

Direct AC-AC Converter Based on Odd Symmetrical No-Differential AC Chopper Legs With Unipolar Modulation Strategy

Tianpeng Du, Dongbo Guo, Mingyue Li, Yonglin Duan, Zhongchen Pei and Guanyu Yan

EasyChair preprints are intended for rapid dissemination of research results and are integrated with the rest of EasyChair.

January 3, 2023

Direct AC-AC Converter Based on Odd Symmetrical No-differential AC Chopper Legs With Unipolar Modulation Strategy

Abstract—This paper proposes a single-phase bipolar type pulsewidth modulation (PWM) direct ac-ac converter with unipolar modulation strategy. This converter is composed of two odd symmetrical two-level no-differential ac chopper legs, which can work in non-inverting and inverting modes for the utility voltage sag or swell compensation. The number switches working at high frequency in each operating mode are minimized thanks to the proposed unipolar modulation strategy, so the total switching losses can be effectively reduced in theory, and the efficiency of the converter is improved. In addition, capacitor symmetry can be realized in the positive and negative half cycle of the input voltage, and the converter has no commutation problems. Finally, on the basis of theoretical analysis, the experimental results under different working conditions are also provided to verify the correctness of the theoretical analysis and the superiority of the proposed topology and modulation strategy.

Index Terms—Direct AC-AC converter, bipolar voltage gain, unipolar modulation strategy, odd symmetrical, switching losses.

I. INTRODUCTION

RECENTLY, with the popularity of renewable energy sources such as wind power and photovoltaics, the voltage fluctuation in the power grid has been exacerbated[1]. To meet the requirements of sensitive loads for voltage stability [2], AC-AC converters are usually used to improve the power quality and provide a stable AC voltage for the loads. Commonly used AC-AC converters can be divided into three types: indirect converters (AC-DC-AC) [3]-[6], matrix converters [7]-[10], direct AC-AC converters [11]-[13]. For the AC-DC-AC converter, two-stage power conversion is required, the existence of the DC link increases the volume and loss of the converter, reduces the conversion efficiency, increases the maintenance cost; For the matrix converter, although the amplitude and frequency of the voltage can be adjusted at the same time, its circuit structure is complex, and the transmission efficiency and voltage gain are relatively low. The direct PWM AC-AC converter has no DC link and only needs a single power conversion, so the efficiency and power density of the converter can be improved, and it has outstanding advantages when only the voltage amplitude needs to be adjusted.

The buck, buck-boost, and Cuk converters proposed in the literature [14]-[16] use fewer switching devices and passive components, which have the advantages of simple structure and high conversion efficiency. However, due to the use of bidirectional switches, during the commutation process switches may experience voltage spikes or current spikes, seriously affecting the reliability of the converter. The switching cell (SC) connected in series by MOSFET and diode [17]-[19], which solves the commutation problem through additional coupling inductors, but the additional diodes and inductors may increase the loss of the converter. The biggest defect of the above converter [14]-[19] is that it can only output unipolar voltage and cannot solve voltage swell and sag at the same time. Therefore, it is necessary to develop a converter that can output bipolar voltage [20]-[30]. In [20]-[22], Z-source topology is adopted to obtain wide-range bipolar voltage gain, but the converter suffers from commutation problems due to the use of bidirectional switches [23-26]. In [20]-[21], a special commutation strategy was applied, but reliable and safe commutation could not be achieved. In [22], an RCD absorption circuit was used to suppress the voltage spike on the switch, which not only complicates the circuit structure but also increases the extra loss. In [27]-[28], the proposed topology adopts fewer active switches and passive components, which can not only output bipolar voltage but also has no commutation problem. However, the converter proposed in [27]-[28] has the problem of the input and output having no common ground. In addition, the converter proposed in [27] has an asymmetric bipolar operating mode. The converter proposed in [29]-[30] not only overcomes the commutation problem but also has a symmetrical bipolar operating mode, and also ensures that the input and output have the same ground, the defect is that the modulated signals are all high-frequency signals. In the working circuit, more switches are working at high frequency, which increases the switching loss.

