A Boost CLLC Converter Controlled by PWM and PFM Hybrid Modulation for Photovoltaic Power Generation

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ABSTRACT When an input terminal is connected to a solar panel with a low voltage level and large power range fluctuation, a Boost type CLLC converter controlled by fixed frequency pulse width modulation (PWM) reduces the utilization of PV electric energy, and the synchronous drive of the secondary side and primary side ensures that the converter operates with low efficiency when the duty cycle is shifted by 0.5. To maximize the utilization of photovoltaic (PV) energy in a PV power generation system and improve the conversion efficiency, this paper studies a Boost CLLC converter with high gain and soft switching controlled by PWM-pulse frequency modulation (PFM) hybrid modulation for PV power generation. By controlling the duty cycle and switching frequency of the primary MOSFETs, the converter can stabilize the output voltage while tracking the PV maximum power point. Under such a hybrid control, the resonant gain of the converter is greater than 1, the voltage gain is large, one of the MOSFETs at the primary side is in a zero-voltage-switching-open (ZVS-ON) state, two diodes are used to realize zero-current-switching (ZCS) soft switching on the secondary side, and the efficiency of the converter is high. In addition, this paper studies the phenomenon of backflow power in the resonant circuit and the influence of a change in the switching frequency on the magnitude of the backflow power. Finally, a 500W experimental prototype was constructed to verify the accuracy of the theoretical analysis and the feasibility of the control method.

INDEX TERMS PWM-PFM hybrid control, boost CLLC, soft-switching and high-gain, backflow power.

I. INTRODUCTION

Among renewable energy sources, photovoltaic (PV) has been the fastest-growing energy technology over the past 15 years [1]. Compared with limited fossil fuels, sunlight is abundant in nature. PV power generation process is clean and pollution-free. Therefore, increasing the proportion of PV power generation in the grid power generation system is of great significance for promoting national economic development and solving the problem of pollution caused by traditional fossil fuel power generation. The PV power range fluctuates widely, with a low output voltage level and wide output voltage range. Thus, a PV interface converter not only needs to adapt to a wide range of voltage input and high voltage gain but also needs to maximize the use of electrical energy and have high conversion efficiency.

Taking two series-connected JKM250P-60 solar PV cells as an example, one finds that such a PV cell has a wide voltage range (usually 20-78V). Therefore, the input characteristics of the PV interface converter need to match the voltage of the solar PV cell. Moreover, soft switching techniques such as zero-voltage-switching (ZVS) and zero-current-switching (ZCS) are expected to be used [14], [15] when designing a PV interface converter to reduce the switching losses of the converter and achieve higher conversion efficiency. The large popularity of resonant converter gain in the application of PV converters is due to the benefits of soft switching and high voltage gain [2]–[5]. According to PV characteristics, we know that without additional control of a PV, the output power cannot reach a
maximum. Studies show that PV interface converters with MPPT function can greatly improve the utilization of PV energy [6]–[8]. In addition, studies reported in the literature [9]–[12] have proposed improvements to the traditional MPPT method and improved the application of the MPPT method. Therefore, a PV interface converter preferably has MPPT functionality.

A critical issue with a PV interface converter is how to adapt it to a wide range of PV voltage output and maximize the use of PV energy. Generally, there are two feasible solutions: a two-stage circuit and single-stage high-gain circuit. When a two-stage circuit is used, a prestige circuit is responsible for implementing MPPT, and a poststage circuit is responsible for boosting to facilitate the grid connection of the inverter. The two-stage circuit solution is suitable for large-scale centralized grid connections for PV and is not suitable for the current distributed power generation concept.

A single-stage high-gain circuit solution is a cost-effective way to solve the above problems and to enhance the conversion efficiency. However, it is difficult for a single-stage circuit to realize MPPT and increase the PV voltage while stabilizing the output voltage, so additional control is needed. Reference [13] added an additional energy storage part to the circuit to form a three-port circuit, and used a hybrid control method to control the current of the energy storage part to balance the changes in the input current brought about by a hybrid control method to realize MPPT and increase the PV voltage while stabilizing the output voltage. This method results in extra hardware cost, and the configuration of the energy storage system capacity needs to be discussed. Therefore, this method is not suitable for small-scale power generation systems, such as distributed roof PV power generation. It is preferable to achieve the above requirements without use of an energy storage unit.

In [14], four typical topologies are compared, including DAB [15], [16] and CLLC [17], [18], and the results show that a half-bridge CLLC has obvious advantages in terms of current stress, transformation efficiency and hardware cost. Reference [5] introduces an integrated CLLC resonant circuit without an energy storage unit by integrating a Buck-Boost circuit and a CLLC resonant circuit using common switches. The input and output are superimposed onto each other to achieve a high gain for the circuit. The circuit also shows soft switching characteristics. However, when the input terminal is connected to a solar panel with a low voltage level and a large power range fluctuation, the driving synchronization for the primary and secondary sides of the converter switch tube leads to a lower efficiency for the duty cycle when it is shifted by 0.5. The converter under fixed frequency PWM control has four switch tubes and eight switch processes. When the duty cycle shifts by 0.5, only two of the switching processes achieve soft switching. The fixed frequency pulse width modulation (PWM) control method reduces the use of PV energy. Therefore, additional control needs to be added to the circuit to improve the utilization of PV power.

