $V_o = 48$ V and $P_o = 500$ W. The design resonant frequency $f_r = 100$ kHz. The assumed inductance ratio $L_m/L_r$ is 6. Since the resonant tank is identical for the operation in low, medium and high voltage ranges as shown in Figure 6, the following design procedures are based on the high input voltage condition ($V_{in,H} = 200–400$ V). The assumed voltage gain of resonant converter is 0.95 at $V_{in} = 400$ V case. Thus, the primary-secondary turn $n_p/n_s$ can be calculated as.

$$n_p/n_s = G_{dc}(f_{sw})V_{in}/V_o = 0.95 \times 400/48 = 7.916$$

The transformer $T$ is implemented by using the magnetic core TDK EE-55 with primary turns $n_p = 16$ and secondary turns $n_s = 2$. Therefore, the actual voltage gains at 200 V and 400 V input cases are rewritten as.

$$G_{dc,max} = n_pV_o/n_sV_{in} = 1.92$$

$$G_{dc,min} = n_pV_o/n_sV_{in} = 0.96$$

According to the winding turns and the load resistor at full load, $R_{eq,dc}$ is calculated in Equation (9).

$$R_{eq,dc} = 2(n_p/n_s)^2R/\pi^2 \approx 60 \Omega$$

It is assumed the quality factor $x = \sqrt{L_r/C_r}/R_{eq,dc} = 0.1$. Then, $L_r$ can be derived in Equation (10).

$$L_r = xR_{eq,dc}/(2\pi f_r) \approx 10 \mu H$$

From the given inductance ratio $L_m/L_r = 6$, $L_m$ is calculated as $L_m = 6 \times L_r = 60 \mu H$. The series resonant capacitance $C_r$ is expressed in Equation (11).

$$C_r = 1/(4\pi^2 f_r^2) \approx 254 \text{nF}$$

The maximum voltage rating of $S_1$–$S_4$ equals $V_{in,max} (= 400$ V). The voltage rating of $D_1$–$D_4$ is equal to $V_o (= 48$ V). Power switches STF40N60M2 (650 V/22 A) are used for $S_1$–$S_4$ and switch IXTP160N075T (75 V/160 A) is used for $S_{ac}$ in the prototype circuit. Power diodes MBR40100PT (100 V/40 A) are adopted in the prototype for the rectifier diodes $D_1$–$D_4$. The selected capacitances $C_{c1}$ and $C_{c2}$ are 940 μF with 100 V voltage rating. The control block of a laboratory prototype is shown in Figure 7a. Two Schmitt voltage comparators (comp 1 and comp 2) with reference voltages at 100 V and 200 V are used to select three input voltage regions. The control unit of the converter is using the integrated
The logic gates such as AND and OR gates are used to generate the control PWM signals of $S_{ac}$, $S_3$ and $S_4$. The relationship between the PWM signals of power switches and the input voltage is provided. It can be observed that $S_{ac}$ is always ON and $S_1$–$S_4$ are active if $50 \, \text{V} < V_{in} < 100 \, \text{V}$. If $200 \, \text{V} > V_{in} > 100 \, \text{V}$, $S_{ac}$ is always OFF and $S_1$–$S_4$ are active. When $V_{in} > 200 \, \text{V}$, $S_{ac}$ and $S_3$ are always OFF, $S_4$ is always ON, and $S_1$ and $S_2$ are active.

Figure 7. Laboratory prototype: (a) circuit diagram and control block; (b) control signal and input voltage.
4. Experimental Results