1

The converter proposed in this paper has symmetrical bipolar working mode, which can effectively solve the problem of voltage sag and swell. In addition, the converter has no commutation problems, and there is no loss of buffer circuit or special commutation strategy. The most prominent advantage of this paper is the proposed modulation strategy, which uses fewer high-frequency modulation signals. Under resistive load, the converter has only one switch operating at high frequency per half-wave and there is no dead time. Under inductive load, there are only two switches operating at high frequency. In addition, the converter uses only one inductor (L_f) , which effectively reduces stray resistance losses and core losses. All the above characteristics of the converter are verified on the built 1kW prototype. This paper is arranged as follows. The description of the proposed topology and PWM modulation principles are arranged in section II. Section III investigates the working process of the proposed converter. Section IV presents the design considerations of the key components in the

converter. The loss of the converter is calculated theoretically in the section V. The experimental results illustrating the performance of the converter are presented in section VI. Finally, the conclusions are highlighted in section VII.

II. DESCRIPTION OF THE PROPOSED TOPOLOGY AND PMW MODULATION PRINCIPLES

A. Topology of the Proposed Converter

The proposed converter consists of two completely odd symmetry two-level non-differential ac chopper legs, which is different from the converter proposed in reference [29]. However, similar to the converter mentioned in reference [29], each ac chopper leg of the proposed converter is also composed of four fully controlled switches (such as IGBT) and one absorption capacitor, as shown in Fig.1. By adjusting the direction of switches and optimizing its modulation strategy, the number of switches working at high frequency in each operating mode is reduced, the switching loss is reduced theoretically. Another advantage is that input capacitor C_{in} is always connected in parallel with one of the absorption capacitor symmetry is achieved.





The converter adopts eight IGBT, among which S_1 ~ S_4 can be selected according to load type and working mode. For resistive loads, the switches S_2 and S_3 can employ diodes instead of full controlled switches when the output voltage of the converter is in phase with the input source voltage, diodes also can be used to replace the full controlled switches S_1 and S_4 when the output voltage of the converter is out of phase with the input source voltage. At this time, there is only one switch working at high frequency in the positive and negative half cycle of input voltage, which not only simplifies the converter structure but also does not exist dead time. For inductive loads, it needs to be replaced with full controlled switches IGBT.

B. Unipolar Modulation Strategy

The unipolar modulation strategy adopted by the converter proposed in this paper is shown in Fig. 2. The biggest advantage of this modulation strategy is that it has fewer high-frequency modulation signals. It can be seen that SF1, SF₂, SF3, and SF₄ always operates at low frequency, so for IGBT, the requirement for switching frequency can be ignored to reduce the conduction voltage drop, thereby reducing the conduction loss.



2

III. OPERATION OF THE PROPOSE D CONVERTER

According to the gain of the converter output voltage, the converter has two working modes, one is that the output voltage and the input voltage are in phase, and the other is that the output voltage and the input voltage are out of phase. The relationship between the input voltage and the output voltage is shown in formula (1). Where U_{in} is defined as the amplitude of the input line frequency ac voltage (50Hz), U_{out} is defined as the amplitude of the amplitude of the output line frequency voltage (50Hz), and *d* is defined as the time interval when switches are turned on during one switching period. The range of *d* is from 0 to 1.

$$J_{out} = \pm dU_{in} \tag{1}$$

A. Operation process in mode 1

In this mode, the converter has two operating stages, as shown in Fig. 3. Under resistive load, there is no reverse power flow. So, two diodes D₂ and D₃ can replace the full controlled switches S₂ and S₃ respectively. When $U_{in}>0$, only switch S₁ operating at high frequency, the switches SF₁, SF₄, SF₂, and S₄ are in the always on state, and the switch SF₃ is in the always off state.



(4)

Stage 1 [t1-t2]

During $[t_1-t_2]$, S₁ is turned on, the output filter inductor (L_j) and load are charged from the input ac source through SF₁ body diode, S₁, S₄ body diode and SF₄, the working process of the converter can be represented in Fig. 4(a). At this time, the inductor current I_L increases linearly to the maximum value and the converter is in active mode, the voltage and current relationship as flows:

$$\begin{cases}
L_{f} \frac{dI_{L}}{dt} = U_{in} - U_{out} \\
L_{f} \frac{(I_{LMax} - I_{LMin})}{d \cdot T_{SW}} = U_{in} - U_{out}
\end{cases}$$
(2)

Stage 2 [t₂-t₃]

During $[t_2-t_3]$, S₁ is turned off and S₂ body diode (D₂) forward bias, the inductor current I_L forms a closed loop through the load, S₄ body diode, SF₄, SF₂, and D₂. The energy stored on the inductor is released through this loop, so the inductor current linearly decreases as shown in Fig. 3. During this time interval, the converter is in the freewheeling mode (passive mode) and the voltage and current relationship as flows:

$$\begin{bmatrix}
L_f \frac{dI_L}{dt} = -U_{out} \\
L_f \frac{(I_{LMin} - I_{LMax})}{(1 - d)T_{SW}} = -U_{out}
\end{bmatrix}$$
(3)

Based on formula (2) and (3), the relationship between the output voltage and the input voltage is shown in formula (4).