Through research of the existing literature, it is found that the main features of a PV interface converter are high conversion efficiency, soft switching, MPPT, low hardware cost and the ability to adapt to a wide range of PV voltage input. In this paper, we propose a Boost-type soft-switching, high-gain CLLC converter with PWM-pulse frequency modulation (PFM) hybrid control for PV power generation. By controlling the duty cycle and switching frequency of the switches, the converter can track the PV maximum power point while stabilizing the output voltage to improve the power generation efficiency of the system. Compared with a traditional synchronous PWM control method [5], the hybrid control method utilizes uncontrolled rectification on the secondary high voltage side, and achieves five soft switching processes in eight switching processes. In a PV grid-connected system, compared with the traditional two-stage topology, the converter studied in this paper reduces the system cost, frees the control degree of freedom for the last stage grid-connected inverter, and helps one to provide more grid services. The resonant gain of the converter under hybrid control is greater than 1, and the DC voltage gain is high, which avoids the limit duty cycle of the converter and improves the stability of the circuit. ZVS for the primary-side switch is simple to implement, and ZCS soft switching is realized under the conditions of a duty cycle, full switching frequency, and the overall converter efficiency is high.

This paper is organized as follows: Section II provides a comprehensive introduction to the theoretical knowledge of the converter. The analysis of the gain characteristics and the soft switch implementation conditions are presented in Section III. Section IV introduces the proposed hybrid PWM-PFM control method and Section V shows the experimental results. Finally, Section VI concludes this paper.

II. CONVERTER WORKING PRINCIPLE AND THEORETICAL ANALYSIS

Fig. 1 shows the Boost-type CLLC converter under hybrid control. The converter consists of three parts, including an integrated Boost half bridge, high voltage side half-bridge rectification and a CLLC resonant tank.

In Fig. 1, specified arrows refer to the reference positive directions of the voltage and current. We define the current flowing through the switch as positive, and the current flowing through the body diode current as negative. The Boost circuit and half-bridge CLLC resonance circuit are integrated together through using common switch tubes, and the PV source and load share a common ground. At the low-voltage input, PV transfers energy to \( C_{b1} \) via the Boost circuit. \( C_{b1} \) transfers energy to \( C_{b2} \) through the CLLC resonant network. The input and output are superimposed together to form two series voltage sources to supply power to the load. The resonant inductors \( L_{r1} \) and \( L_{r2} \) are the transformer leakage inductance and external discrete inductance, respectively, \( C_{r1} \) and \( C_{r2} \) are the resonant capacitors, and the resonance parameters are the same. \( L_{m} \) is the excitation inductance of the transformer, and the transformer ratio is 1:1. \( S_1-S_2 \) are switching.
FIGURE 1. The boost-type CLLC converter.

MOSFETs, and their parasitic capacitance is $C_{oss_1}$-$C_{oss_2}$. $D_{o1}$-$D_{o2}$ is the body diode of the switch MOSFET. $C_1$-$C_4$ are bridge-arm capacitors, and $L_b$ is a Boost inductor. $U_{r1}$ and $U_{r2}$ are the primary and secondary resonance voltages, $i_{r1}$ and $i_{r2}$ are the primary and secondary resonance currents, and $u_{bc}$ and $u_{de}$ are the primary and secondary resonance tank voltages, $u_{tank_1}$ and $u_{tank_2}$, respectively. The circuit has two resonance frequencies $f_r$ and $f_m$. The first is the resonance frequency of the two elements $L_r$ and $C_r$, $f_r = 1/(2\pi((L_r C_r)^{1/2})$. The other is the three-element resonance frequency $f_m$, with the magnetizing inductance $L_m$ that participates in resonance, $f_m = 1/(2\pi((L_r + L_m) C_r)^{1/2})$. The normalized switching frequency is $f_n = f_s/f_r$. We define the $S_2$ duty cycle as $D$, the $S_1$ drive signal is complementary to $S_2$, and the secondary side control method is an uncontrolled rectifier. Fig. 2 shows the key waveforms for the converter under PWM-PFM hybrid control.

A. ANALYSIS OF THE WORKING MODE

Taking $f_s < f_r$ as an example to explain the operation process of the converter. In a single switching cycle $T$, there are 6 working modes $[t_0, t_6]$. The equivalent circuit for each mode is shown in Fig. 3. Each mode corresponds to (a)-(f) in Fig. 3. For example, Fig. 3 (a) represents Mode 1.

1) MODE 1 ($t_0$-$t_1$)

Before time $t_0$, $S_1$ and $S_2$ are commutated, and the parasitic capacitance $C_{oss_2}$ of $S_2$ is charged to $U_{b1}$ and $S_2$ is turned off. The $S_1$ parasitic capacitance $C_{oss_1}$ is discharged to 0, and $S_1$ is turned on. The Boost inductor current $\Delta i_{Lb} > 0$, enters the charging state. The secondary resonance current $i_{r2}$ is negative, and power is supplied to the load through $D_{o1}$, and $D_{o2}$ is in an off state. The primary resonance tank voltage is $u_{tank_1} = U_{b1}/2$, the secondary resonance tank voltage is $u_{tank_2} = U_{b2}/2$, and the excitation inductance clamp is located at $u_{tank_2}$, which does not participate in resonance, and the excitation current $i_{Lm}$ decreases linearly. At this time, the bridge capacitor $C_1$ is discharged to the CLLC resonance.

FIGURE 2. The key waveforms of the converter, (a) $f_s < f_r$ and (b) $f_s = f_r$.

FIGURE 3. Modal equivalent circuit diagram.
network, and the resonance network charges the bridge capacitor \( C_3 \), the bus capacitor \( C_{b1} \) charges \( C_2 \), and \( C_4 \) discharges to \( C_{b2} \). In this stage, \( L_{r1} \) and \( C_{r1} \) participate in resonance, and the resonance angular frequency is \( \omega_r \).