The experimental waveforms are provided to confirm the effectiveness of the presented resonant converter. Figures 8–10 give the experimental results of PWM signals of $S_1$–$S_4$ for low, medium and high input voltage regions, respectively. For low voltage region operation, $S_{ac}$ always turns on, the $2n_s$ secondary turns are connected to load and diodes $D_3$ and $D_4$ are the reverse biased. The gating signals $v_{S1,g}^{-}v_{S4,g}$ under 50 V and 90 V input conditions are provided in Figure 8a,b. The converter operated at $V_{in} = 50$ V condition has the low switching frequency compared to 90 V input condition. Figure 8c,d provide the test waveforms of $S_1$ at 20% and 100% output power under 50 V input condition. Similarly, the test waveforms of $S_1$ at 20% and 100% output power under 90 V input condition are provided in Figure 8e,f. One can observe the switch $S_1$ is tuned on at zero voltage switching for both 50 V and 90 V input conditions from 20% output power. Power switches $S_2$–$S_4$ have the similar switching characteristics as switch $S_1$. Thus, the soft switching turn-on of $S_1$–$S_4$ can be achieved under low voltage input region. In the same manner, the measured results of $S_1$–$S_4$ for medium and high voltage input regions are shown in Figures 9 and 10, respectively. For high voltage input region, the half bridge LLC resonant converter is activated and controlled. Thus, $S_3$ and $S_4$ are always turn-off and turn-on as shown in Figure 10a,b, respectively. Figure 11 gives the test waveforms $v_{ab}$, $v_{Cr}$ and $i_{tr}$ at $V_{in} = 50$ V, 110 V, 190 V and 400 V conditions. In the same manner, the experimental results of the rectifier diode currents, load voltage and load current at $V_{in} = 50$ V, 110 V, 190 V and 400 V conditions are shown in Figure 12. For 50 V input case, the studied converter is controlled at low voltage input region and $S_{ac}$ is turned on. The turns-ratio of transformer $T$ is $n_{ph}/2n_s$. Diodes $D_3$ and $D_4$ are off. Since the studied circuit has much more voltage gain at $V_{in} = 50$ V than $V_{in} = 100$ V, the circuit operated at $V_{in} = 50$ V has less switching frequency. It can observe that $D_1$ and $D_2$ turn off without the reverse recovery current loss shown in Figure 12a. For $V_{in} = 110$ V (Figure 12b) and 190 V (Figure 12c) conditions, the circuit is operation in medium input voltage range. $S_{ac}$ is in the off-state and the transformer turns-ratio is $n_p/n_s$. Diodes $D_1$ and $D_2$ are inactive. The voltage gain of the converter operated at $V_{in} = 110$ V is greater than $V_{in} = 190$ V condition. Therefore, the switching frequency at 110 V input (Figure 11b) is less than 190 V input condition (Figure 11c). Since $f_{sw}$ (switching frequency) at 110 V input is lower than $f_r$ (resonant frequency), $D_3$ and $D_4$ are turned off under zero-current switching shown in Figure 12b. For $V_{in} = 400$ V input, the resonant converter is operation at high voltage input region. $S_3$ and $S_{ac}$ are always turn-off and switch $S_4$ is always in the on-state. One can observe there is a dc voltage value ($V_{in}/2 = 200$ V) on voltage $v_{Cr}$ shown in Figure 11d. Owing to $f_{sw} > f_r$ at 400 V input, $D_3$ and $D_4$ are turned off with hard switching shown in Figure 12d. The experimental results of $V_{in}$, $V_{Ce1}$, $V_{Ce2}$ and $I_o$ for different input voltage conditions are provided in Figure 13. It observes that $V_{Ce1}$ and $V_{Ce2}$ are balanced well each other under different input voltage conditions. Figure 14a gives the measured input voltage $V_{in}$, the gating voltage $v_{Sacc,g}$ and output voltage $V_o$ between 50 V (low voltage region)–130 V (medium voltage region)–250 V. When $V_{in}$ is lower or greater than 100 V, $S_{ac}$ turns on (low input voltage range) or off (medium input voltage range). Figure 14b gives test waveforms of $V_{in}$, $v_{S3,g}$ and $v_{S4,g}$ between $V_{in} = 0$ V and 250 V. If $V_{in}$ is lower (or greater) than 200 V, $S_3$ is active (or always turn-off) and $S_4$ is active (or always turn-on). The test PWM signals of $S_3$, $S_4$ and $S_{ac}$ and input voltage $V_{in}$ shown in Figure 14 are agreed with the theoretical waveforms shown in Figure 7.
Figure 8. Experimental results of PWM signals for low input voltage region: (a) $S_1$–$S_4$ at 50 V input and 100% power; (b) $S_1$–$S_4$ at 90 V input and 100% power; (c) $S_1$ voltage and current at 50 V input and 20% power; (d) $S_1$ voltage and current at $V_{in} = 50$ V input and 100% power; (e) $S_1$ voltage and current at 90 V input and 20% power; (f) $S_1$ voltage and current at 90 V input and 100% power.
Figure 9. Experimental results of PWM signals for medium input voltage region: (a) $S_1$–$S_4$ at 110 V input and 100% power; (b) $S_1$–$S_4$ at 190 V input and 100% power; (c) $S_1$ voltage and current at 110 V input and 20% power; (d) $S_1$ voltage and current at $V_{in} = 110$ V input and 100% power; (e) $S_1$ voltage and current at 190 V input and 20% power; (f) $S_1$ voltage and current at 190 V input and 100% power.

Figure 10. Cont.
Figure 10. Experimental results of PWM signals for high input voltage region: (a) $S_1$–$S_4$ at 210 V input and 100% power; (b) $S_1$–$S_4$ at 400 V input and 100% power; (c) $S_1$ voltage and current at 210 V input and 20% power; (d) $S_1$ voltage and current at $V_{in} = 210$ V input and 100% power; (e) $S_1$ voltage and current at 400 V input and 20% power; (f) $S_1$ voltage and current at 400 V input and 100% power.

Figure 11. Measured primary-side voltage and current waveforms at 100% load: (a) 50 V input (low voltage range); (b) 110 V input (medium voltage range); (c) 190 V input (medium voltage range); (d) 400 V input (high voltage range).
Figure 12. Experimental results of the secondary-side currents and load voltage at 100% load; (a) 50 V input (low voltage range); (b) 110 V input (medium voltage range); (c) 190 V input (medium voltage range); (d) 400 V input (high voltage range).