Fig. 4. Two effective switching states of model under pure resistive load. (a) *stage* 1. (b) *stage* 2.

Under resistive-inductive load, there will be a reverse power flow. When $U_{out}>0$, $i_o<0$, if S₁ turns on and S₂ turns off, the current will form a closed loop through S₁ body diode, SF1, SF₄ body diode, and S₄, as shown in Fig. 5(a); if S₁ turns off and S₂ turns on, current will form a closed loop through S₂, SF₂ body diode, SF_4 body diode, and S_4 , as shown in Fig. 5(b). When the power flows forward, the flow path is the same as that under resistive load, so it will not be described.



Fig. 5. Two effective switching states of model under resistive-inductive load. (a) stage 1. (b) stage 2.

B. Operation process in mode 2

In this mode, the converter also has two operating stages, and the output voltage of the converter is out of phase with the input voltage. Under resistive load, two diodes can also be used as switches S_1 and S_4 . Similar to the analysis when the converter operates in mode 1, the positive half wave of the input voltage is still taken as an example for analysis in mode 2. In this mode, only switch S_2 operating at high frequency, the switches SF_1 , SF_2 , SF_3 , and S_3 are always on, and SF_4 , S_4 are always off.



Stage 1 [t1-t2]

The switch S_2 is on state during this period, the inductor and load are charged from the source through SF_3 , S_3 body diode, S_2 and SF_2 body diode. The converter is also in active mode, and the working process is shown in Fig. 7(a). At this time, the voltage relationship of the working loop is shown in formula (5). (7)

$$\begin{cases} -L_f \frac{dI_L}{dt_1} = U_{in} + U_{out} \\ -L_f \frac{(I_{LMax} - I_{LMin})}{dT_{SW}} = U_{in} + U_{out} \end{cases}$$
(5)

Stage 2 [t2-t3]

When S₂ turns off, the energy stored on the inductor is released, the working process is shown in Fig. 7(b). At this time, the voltage relationship of the working loop is shown in formula (6).

$$\begin{cases} L_f \frac{dI_L}{dt_2} = -U_{out} \\ -L_f \frac{(I_{LMin} - I_{LMax})}{(1 - d) \cdot T_{SW}} = -U_{out} \end{cases}$$
(6)

Based on formula (5) and formula (6), the relationship between the output voltage and the input voltage is shown in formula (7).



Fig. 7. Two effective switching states of mode2 under pure resistive load. (a) stage 1. (b) stage 2.

Under resistive-inductive load, when $U_{out} < 0$, $i_o > 0$, if S₂ turns on and S_1 turns off, the working circuit is shown in Fig. 8(a); if S₂ turns off and S₁ turns on, the working circuit is shown in Fig. 8(b).



and freewheeling state (passive mode). In order to ensure that the converter always works in CCM (current continuous mode), it is necessary to ensure the
$$I_{Lmin} > 0$$
 under the freewheeling state. According to the principle of conservation, the energy W_L released by the inductor in the freewheeling state is equal to the energy W_{load} consumed by the load.

ar

it

$$\begin{cases} W_L = \frac{1}{2}LI_{Lmax}^2 - \frac{1}{2}LI_{Lmin}^2 \\ W_{load} = U_{out} \cdot I_{out} \cdot (1-d) \cdot T_{SW} \end{cases}$$
(8)

the principle of conservation, the energy W_L

Based on the formulas (2)-(8), the minimum value of filter inductor shall meet the formula (9), where r is the maximum current ripple.

$$L \ge \frac{U_{out} \cdot (1-d)}{r \cdot f_s \cdot I_{out}} \tag{9}$$

B. Design of output filter capacitors

In order to ensure the quality of the output waveforms, an output filter capacitor is designed to limit voltage ripple. Based on maximum voltage ripple k_u , the output filter capacitor shall meet the formula (10). At the same time, to absorb the stray energy in the circuit, a clamping capacitor is added to each ac chopper leg.

$$C_f \ge \frac{i_{0\max}}{k_u f_s U_{out}} \tag{10}$$

Based on the above design considerations, some key parameters of the proposed converter are shown in Table I below.