2) MODE 2 (\( t_1 \)-\( t_2 \))
In this stage, the primary switch does not operate, \( i_{Lb} \) rises linearly, at time \( t_1 \), \( i_{r1} \) is equal to \( i_{Lm} \), the secondary switch switches: the \( D_{o1} \) current is reduced to 0, and ZCS is turned off, the value of the side resonance current \( i_{L2} \) is greater than 0, and \( D_{o2} \) realizes ZCS on. \( C_1\cdot C_4 \) and \( C_{b1}\cdot C_{b2} \) are consistent with the previous modal states. In this stage, \( L_m \) is no longer clamped, and participates in resonance with \( L_{r1} \) and \( C_{r1} \), with a resonance frequency of \( f_m \). The resonance angular frequency \( \omega_m \) is \( \omega_m = \omega_r (m + 1)^{1/2} \), where \( m = L_{o1}/L_r \).

3) MODE 3 (\( t_2 \)-\( t_3 \))
This time period is the dead time for \( S_1 \) and \( S_2 \) commutation. In the dead time, \( i_{Lb} \cdot i_{r1} > 0 \), \( i_{Lb} \) and \( i_{r1} \) work together to charge the parasitic capacitance \( C_{oss1} \) of \( S_1 \) to \( U_{b1} \), and the parasitic capacitance \( C_{oss2} \) of \( S_2 \) is discharged to 0. \( S_2 \) has a body diode freewheeling before the driving signal arrives, \( S_2 \) achieves ZVS-ON.

4) MODE 4 (\( t_3 \)-\( t_4 \))
\( S_2 \) is turned on at \( t_3 \), the Boost inductor is discharged, and the inductor current \( i_{Lb} \) decreases linearly. The primary tank voltage is \( u_{tank1} = U_{b1}/2 \) and the secondary tank voltage is \( u_{tank2} = U_{b2}/2 \). The excitation inductor clamp is located at \( u_{tank2} \) and does not participate in resonance. The excitation current \( i_{Lm} \) increases linearly. At this time, the bridge capacitor \( C_3 \) is discharged to the CLLC resonance network, and the resonance network charges the bridge capacitor \( C_1 \); the bus capacitor \( C_{b2} \) charges \( C_4 \), and \( C_2 \) discharges to \( C_{b1} \).

5) MODE 5 (\( t_4 \)-\( t_5 \))
The primary switch does not operate at time \( t_5 \), \( i_{Lb} \) linearly decreases, \( i_{r1} \) and \( i_{Lm} \) are equal at time \( t_5 \), the secondary switch switches the current, the \( D_{o2} \) current is reduced to 0, and ZCS is turned off. The absolute value of the side resonance current \( i_{L2} \) is greater than 0, and \( D_{o1} \) turns on ZCS. \( C_1\cdot C_4 \) and \( C_{b1}\cdot C_{b2} \) are consistent with the previous modal states. In this stage, \( L_m \) is no longer clamped and participates in resonance with \( L_{r1} \) and \( C_{r1} \).

6) MODE 6 (\( t_5 \)-\( t_6 \))
This time period is the dead time for \( S_1 \) and \( S_2 \) commutation. In the dead time, the parasitic capacitance \( C_{oss1} \) of \( S_2 \) is charged to \( U_{b1} \), and \( S_2 \) is turned off; the parasitic capacitance \( C_{oss2} \) of \( S_1 \) is discharged to 0, and \( S_1 \) is turned on.

**B. ANALYSIS OF THE STATE-TRAJECTORY**

Ignoring the dead time, the state-trajectory diagram is utilized to guide the establishment of time-domain expressions for each parameter of the converter. With the changes in the normalized switching frequency \( f_n \), the converter has two typical operating states, \( f_n = 1 \) and \( f_n < 1 \). Fig. 4 shows a state-trajectory diagram for the two operating states. From Fig. 4(a), we can see that at \( f_n = 1 \), the state-trajectory diagram is a perfect circle and the converter has only two modes, mode 1 and mode 4. This is because the magnetizing inductance \( L_m \) does not participate in resonance, and only \( L_r \) and \( C_r \) complete the resonance once in a half cycle. At time \( t_0 \) in Fig. 4(a), \( i_{r1}(t_0) > 0 \), as the trajectory changes clockwise, the resonance voltage \( u_r \) reaches a maximum value when \( i_{r1} = 0 \). In addition, \( i_{r1}(t_0) = -i_{r1}(t_3) \). At time \( t_3 \), \( i_{r1}(t_3) < 0 \), if switch \( S_2 \) is turned on within this period of time, then \( i_r \) is negative and \( i_r \) continues to flow through the \( S_2 \) body diode, then \( S_2 \) will realize ZVS-ON.

From Fig. 4(b), the normalized switching frequency \( f_n < 1 \), we can see that the state trajectory is no longer a perfect circle, and there are 4 models in a cycle. Mode 1 and Mode 4 are quasi-sine waves in \( f_n < 1 \) and \( f_n = 1 \). Both Mode 2 and Mode 5 are approximated as a straight line because \( L_m \) participates in resonance with \( L_r \) and \( C_r \); the straight line means that the resonance voltage in these modes changes but the resonance current is approximately unchanged; the increment of the resonant current can be ignored when solving differential equations approximately. In Fig. 4(b), as the normalized switching frequency \( f_n \) continues to decrease, the duration of Mode 2 and Mode 5 will gradually increase. At time \( t_3 \), if switch \( S_2 \) is turned on within this period of time, ZVS-ON will be realized.