Figure 13. Measured results of $V_{in}$, $V_{C01}$, $V_{C02}$ and $I_o$ at 100% load: (a) 50 V input (low voltage range); (b) 110 V input (medium voltage range); (c) 190 V input (medium voltage range); (d) 400 V input (high voltage range).
5. Conclusions

The hybrid LLC converter with three equivalent sub-circuit topologies is proposed and discussed to realize wide soft switching turn-on and wide voltage input operation. According to the switching status of power devices, the full bridge and half bridge LLC circuit with variable transformer turn-ratio are operated to accomplish wide voltage operation ($V_{in} = 50–400$ V). Owing to the LLC circuit tank, power switches have soft switching characteristic at turn-on instant. The presented resonant converter can be applied to dc-dc converters with wide voltage variation capability such as power units in PV power converters, power servers with large hold-up times and battery chargers and dischargers. The experimental results are demonstrated and provided to confirm the effectiveness of the adopted circuit topology.

Author Contributions: B.-R.L. designed and evaluated this project and was also responsible for writing this paper. Y.-C.L. measured the experimental waveforms. All authors have read and agreed to the published version of the manuscript.

Funding: This research is supported by the Ministry of Science and Technology (MOST), Taiwan, under grant number MOST 108-2221-E-224-022-MY2.

Acknowledgments: The authors are grateful to the all the editor and the reviewers for their valuable suggestions to improve this paper.

Conflicts of Interest: The author declares no conflict of interest.

References

1. Kim, J.Y.; Kim, H.S.; Baek, J.W.; Jeong, D.K. Analysis of effective three-level neutral point clamped converter system for the bipolar LVDC distribution. Electronics 2019, 8, 691. [CrossRef]
2. Almalaq, Y.; Matin, M. Three topologies of a non-isolated high gain switched-capacitor step-up cuk converter for renewable energy applications. Electronics 2018, 7, 94. [CrossRef]
3. Lin, B.R. Phase-shift pwm converter with wide voltage operation capability. Electronics 2020, 9, 47. [CrossRef]
4. Lin, B.R. Analysis of a dc converter with low primary current loss and balance voltage and current. *Electronics* 2019, 8, 439. [CrossRef]

5. Steigerwald, R.L. A comparison of half-bridge resonant converter topologies. *IEEE Trans. Power Electron.* 1988, 3, 174–182. [CrossRef]

6. Lin, B.R.; Chu, C.W. Hybrid full-bridge and LLC converter with wide ZVS range and less output inductance. *IET Power Electron.* 2016, 9, 377–384. [CrossRef]

7. Lee, J.B.; Kim, J.K.; Baek, J.I.; Kim, J.H.; Moon, G.W. Resonant capacitor on/off control of half-bridge LLC converter for high efficiency server power supply. *IEEE Trans. Ind. Electron.* 2016, 63, 5410–5415. [CrossRef]

8. Sun, W.; Xing, Y.; Wu, H.; Ding, J. Modified high-efficiency LLC converters with two split resonant branches for wide input-voltage range applications. *IEEE Trans. Power Electron.* 2018, 33, 7867–7870. [CrossRef]

9. Hu, H.; Fang, X.; Chen, F.; Shen, Z.J.; Batarseh, I. A modified high-efficiency LLC converter with two transformers for wide input-voltage range applications. *IEEE Trans. Power Electron.* 2013, 28, 1946–1960. [CrossRef]

10. Lu, J.; Kumar, A.; Afridi, K.K. Step-down impedance control network resonant DC-DC converter utilizing an enhanced phase-shift control for wide-input-range operation. *IEEE Trans. Ind. Appl.* 2018, 54, 4523–4536. [CrossRef]

11. Jeong, Y.; Kim, J.K.; Lee, J.B.; Moon, G.W. An asymmetric half-bridge resonant converter having a reduced conduction loss for DC/DC power applications with a wide range of low-input voltage. *IEEE Trans. Power Electron.* 2017, 32, 7795–7804. [CrossRef]

12. Lin, B.R. Resonant converter with soft switching and wide voltage operation. *Energies* 2019, 12, 3479. [CrossRef]

13. Lin, B.R. Series resonant converter with auxiliary winding turns: Analysis, design and implementation. *Int. J. Electron.* 2018, 105, 836–847. [CrossRef]

14. Lin, B.R. Resonant converter with wide input voltage range and input current ripple free. *IET Proc. Electron. Lett.* 2018, 54, 1086–1088. [CrossRef]

**Publisher’s Note:** MDPI stays neutral with regard to jurisdictional claims in published maps and institutional affiliations.

© 2020 by the authors. Licensee MDPI, Basel, Switzerland. This article is an open access article distributed under the terms and conditions of the Creative Commons Attribution (CC BY) license (http://creativecommons.org/licenses/by/4.0/).