Fig. 8. Two effective switching states of mode2 under resistive-inductive load

When $U_{in} < 0$, the analysis method is the same as when $U_{in} >$ 0. To avoid repetition, the analysis is no longer carried out. It can be seen from the above analysis that no matter under resistive load or inductive load, the current loop always exists, so there is no commutation problem in the converter. Moreover, the converter has the ability to output bipolar voltage gain, which can simultaneously address the problem of voltage sag and swell in the power grid.

IV. DESIGN CONSIDERATIONS OF KEY COMPONENTS OF CONVERTER

Under the two operating modes, the output filter inductor has only two working states in each switching cycle: active state

A. Design of output filter inductors

TABLE I	
EXPERIMENTAL PARAMETERS	
Parameters	Values
Input voltage U _{in}	[120-220Vrms]/50Hz
Output power P_O	1kW
Output voltage Uout	150 Vrms
Switching frequency $f_{\rm S}$	10kHz
Inductance L_f	0.5mH
Output capacitor C_f	10µF
Capacitors $C_1 \sim C_2$	10µF
Output load {R, R&L}	$\{20\Omega, 20\Omega\&25m\}$

V. CONVERTER LOSS THEORETICAL ANALYSIS

The converter power loss mainly includes IGBT loss and stray resistance loss. IGBT loss includes conduction loss and switching loss. Based on the IGBT datasheet, the function relation between power loss and junction temperature, voltage, current, and other parameters are obtained by the numerical fitting method, and then IGBT loss is analyzed and calculated.

A. Theoretical calculation of conduction loss

The conduction loss can be calculated according to the formula (12).

$$P_{S} = \frac{1}{T} \int_{0}^{T} V_{CE} I_{C} dt \qquad (12)$$

After linear fitting of IGBT output characteristic curve, the relationship between on-state voltage drop (V_{CE}) and collector current (I_C) can be expressed as follows:

$$V_{CE} = V_{CEO} + I_C R_{CE} \tag{13}$$

Where V_{CEO} is the threshold voltage of IGBT after linear fitting, R_{CE} is the forward on-resistance of IGBT after linear fitting. These two fitting parameters are affected by temperature change, which can be expressed as follows:

$$\begin{cases} V_{CEO} = V_{CEO_{25^{\circ}C}} + K_{V_{T}} (T_{J} - 25^{\circ}C) \\ R_{CE} = R_{CE_{25^{\circ}C}} + K_{R} (T_{J} - 25^{\circ}C) \end{cases}$$
(14)

Where K_{VT} and K_R are the junction temperature coefficients of on-resistance and threshold voltage respectively, and T_J represents the actual junction temperature. Based on the formulas (12)-(14), The conduction loss can be expressed as follows:

$$P_{S} = V_{CEO}I_{Cavg} + R_{CE}I_{Crms}^{2}$$
(15)

Where I_{Cavg} represents the average value of collector current and I_{Crms} represents the effective value of collector current. The calculation process of diode conduction loss is similar to that of IGBT, and its conduction loss can be expressed as the formula (16). The total conduction loss is shown in formula (17).

$$P_F = V_{FO}I_{Favg} + R_F I_{Frms}^2$$
(16)

$$P_{\rm T} = P_{\rm F} + P_{\rm S} \tag{17}$$

B. Theoretical calculation of switching loss

The IGBT switching process is affected by many factors such as gate resistor and junction temperature. The switching loss provided in the datasheet is tested under rated conditions, so it can not be used directly. It should be corrected according to the actual working conditions. The energy loss of IGBT in one switching cycle is shown in formula (18).

$$E_{ts} = E_{ts_i_c}^* \cdot K_{T_i} \cdot K_{R_c} \cdot K_{V_{CF}}$$

$$\tag{18}$$

The E_{ts} includes IGBT switching loss and diode reverse recovery loss, and $E^*_{ts_ic}$ is the corresponding energy loss under different collector current. The values of junction temperature influence coefficient (K_{TJ}), voltage influence coefficient (K_{VCE}), and grid resistance influence coefficient (K_{RG}) are shown in formula (19):

$$\begin{cases} K_{V_{CE}} = \left(\frac{V_{CE_red}}{V_{CE_ref}}\right)^{N_{Tr_VCE}} \\ K_{T_{j}} = \left[1 - N_{T_{j}}(T_{J_ref} - T_{J_real})\right] \\ K_{R_{G}} = \left(\frac{R_{G_real}}{R_{G_ref}}\right)^{N_{Tr_R_{G}}} \end{cases}$$
(19)

Where T_{J_ref} , V_{CE_ref} , and R_{G_ref} are junction temperature, collector-emitter voltage, and gate resistor under test conditions respectively; the T_{J_real} , V_{CE_real} , and R_{G_real} are junction temperature, collector emitter voltage, and gate resistor in the actual circuit respectively.