Without ignoring the increase in resonance current, based on the above mode and state trajectory analysis, we can obtain the KVL Equation (1) and the solution of the KVL equation in 4 modes: Equation (2), (3), (4) and (5). \( U_{in} \) represents the input voltage of the resonant tank and \( U_{out} \) represents the output voltage of the resonant tank. Taking mode 1 as an example, the input voltage of the resonant tank is \( U_{in} = U_{b1}/2 \) and the output voltage of the resonant tank is \( U_{out} = U_{b2}/2 \).

\[
\begin{align*}
U_{in} + L_r \frac{di_r}{dt} + u_r &= U_{out} \\
U_{in} + L_m \frac{di_{Lm}}{dt} &= U_{out} \\
C_r \frac{du_r}{dt} &= i_r
\end{align*}
\]
The polarity of the resonant current, the transmitted power will lead to an increase in the power output, which will increase the power circulating current and current stress in the converter, which will also increase the losses in the power devices and magnetic components and reduce the converter efficiency. Since the backflow power is not consumed by the load in the circuit, studying the relationship between the switching frequency and the backflow power is of great significance for the efficiency.

As shown in Fig. 2, the backflow power occurs in the positive and negative half cycles of the converter, respectively. According to the symmetry of the circuit operation, the backflow power is equal in the two half cycles. That is, in one cycle, the backflow power is twice the backflow power in a half cycle.

By setting the resonant current zero crossing angle to \( \alpha \), \( i_{r_1} (\alpha) = 0 \), and referring to Equation (2), we can obtain

\[
\alpha = \arctan \left( \frac{2i_r(t_0)Z_r}{2u_r(t_0) - (U_b2 - U_b1)} \right)
\]

By defining the backflow power as the product of the resonant tank input voltage and the resonant current of the opposite polarity, we obtain

\[
P_{\text{back}} = \frac{2}{T} \int_0^T u_{\text{tank}} \times i_{r_1}(t) \, dt
\]

\[
= \frac{U_b1}{T} [i_{r_1}(t_0) \sin \alpha + \cos(\alpha - 1) 2u_r(t_0) - (U_b2 - U_b1)]
\]

The transmission power under hybrid control is

\[
P = \frac{2}{T} \int_0^{T/2} u_{\text{tank}} \times i_{r_1}(t) \, dt
\]

\[
= \frac{U_b1}{T} [i_{r_1}(t_0) \sin \left( \frac{\omega_r T}{2} \right) + \cos \left( \frac{\omega_r T}{2} - 1 \right) 2u_r(t_0) - (U_b2 - U_b1)]
\]

Considering the correlation between the backflow power and the transmission power, we define the backflow power ratio \( M_{\text{back}} \) as

\[
M_{\text{back}} = \frac{P_{\text{back}}}{P} = \frac{\int_0^T u_{\text{tank}} \times [i_{r_1}(t)] \, dt}{\int_0^{T/2} u_{\text{tank}} \times i_{r_1}(t) \, dt}
\]

Referring to (9), to calculate the value of backflow power ratio, we must know the value of the unknown variable \( i_{r_1}(t_0) \), \( u_{r_1}(t_0) \), and \( \alpha \). According to the mode analysis for the converter and Fig. 2, we can see that \( i_{r_1}(t_0) = i_{L_m} \). At this time \( i_{r_1}(t_0) = i_{L_m} \), so \( i_{r_1}(t_0) = -i_{r_1}(\pi) = i_{L_m} \max \), given that \( i_{r_1}(t_0) \) can be calculated using Equation (10).

\[
\begin{align*}
M_{\text{back}} &= \frac{\int_0^T u_{\text{tank}} \times [i_{r_1}(t)] \, dt}{\int_0^{T/2} u_{\text{tank}} \times i_{r_1}(t) \, dt} \\
&= \frac{\int_0^{T/2} u_{\text{tank}} \times i_{r_1}(t) \, dt}{\int_0^{T/2} u_{\text{tank}} \times i_{r_1}(t) \, dt}
\end{align*}
\]

\[
\begin{align*}
\alpha &= \arctan \left( \frac{2i_r(t_0)Z_r}{2u_r(t_0) - (U_b2 - U_b1)} \right) \\
\end{align*}
\]

C. ANALYSIS OF THE BACKFLOW POWER
The backflow power is shown as the black portion in Fig. 2. When the input voltage of the resonant tank is opposite to the polarity of the resonant current, the transmitted power is negative and the power flows back to the power supply.
According to the analysis of Mode 1, the resonance current is equal to the excitation current at time \( t_1 \), \( i_{r1}(t_1) = i_{Lm}(t_1) \), so we can obtain \( u_{r1}(t_0) \):

\[
u_{r1}(t_0) = \frac{U_{b2}\pi Z_r}{4L_m} \left[ \frac{\cos(\omega_r t_1) + 1}{\sin(\omega_r t_1)} \right] + \frac{U_{b2} - U_{b1}}{2} \tag{11}\]

Referring to (10) and (11), we can obtain

\[
a = \arctan \left[ \frac{\sin(\omega_r t_1)}{1 + \cos(\omega_r t_1)} \right] \tag{12}\]

The value of the return power is related to \( a \), and the value of \( a \) is related to \( t_1 \), so a solution for \( t_1 \) is needed. Based on the mode analysis and the converter waveforms, we know that the waveform of the converter is symmetrical, so \( f(t_0) = -f(t_2) \), \( f(x) \) can be \( i_{r1,2} \), \( u_{r1,2} \), and \( i_{Lm} \).

Referring to (2) and (3), we eliminate the variables \( f(t_1) \), and substitute the solution for \( f(t_0) \) to obtain

\[
\begin{align*}
U_{b2} &= \left\{ \frac{1}{\sqrt{m+1}} \left[ \frac{\cos(\omega_r t_1) + 1}{\sin(\omega_r t_1)} \right] + \frac{2L_m}{Z_r \sqrt{m+1}} \right\} \\
\frac{(2U_{b1} - U_{b2})2L_m}{2U_{b2}Z_r \pi} &= \left\{ \left[ \frac{\cos(\omega_r t_1) + 1}{\sin(\omega_r t_1)} \right] + \frac{2L_m}{Z_r \pi} \right\} \\
&\times \cos(\omega_r t_2) - \sqrt{m+1} \sin((\omega_r t_2)) \\
&+ \left[ \frac{\cos(\omega_r t_1) + 1}{\sin(\omega_r t_1)} \right] \tag{13}\end{align*}
\]

We utilize the fsolve function in the mathematical tool MATLAB to calculate the numerical solution of Equation (9). The form of the equation solved by fsolve is \( F(X) = 0 \), the usage is

\[
x = \text{fsolve}(f(x), x_0) \tag{14}\]

where \( f(x) \) is the equation and \( x_0 \) is the initial value of the variable. The most important thing in the solution process is to determine the iterative algorithm and the initial values for the variables. The iterative algorithm utilizes the Trust-Region with a Dogleg method. The advantages of this algorithm are high reliability, strong effectiveness and convergence. The setting of the first initial value is obtained by cursor measurements in Simulink. The calculation results are shown in Table 1.

From the above calculations, we know that calculating the backflow power is complicated. One simple method is to obtain the backflow power by utilizing the simulation method. Compared with the numerical calculation method, the simulation method can simplify the calculation without calculating the initial values of the variables and considering the mutual coupling between variables, for accurately obtaining the backflow power. The simulation results are shown in Table 1.

From Table 1, we can analyze the following:

1) When \( f_n = [0.6, 1] \), as \( f_n \) decreases, \( M_{\text{back}} \) decreases and the backflow power decreases. The simulation results are similar to the calculation results, which also proves the correctness of the above theoretical analysis.

| \( f_n(N) \) | \( M_{\text{back}} \) by calculation (%) | \( M_{\text{back}} \) by simulation (%) |
|-------------|----------------|----------------|
| 1.0         | 8.92           | 8.770          |
| 0.9         | 8.84           | 8.681          |
| 0.8         | 8.53           | 8.488          |
| 0.7         | 8.14           | 8.059          |
| 0.6         | 8.03           | 7.036          |

2) When \( f_n = 0.6 \), the error between the calculation result and the simulation result is approximately 1%. This is because when we calculate \( i_{Lm}(t) \), we neglect the increment of \( i_{Lm}(t) \) in Mode 2, 5 in order to simplify the calculation. According to the state trajectory diagram, we can see that when \( f_n \) gradually decreases, the duration of Mode 2, 5 increases. This means that the error generated during the approximate solution for \( i_{Lm}(t) \) will gradually increase.

### III. GAIN CHARACTERISTICS AND IMPLEMENTATION CONDITIONS OF SOFT-SWITCHING

#### A. ANALYSIS OF THE GAIN CHARACTERISTICS

We set the input terminal voltage to be \( U_L \). From the power balance, we can know that the output voltage of the Boost part is \( U_{b1} = U_L / D \). The input voltage of the resonant tank is a square wave with a value of \( U_{\text{tank}1} = \pm U_{b1}/2 \). This square wave transfers energy to \( C_{L2} \) via the CLLC resonance network, forming two equivalent series voltage sources to supply power to the load together. Fig. 5 shows the first-harmonic-approximation (FHA) AC equivalent model realized by utilizing an FHA method.

![FIGURE 5. FHA AC equivalent model.](image)

To simplify the design of the resonance parameters, we define the primary and secondary resonance parameters to be consistent: \( C_{r1} = C_{r2} = C_r \), \( L_{r1} = L_{r2} = L_r \). The transformer turns ratio is \( n_1: n_2 = 1:1 \). From the FHA equivalent model in Fig. 5, we can know that the transfer function of the resonant network is

\[
M_{\text{cllc}}(s) = \frac{sL_m//\left(1 + \frac{1}{sC_r} + Z_e\right)\times Z_e}{\left(sL_r + \frac{1}{sC_r}\right)//\left(sL_m + \left(1 + \frac{1}{sC_r} + Z_e\right)\times Z_e\right)} \tag{15}\]

where \( Z_e = 4R_H/\pi^2 \) is the equivalent AC load. We define the circuit’s inductance ratio to be \( L_r/L_m = k \) with a figure of merit of \( Q = Z_r/Z_e \), where \( Z_r \) is the characteristic impedance \( Z_r = (L_r/C_r)^{1/2} \). Normalizing the switching frequency
where

\[
\begin{align*}
A(f_n) &= -f_n^2(1 + k) + k \\
B(f_n) &= f_n^4 - f_n^2(2 + k) + k
\end{align*}
\]

To improve the overall efficiency of the converter, the design should consider reducing the loss of the converter. Due to the soft switching characteristics of the CLLC converter, the conduction loss for the switching tube accounts for the largest proportion of the total converter loss. Ignoring the dead time, the fundamental and secondary component resonance currents \(i_{L1,\text{rms}}\) and \(i_{L2,\text{rms}}\) for the CLLC converter can be obtained by using the fundamental component method

\[
i_{L1,\text{rms}} = \frac{1}{2} \sqrt{\frac{U_0}{2} \left[ \left( \frac{T_s}{2L_{\text{m}}} \right)^2 + \left( \frac{\pi}{R} \right)^2 \right]}
\]

\[
i_{L2,\text{rms}} = \frac{U_0}{4} \sqrt{\left( \frac{5\pi^2 - 48}{12\pi^2} \right) \left( \frac{T_s}{L_{\text{m}}} \right)^2 + \left( \frac{1}{R} \right)^2}
\]