The values of N_{TJ} , N_{Tr_VCE} , and N_{Tr_RG} are obtained by numerical fitting the E_{ts} - T_j , E_{ts} - V_{CE} , and E_{ts} - R_G curves in the datasheet respectively. Assuming that the switching frequency is f_s , the switching loss power can be expressed as follows:

$$P_{ts} = f_s \cdot E^*_{ts_i_c} \left(\frac{V_{CE_real}}{V_{CE_ref}} \right)^{N_{Tr_VCE}} \cdot \left(\frac{R_{G_real}}{R_{G_ref}} \right)^{N_{Tr_R_G}}$$
(20)

$$\cdot [1 - N_{T_J} \left(T_{J_ref} - T_J \right)]$$

Similarly, the turn-on loss is shown in formula (21).

$$P_{ON} = f_s \cdot E^*_{on_i_c} \left(\frac{V_{CE_real}}{V_{CE_ref}} \right)^{N_{Tr_VCE_om}} \cdot \left(\frac{R_{G_real}}{R_{G_ref}} \right)^{N_{Tr_R_{G_om}}}$$
(21)

$$\left[1 - N_{T_J} \left(I_{J_ref_on} - I_J\right)\right]$$

The turn-off loss is shown in formula (22).

$$P_{OFF} = f_{s} \cdot E_{off_i_{c}}^{*} \left(\frac{V_{CE_real}}{V_{CE_ref}} \right)^{N_{Tr_V_{CE_eff}}} \cdot \left(\frac{R_{G_real}}{R_{G_ref}} \right)^{N_{Tr_R_{G}_eff}}$$
(22)

$$\cdot [1 - N_{T_{J}} (T_{J_ref_off} - T_{J})]$$

Fig. 9 shows the switching loss corresponding to different collector currents under rated conditions provided in the datasheet. The Eon includes turn-on loss and diode reverse recovery loss, Eoff is the turn-off loss, and Ets* is the total switching loss.



Fig. 9. Switching energy loss as a function of collector current ($V_{GE} = 0/15V$, $T_J = 175^{\circ}C$, $R_G = 5\Omega$, $V_{CE} = 400V$)

C. The converter power loss distribution

In order to determine the various losses of the proposed converter, the parameters for loss analysis are given in Table II. The operating conditions assume that U_{in} =200 Vrms, U_{out} =150Vrms, P_o =1000W, f_s =10k, and T_J =75°C. PSIM software

can be used to measure the collector-emitter voltage (V_{CE}) and the collector current (I_C) of the switch, and then the conduction loss can be calculated through formulas (12) to (17). Combining formula (20), Table II and Fig. 9, the switching loss can be calculated.

TABLE II	
PARAMETERS FOR LOSS ANALYSIS	
Parameter	Value
IGBT	IKW75N60T
f_s	10 kHz
$V_{CEO_{25^{\circ}C}}, R_{CE_{25^{\circ}C}}(I_C < 15A)$	0.6 V , $20 \text{m}\Omega$
$V_{CEO_25^{\circ}C}, R_{CE_25^{\circ}C}(I_C>15\text{A})$	$0.75~V$, $8~m\Omega$
K_{VT} , K_R	-0.001, 0.05
$V_{FO_{25^{\circ}C}}, R_{F_{25^{\circ}C}}(I_C < 15A)$	$0.65V$, 22 m Ω
V _{CE_ref} , R _{G_ref} , T _{J_ref}	400V, 5Ω, 175°C
N_{Tr_VCE} , N_{Tr_RG} , N_{TJ}	1.6 , 0.46 , 0.00162
ESR of capacitor C_f	2.5 mΩ
Parasitic resistances of L_f	$140 \text{ m}\Omega$
P _n P _Z PL _r PC _r Others 58.31%	

Fig. 10. Loss distribution of the proposed converter

The loss distribution of the proposed converter is presented in Fig. 10. It can be seen that approximately 59% of the power loss comes from conduction loss, this is because IGBTs have fixed threshold voltage (V_{CEO}) and most of the switches in the working loop are always on.

VI. ANALYSIS OF EXPERIMENTAL RESULTS

To validate the rationality of the proposed converter, a 1kW prototype is built for testing based on theoretical analysis, shown in Fig. 11. The prototype was tested in two working modes, including resistive load and resistive-inductive load. The specific experimental parameters are summarized in Table I.