Increasing the field inductance reduces the effective value of the current and reduces the conduction loss. According to the working waveform of the CLLC converter, we can know that the switching tube is turned off at the peak value of the magnetizing inductance, which increases the magnetizing inductance and reduces the switching-off current of the switching tube. Fig. 6(c) shows the effective value of the current with respect to the value of \(L_{\text{m}}\) referring to (17) and (18). Given the input and output conditions in the figure, we can know that the effective value of the resonance current for the primary and secondary sides decreases with increasing excitation inductance \(L_{\text{m}}\). The input of the converter is a low voltage PV, so it is necessary to ensure that the voltage gain of the resonant tank is large. To obtain sufficient voltage gain for the converter and ensure the minimum effective value for the resonant current. According to Fig. 6(c), the excitation inductance value should be 250\(\mu\)H.

The DC gains \(M_{\text{dc}}\) for the converters described by (15) and (16) are

\[
M_{\text{dc}} = \frac{M_{\text{ecll}} \sqrt{2} U_{b1}/\pi}{U_{b1} |D - 0.5| + \sqrt{2} U_{b1}/\pi}
\]

The final converter gain \(G\) is

\[
G = \frac{U_H}{U_L} = \frac{\sqrt{2}/\pi(1 + M_{\text{ecll}}) + |D - 0.5|}{\sqrt{2}/\pi + |D - 0.5|D}
\]

Referring to (20), the three-dimensional gain curve for the converter in Fig. 6(d) can be obtained. In this article, the inductance ratio \(k\) is 0.2. Under the rated output power, \(Q\) is 0.055-0.221 according to different output voltages (200V-400V). Referring to (16) and (20), when the normalized switching frequency \(f_n\) is [0.6, 1], we know that the resonance gain is [0.976, 1.251]-[0.999, 1.526].

In the PFM-PWM hybrid modulation converter, at the duty cycle [0.3, 0.7], the normalized switching frequency is \(f_n = 0.6\) and the gain is [6.7, 2.8]. The fixed-frequency PWM gain is [5.6, 2.4] at a duty cycle of [0.3, 0.7]. Compared with the traditional Boost converter gain [3.3, 1.4], it has obvious advantages.
B. IMPLEMENTATION CONDITIONS OF SOFT-SWITCHING

According to mode analysis, the key to achieve ZVS soft switching is in the parasitic capacitance of the MOSFET being discharged to zero before turning on, and the anti-parallel diode being turned on. The key to achieving ZCS soft switching is the magnetizing inductance participating in resonance, \( i_{r1} = I_{Lm} \) and \( i_{r2} = 0 \), the secondary diode being cut off in reverse, and the current flow being zero. We named the commutation moment \( t_{S12} \) as \( S_1 \) is turned off and \( S_2 \) is turned on to analyze the soft switching implementation conditions:

\[
\begin{align*}
    i_{ZVS}(t_{S12}) &= i_{Lb}(t_{S12}) - i_{Lm}(t_{S12}) > 0 \\
    i_{ZCS}(t_{D21}) &= i_{r1}(t_{D21}) - i_{Lm}(t_{D21}) = 0 \\
    i_{ZCS}(t_{D12}) &= i_{r1}(t_{D12}) - i_{Lm}(t_{D12}) = 0
\end{align*}
\]  

(21)

Before \( t_{S12}, S_1 \) is turned on. The inductor current \( i_{Lb} \) rises linearly and reaches a peak value at \( t_{S12} \). Before time \( t_{S12} \), the converter is in the mode \( (t_1-t_2) \) phase. The circuit is in the resonance state of the three resonance elements. The resonance current and the excitation current are approximately equal, and \( i_{r1} (t_{S12}) \) can be calculated from the excitation current. The excitation current decreases linearly before \( t_{S12} \), and reaches a minimum value at \( t_{S12} \).

\[
    i_{r1}(t_{S12}) \approx i_{Lm, min} = -\frac{U_{b2} \pi}{4L_m} \tag{22}
\]

When the inductance \( L_b \) is infinite and the inductor current is continuous, the average inductor current \( i_{Lb} \) is approximately equal to the input current. We can determine the average boost inductor current \( i_{Lb} \) according to the converter output power \( P_o/U_L \). The peak inductor current can be obtained from

\[
    i_{Lb, max} = i_0 + \frac{\Delta i_{Lb}}{2} = P_0 \frac{U_L(1-D)}{2f_is_{Lb}} \tag{23}
\]

Based on (21), (22), and (23), we can obtain

\[
    i_{ZVS}(t_{S12}) = i_{Lb}(t_{S12}) - i_{Lm}(t_{S12}) = \frac{P_0}{U_L} + \frac{U_L(1-D)}{2f_is_{Lb}} + \frac{U_{b2} \pi}{4L_m} \tag{24}
\]

The Boost inductance parameter value in this paper is large, and \( i_{ZVS}(t_{S12}) \) is greater than zero. Therefore, the switch \( S_2 \) can always obtain ZVS-ON. The secondary side adopts uncontrolled rectification. When \( f_n < 1 \), the magnetizing inductance participates in resonance, \( i_{r1} = i_{Lm} \), and \( D_{r1,2} \) can realize ZCS. When \( f_n = 1 \), the converter works in a quasi-resonant state, the secondary diode turns on and off within the dead time, and ZCS can be realized.