Fig. 11. Photograph of experimental prototype

The Fig. 12 shows the results of the converter under operating mode 1 when the load is pure resistive load, the input voltage is 200 Vrms, and the output voltage is 150 Vrms. The partial signal waveforms of the switches are shown in Fig. 12 (a). It can be seen that only one switch operates at high

frequency. Fig. 12(b) shows the waveforms of input voltage, output voltage and output current, which are highly sinusoidal, and there is almost no phase shift between the output voltage and the output current. Fig. 12(c) shows the high-frequency voltage before filtering (U_{ab}) and the output filter inductor current (I_L). It can be seen from the partial enlarged detail that there are two change processes in the inductor current. When S₁ turns on, the inductor and load are charged from the source, and the voltage before filtering (U_{ab}) is equal to the input voltage (U_{in}). When S₁ turns off, the inductor will be freewheeling, and the voltage before filtering (U_{ab}) is approximately equal to zero. The experimental results are consistent with theoretical analysis.

6



Fig. 12. Experimental waveformss of model under resistive load $(U_{in}=200$ Vrms, $U_{out}=150$ Vrms, $R=20\Omega$)

The Fig. 13 shows the results waveforms of the converter under operating mode 1 when the load is resistive-inductive load. Fig. 13 (a) shows the switches signal waveforms. To prevent the short circuit of the bridge arm, the dead time of the 2μ s is set between S₁ and S₂. Fig. 13(b) shows the waveforms of input voltage, output voltage, and output current under resistive-inductive loads condition. It can be seen that there is an obvious phase offset between the output voltage (U_{out}) and the output current (i_0), which verifies that the converter can realize bidirectional power flow.



Fig. 13. Experimental waveformss of model under resistive-inductive load $(U_{in}=200$ Vrms, $U_{out}=150$ Vrms, $R=20\Omega$, L=25mH)

The Fig. 14 shows the results waveforms of the converter under operating mode 2 when the load is pure resistive load. Fig. 14 (a) shows the waveforms of input voltage (U_{in}), output voltage (U_{out}), output filter inductor current (I_L) and output current (i_0). As can be seen from Fig. 14(a), U_{in} and U_{out} are out of phase, that is, their phases are 180° different from each other. I_L is always higher than zero, so the converter operates in CCM. Similar to mode 1, the inductor current (I_L) also has two changes processes in mode 2. The voltage across the switches S₁ and S₂ and the input voltage are presented in Fig. 14(b). It can be seen that the peak voltage stress on switches S₁ and S₂ is equal to the input voltage.





7

Fig. 14. Experimental waveformss of mode2 under resistive load(U_{in} =200 Vrms, U_{out} =150Vrms, R=20 Ω)

The Fig. 15 shows the waveforms of the converter under operating mode 2 when the load is resistive-inductive. Fig. 15(a) shows the waveforms of input voltage (U_{in}) , output voltage (U_{out}) , and output current (i_0) . From Fig. 15(b), it appears that the peak voltage stress on capacitors C₁ and C₂ is equal to the input voltage, so there is no overvoltage problem. In addition, it can be seen that in the positive half cycle of the input voltage, the input capacitor C_{in} is connected in parallel with the clamping capacitor C_1 , and in the negative half cycle, the input capacitor C_{2} , capacitance symmetry is realized in each working process.





The Fig. 16 shows the waveform of the experimental results when the converter output voltage is changed dynamically. It can be seen that the waveforms quality of the converter is good, which verifies that the converter has good dynamic performance and the converter has the ability to output bipolar voltage.



Fig. 16. Experimental waveforms of the converter switching between mode 1 and mode 2.

FLUKER MORMA 5000 POWER ANALYZER is used to analyze the efficiency of the converter under two working modes. The efficiency curve of the converter under different input voltages is shown in Fig. 17. The input voltage ranges from 80 Vrms to 180 Vrms, and the load includes pure resistive and resistive-inductive.

Under pure resistive load, the peak efficiency of the converter is 97.47%. Under resistive-inductive load, the peak efficiency of the converter is 96.36%. Consistent with the theoretical analysis, the efficiency of the proposed converter is significantly improved due to the reduction of the number of high-frequency operating switches.



Fig. 17. Efficiency of the converter under two working modes.

VII. CONCLUSION

This paper presents a single-phase reliable bipolar direct AC-AC converter consisting of two ac chopper legs with odd symmetry and an unipolar modulation strategy. The converter has the ability to output bipolar voltage gain, which makes it very suitable for addressing the problem of power grid voltage swell and sag. The converter has high reliability due to it overcomes the commutation problems. Thanks to the unipolar modulation strategy, the number switches working at high frequency in each operating mode are optimized, so the total switching losses can be effectively reduced in theory. Moreover, capacitor symmetry can be realized in the positive and negative half cycle of the input voltage. A detailed theoretical analysis including converters working process, design consideration and power losses calculation has been given, and validated by experimental results.