To achieve ZVS switching, the MOSFET output capacitor \( C_{oss} \) must be fully charged and discharged during the dead time. The minimum commutation current required in the dead time can be calculated using (25).

\[
    i_{dead} \geq 16C_{oss}f_is_{Lm} \tag{25}
\]

IV. CONTROL METHOD

Common MPPT methods include constant voltage tracking (CVT), the perturbation and observation method (P&O), and incremental conductance method (IC). Compared with CVT, the P&O and IC methods have higher tracking efficiency. If the P&O or IC method is used, under hybrid control, when the system starts to run, one first controls the duty cycle to track the maximum PV power point. After repeated adjustment, the PV output power is maximum, and then one controls the switch frequency to stabilize the output voltage. However, when the switching frequency changes, the gain of the converter changes, which will affect the input current, leading to failure of the MPPT, and failure of the constant voltage output. If the switching frequency changes according to the output voltage, circuit gain and current duty cycle, when the PV reaches the maximum power point, the system can run stably. However, when the external conditions are constant (to a certain extent), the P&O and IC methods will adjust frequently to determine the current maximum power point for the PV. The frequent adjustment of these two methods results in frequent changes in the duty cycle, and the delay in the actual controller calculation will result in fluctuation of the output voltage, resulting in the degradation of the output voltage and power quality. In fact, in most cases, the chain reaction produced by these two methods will cause the system to not work properly at all. Even if we reduce the sensitivity and add a time delay for the P&O or IC methods, the converter will continue to oscillate, resulting in failure of the constant voltage output and MPPT.

According to the PV single-diode equivalent model of solar panels

\[
    I_{pv} = I_{ph} - I_0 \left\{ \exp \left[ q \left( \frac{U_{pv} + IR_s}{nkT} \right) \right] - 1 \right\} \tag{26}
\]

where \( I_{pv} \) is the output current of the PV cell panel, \( I_0 \) is the saturation current, \( q \) is the electronic charge constant \( (q = 1.6 \times 10^{-19} C) \), \( U_{pv} \) is the output voltage of the PV cell, \( n \) is the fitting coefficient for the diode characteristics, \( k \) is the Hertzian constant \( (k = 1.38 \times 10^{-23} J/K) \), and \( T \) is the absolute ambient temperature. Referring to (26), MATLAB was used to obtain the PV output characteristic, as shown in Fig. 7.

From Fig. 7, we can see that the PV output power is more affected by light intensity than by temperature. If the difference in PV characteristics generated by temperature is ignored, the open-circuit voltage \( U_{oc} \) is considered to be approximately constant, and the PV output power is only related to the incident light. The CVT method is an approximate MPPT method. When the temperature is constant, the voltage corresponding to the maximum power point of the PV cell under different lighting conditions falls almost on the same working voltage \( U_m \). There is an approximately proportional relationship between the operating voltage \( U_m \) and the open circuit voltage \( U_{oc}, U_{oc} = K \times U_{oc} \), and the ratio \( K \) is 0.71-0.8 [20]. If the output voltage of the PV cell is always kept equal to this working voltage \( U_m \), then the PV system can keep working at the maximum power point. When
using CVT, a constant voltage can be achieved. Therefore, the MPPT control method used in this article is CVT.

The control block diagram is shown in Fig. 8. The specific implementation of control is as follows: At rated power, the DSP first reads the PV operating voltage $U_m$ and the reference output $U_{ref}$, $U_{ref} / U_m = G$, where G is a function of duty cycle and frequency, $G = f(f_n, D)$; according to our analysis, we can know when $f_n = 1$, where the converter shows the highest efficiency. We first assume $f_n = 1$, and obtain the duty cycle $D$ according to the function $G$. If $D$ is not within $[0.3, 0.7]$, one proves that the current converter gain is not big enough and that the resonant tank is needed to increase the additional voltage gain. Then, $f_n = f_n - \Delta f$, and the duty cycle D is calculated, following which one can determine whether $D$ is within the interval, with the program looping until this condition is satisfied. However, due to the loss of voltage gain due to the dead time and the presence of the on-state impedance of the converter, the output voltage $U_H < U_{ref}$.

Considering this situation, we set a time $T_L$ for this program cycle; when $t = T_L$, the control loop changes, and the control loop becomes a single-voltage closed-loop PI control. The reference voltage $U_{ref}$ is compared with the output voltage $U_H$ and then sent to the PI controller to change the switching frequency to achieve the desired output voltage.

V. EXPERIMENTAL VALIDATION

A 500W experimental prototype is implemented and tested to verify the performances of the converter and the effectiveness of the hybrid control. The experimental test bench is shown in Fig. 9. The oscilloscope used in the laboratory is Yokogawa DL850E digital recorder. The active current probe model is Yokogawa 701933, and the voltage probe model is Yokogawa 700929. $S_1$-$S_2, D_{o1}$-$D_{o2}$ models are FCH104N60F, drivers are opt coupler FOD3184S. Magnetic core components: Boost inductance is ferrosilicon magnetic ring inductance, high frequency transformer is PC40EE55, and resonance inductance is PC40PQ35. The control algorithm is implemented in TMS320F28335 from Texas Instruments. The final determined converter parameters are shown in Table 2. Different control conditions and hybrid control are tested.

Fig. 10 shows the waveforms for different duty cycles and switching frequencies. Fig. 10 (a) shows the $u_{tank,1,2}, i_{S_1,2}, u_{D_{o1,2}}$ and the drive($S_1,2$) waveforms at $D = 0.5$ and $f_n = 0.9$. It is obvious that $S_2$ realizes ZVS and $D_{o1,2}$ realizes ZCS during the turning-on process; the soft switch is consistent with the above analysis. The low-voltage terminal is connected to a 47V/4.59A DC power supply; the high-voltage terminal is connected to a 152 $\Omega$ resistor load, and the output voltage is 179.07 V. The actual converter voltage gain is 3.81, and the conversion efficiency is 97.78%. Based on the theoretical voltage gain analysis, the theoretical gain at $D = 0.5$ and
TABLE 2. Parameters of the converter.

| Parameter | Value | Parameter | Value |
|-----------|-------|-----------|-------|
| $U_I$     | 20–80V| $C_1$–$C_4$| 470μF |
| $U_H$     | 100–400V| $C_{51}$–$C_{52}$| 1 880μF |
| $P_{in}$  | 500 W | $C_5$–$C_7$| 970mF |
| $L_m$     | 255μH | $n_1$, $n_2$| 15:15 |
| $L_{11}$  | 50μH  | $f_1$ | 22.8kHz |
| $L_{12}$  | 50μH  | $f_1$ | 13.7–22.8kHz |
| $I_o$     | 1 000μH| $t_{dead}$| 1us |

$f_n = 0.9$ is 4.076. There is an error between the theoretical and the actual gain of 0.2 because of the dead time gain loss and the on-state impedance of the converter.

Upon changing the control condition, Fig. 10 (b) shows the waveforms at $D = 0.5$ and $f_n = 0.8$. We can see that $S_2$ realizes ZVS-ON, and $D_{o1}$, $D_{o2}$ realize ZCS. The input voltage is 48 V, the output voltage is 187.43V. The actual converter gain is 3.90, which is greater than 3.81 at $D = 0.5$ and $f_n = 0.9$. The realization of soft switching and the change in the voltage gain are consistent with the analysis. Under this control condition, the conversion efficiency is 96.3%, which is also extremely high.

To verify the conversion efficiency of the converter and the realization of the soft switching under heavy load condition, Fig. 10(c) shows the $u_{tank1}$, $i_{S1}$, $i_{D1}$, and the drive($S_{1,2}$) waveforms at 504W. It is obvious that the $S_2$ realizes ZVS and $D_{o1,2}$ realize ZCS during its turning-on process, and the efficiency is 94.1%.

When the PV voltage is relatively low, the converter needs to provide a large voltage gain to ensure that the output voltage meets the requirements. To verify this, Fig. 10(d) shows the scope waveforms at an input DC voltage of 15V. At this low input voltage, the output voltage is 98.2V at $D = 0.3$ and $f_n = 0.6$, the converter voltage gain is 6.55, which means that if the low voltage input $U_L$ is within [20, 80]V, the converter can raise $U_L$ to be large enough to meet the output voltage $U_H = [200, 400]$V. As shown, $S_2$ realizes ZVS-ON, and $D_{o1}$, $D_{o2}$ realize ZCS, and the efficiency is 92.75%, which means that under extreme control conditions, the converter also shows a good conversion efficiency.

To verify the effectiveness of the hybrid control, Fig. 11 shows the input $i_{pv}$, $u_{pv}$ and output $u_H$, $i_H$ waveforms. In Fig. 11 (a), the load suddenly decreases, the input current $i_{pv}$ is reduced from 3A to 1A within 1s, and the waveforms for the input voltage $u_{pv}$ and output voltage $u_H$ maintain a straight line. This means that the PV is always in a MPPT state, and the hybrid control keeps the output voltage $u_H$ constant to meet the reference voltage $u_H_{-ref}$. Fig. 11(b) shows the waveforms for an increased load. The output voltage for both modes is stable at 130V, which verifies the effectiveness of the hybrid control.

When power flows forward, the efficiency of the converter at different powers is shown in Fig. 12. Because the resonant tank voltage in this paper is greater than 1, the selected excitation inductance $L_m$ is small in order to obtain a higher
AC gain. In addition, the excitation current $i_{Lm}$ is mainly determined by $L_m$, the excitation inductance is small, and the excitation current is large [18]. Therefore, at light load for the converter, $i_{Lm}$ occupies a high proportion of loss and the converter efficiency is low. However, as the rated power of the converter increases, the input current increases, the proportion of $i_{Lm}$ loss decreases, and the efficiency increases. At a light load of 200W for the converter, the maximum efficiency is 97.78%. After 200W, the rated power and input current increase because of the hard switching of $S_1$, and the switching loss for $S_1$ keeps on increasing, so the efficiency curve decreases accordingly.

VI. CONCLUSION

Aimed at PV power generation systems with low input voltage and fluctuating power, this paper studies a Boost-type soft switching and high gain CLLC converter under PWM-PFM hybrid control for PV power generation; this system solves the problem of low utilization of PV energy and large switching loss for a half-bridge CLLC converter under fixed-frequency PWM control. In addition, the converter also shows the following advantages:

1) Under hybrid control, the boost ratio is high, which avoids the limit duty cycle and wide switching frequency of traditional control methods. The system also supports a simplified magnetic design;

2) Uncontrolled rectification of the secondary side saves the drivers of the secondary side, which can reduce the hardware cost;

3) The influence of a change in switching frequency on the backflow power in the converter is analyzed, which makes up for the lack of research into backflow power in resonant converters;

4) In PV grid-connected power generation systems, the converter shows both MPPT and regulated output characteristics, which frees the control freedom for grid-connected inverters, facilitating the provision of more grid services.

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