REFERENCES

 F. C. L. Trindade, K. V. do Nascimento and J. C. M. Vieira, "Investigation on Voltage Sags Caused by DG Anti-Islanding Protection," *IEEE Transactions on Power Delivery*, vol. 28, no. 2, pp. 972-980, April 2013.

- [2] J. Kaniewski, P. Szczesniak, M. Jarnut and G. Benysek, "Hybrid Voltage Sag/Swell Compensators: A Review of Hybrid AC/AC Converters," *IEEE Industrial Electronics Magazine*, vol. 9, no. 4, pp. 37-48, Dec. 2015.
- [3] A. E. L. da Costa, C. B. Jacobina, N. Rocha, E. R. C. da Silva and A. V. d. M. Lacerda Filho, "A Single-Phase ac-dc-ac Unidirectional Three-Leg Converter," *IEEE Transactions on Industrial Electronics*, vol. 68, no. 5, pp. 3876-3886, May 2021.
- [4] L. Yang, H. Zhao, S. Wang and Y. Zhi, "Common-Mode EMI Noise Analysis and Reduction for AC–DC–AC Systems With Paralleled Power Modules," *IEEE Transactions on Power Electronics*, vol. 35, no. 7, pp. 6989-7000, July 2020.
- [5] T. Li, J. Chen, P. Cong, X. Dai, R. Qiu and Z. Liu, "Online Condition Monitoring of DC-Link Capacitor for AC/DC/AC PWM Converter," *IEEE Transactions on Power Electronics*, vol. 37, no. 1, pp. 865-878, Jan. 2022.
- [6] A. Ebrahimian, S. Vahid, N. Weise and A. EL-Refaie, "Two Level AC-DC-AC Converter Design with a New Approach to Implement Finite Control Set Model Predictive Control," 2021 22nd IEEE International Conference on Industrial Technology (ICIT), 2021.
- [7] Y. Sun, W. Xiong, M. Su, X. Li, H. Dan and J. Yang, "Carrier-Based Modulation Strategies for Multimodular Matrix Converters," *IEEE Transactions on Industrial Electronics*, vol. 63, no. 3, pp. 1350-1361, March 2016.
- [8] L. Qiu, L. Xu, K. Wang, Z. Zheng and Y. Li, "Research on Output Voltage Modulation of a Five-Level Matrix Converter," *IEEE Transactions on Power Electronics*, vol. 32, no. 4, pp. 2568-2583, April 2017.
- [9] M. Vijayagopal, C. Silva, L. Empringham and L. de Lillo, "Direct Predictive Current-Error Vector Control for a Direct Matrix Converter," *IEEE Transactions on Power Electronics*, vol. 34, no. 2, pp. 1925-1935, Feb. 2019.
- [10] S. Jayaprakasan, A. S and R. Ramchand, "Analysis of Current Error Space Phasor for a Space Vector Modulated Indirect Matrix Converter," *IEEE Transactions on Industrial Electronics*, vol. 69, no. 6, pp. 5680-5689, Jun. 2022.
- [11] G. H. P. Ooi, A. I. Maswood and Z. Lim, "Five-Level Multiple-Pole PWM AC-AC Converters With Reduced Components Count," *IEEE Transactions on Industrial Electronics*, vol. 62, no. 8, pp. 4739-4748, Aug. 2015.
- [12] K. Basu and N. Mohan, "A Single-Stage Power Electronic Transformer for a Three-Phase PWM AC/AC Drive With Source-Based Commutation of Leakage Energy and Common-Mode Voltage Suppression," *IEEE Transactions on Industrial Electronics*, vol. 61, no. 11, pp. 5881-5893, Nov. 2014.
- [13] A. A. Khan, H. Cha and H. Kim, "Magnetic Integration of Discrete-Coupled Inductors in Single-Phase Direct PWM AC-AC Converters," IEEE Transactions on Power Electronics, vol. 31, no. 3, pp. 2129-2138, March 2016.
- [14] S. Sharifi, M. Monfared and A. Nikbahar, "Highly Efficient Single-Phase Direct AC-to-AC Converter With Reduced Semiconductor Count," *IEEE Transactions on Industrial Electronics*, vol. 68, no. 2, pp. 1130-1138, Feb. 2021.
- [15] Jong-Hyun Kim, Byung-Duk Min, Bong-Hwan Kwon and Sang-Chul Won, "A PWM buck-boost AC chopper solving the commutation problem," *IEEE Transactions on Industrial Electronics*, vol. 45, no. 5, pp. 832-835, Oct. 1998.
- [16] J. Hoyo, J. Alcala and H. Calleja, "A high quality output AC/AC Cuk converter," 2004 IEEE 35th Annual Power Electronics Specialists Conference (IEEE Cat. No.04CH37551), 2004, pp. 2888-2893.
- [17] A. A. Khan, H. Cha and H. F. Ahmed, "An Improved Single-Phase Direct PWM Inverting Buck–Boost AC–AC Converter," *IEEE Transactions on Industrial Electronics*, vol. 63, no. 9, pp. 5384-5393, Sept. 2016.
- [18] H. Shin, H. Cha, H. Kim and D. Yoo, "Novel Single-Phase PWM AC–AC Converters Solving Commutation Problem Using Switching Cell Structure and Coupled Inductor," *IEEE Transactions on Power Electronics*, vol. 30, no. 4, pp. 2137-2147, April 2015.
- [19] A. A. Khan, H. Cha and H. Kim, "Magnetic Integration of Discrete-Coupled Inductors in Single-Phase Direct PWM AC-AC Converters," *IEEE Transactions on Power Electronics*, vol. 31, no. 3, pp. 2129-2138, March 2016.
- [20] Y. Tang, C. Zhang and S. Xie, "Z-Source AC-AC Converters Solving Commutation Problem," 2007 IEEE Power Electronics Specialists Conference, 2007, pp. 2672-2677.
- [21] L. He, S. Duan and F. Peng, "Safe-Commutation Strategy for the Novel Family of Quasi-Z-Source AC–AC Converter," *IEEE Transactions on Industrial Informatics*, vol. 9, no. 3, pp. 1538-1547, Aug. 2013.

9

- [22] H. F. Ahmed, H. Cha, A. A. Khan and H. Kim, "A Family of High-Frequency Isolated Single-Phase Z-Source AC–AC Converters With Safe-Commutation Strategy," *IEEE Transactions on Power Electronics*, vol. 31, no. 11, pp. 7522-7533, Nov. 2016.
- [23] Minh-Khai Nguyen, Young-Gook Jung and Young-Cheol Lim, "Singlephase AC/AC converter based on quasi-Z-source topology," 2009 IEEE International Symposium on Industrial Electronics, 2009, pp. 261-265.
- [24] M. Nguyen, Y. Lim and Y. Kim, "A Modified Single-Phase Quasi-Z-Source AC-AC Converter," *IEEE Transactions on Power Electronics*, vol. 27, no. 1, pp. 201-210, Jan. 2012.
- [25] L. He, J. Nai and J. Zhang, "Single-Phase Safe-Commutation Trans-Z-Source AC–AC Converter With Continuous Input Current," *IEEE Transactions on Industrial Electronics*, vol. 65, no. 6, pp. 5135-5145, June 2018.
- [26] P. Li and Y. Hu, "Unified Non-Inverting and Inverting PWM AC-AC Converter With Versatile Modes of Operation," *IEEE Transactions on Industrial Electronics*, vol. 64, no. 2, pp. 1137-1147, Feb. 2017.
- [27] H. F. Ahmed, H. Cha, A. A. Khan and H. Kim, "A Novel Buck–Boost AC– AC Converter With Both Inverting and Noninverting Operations and Without Commutation Problem," *IEEE Transactions on Power Electronics*, vol. 31, no. 6, pp. 4241-4251, June 2016.
- [28] H. F. Ahmed, M. S. El Moursi, H. Cha, K. Al Hosani and B. Zahawi, "A Reliable Single-Phase Bipolar Buck–Boost Direct PWM AC–AC Converter With Continuous Input/Output Currents," *IEEE Transactions* on Industrial Electronics, vol. 67, no. 12, pp. 10253-10265, Dec. 2020.
- [29] C. Liu et al., "Novel Bipolar-Type Direct AC-AC Converter Topology Based on Non-Differential AC Choppers," *IEEE Transactions on Power Electronics*, vol. 34, no. 10, pp. 9585-9599, Oct. 2019.
- [30] Y. Wang et al., "An Improved Bipolar-Type AC–AC Converter Topology Based on Nondifferential Dual-Buck PWM AC Choppers," *IEEE Transactions on Power Electronics*, vol. 36, no. 4, pp. 4052-4065, April 2021.