An AC Z-Source Converter Based on Gamma Structure With Safe-Commutation Strategy

Mohamad Reza Banaei, Rana Alizadeh, Nazanin Jahanyari, and Ebrahim Seifi Najmi

***Abstract*—This paper proposes a novel single-phase ac–ac Z-source converter based on the Γ-structure. The proposed converter uses a coupled transformer to produce the required gain. The obtained gain can be tuned by varying the turn ratio γ_Γ of the transformer within the narrow range of $1 < \gamma_\Gamma < 2$. Also, the gain factor increases with decrease of the transformer turns ratio, γ_Γ. Small transformer turn ratio results in more practical topology in some voltage gains. It makes the transformer design easier and reduces construction cost. Also, in the proposed topology, the input and output voltages share the same ground. In order to avoid voltage and current spikes on the switches, a commutation strategy is proposed and safe commutation is achieved without using a snubber circuit. The operating principles and steady-state analyses are presented to verify the operational concept of the proposed topology. Also, experimental results are given to verify the validation of the proposed converter.**

***Index Terms*— Quasi-Z-source, trans-Z-source, Z-source, Γ-Z-source.**

I. Introduction

MODERN power electronic applications usually require boosted voltage at their output side. Indirect ac–ac converters with dc link [1], matrix converters [2], and direct PWM ac–ac converters [2]–[5] are the most utilized ac power conversion systems. The disadvantage of back-to-back converters is that the diode-rectifier link causes power-line pollution. Matrix converters have the disadvantage of lower voltage ratio, which is 0.866. They also own complex structure and commutation strategy. However, they can provide output voltage with variable frequency, but for applications where only voltage regulation is needed, direct PWM ac–ac converters have the merits such as single-stage conversion, simpler topology, easier control, higher efficiency, smaller size, lower cost, and lower current harmonics in line.

Traditional single-phase Z-source ac–ac converters (ZSACs) have some advantages such as providing a larger range of output voltage in buck–boost mode, with reversing or maintaining phase angle [6], [7]. However, the main drawback of conventional voltage-fed ZSAC [6] is that the ground in input and output voltages is not shared. Hence, the feature that the output voltage reverses or maintains its phase angle with the input voltage is not supported well. Direct PWM ac–ac converters can be derived from dc–dc topologies by substituting all of the unidirectional switches with bidirectional devices [8]. The authors in [9] and [10] use direct PWM ac–ac converters to overcome voltage sags and swells as static VAR compensators in power systems. The performance of ac–ac converters can be improved significantly using safe-commutation strategy in switching with PWM control as presented in [11]. A class of single phase PWM ac–ac power converters with simple topologies have been presented in [12]–[18]. These include buck, boost, buck–boost, and Cuk converters. Z-source converters applied to dc–ac inverters and ac–ac converters have recently been proposed in [6] and [17]–[19]. The research on the dc–ac inverters consists of modeling, control [20], [21], and PWM strategy [24]. Trans Z-source inverters have recently been proposed to improve the voltage gain using coupled inductors [18], [19]. Quasi-Z-source converters have been presented in [17] and [22]. New Z-source inverters with ability of operation in buck and boost modes are possible by using an impedance network between the power source and converter circuit [23]. Various ZSACs based on Z-source dc–ac are presented considering single-phase [25], [18] and three-phase topologies [26].

In this paper, a single-phase ac–ac Γ-Z-source converter is proposed, which has buck–boost capability and can maintain or reverse the output phase angle. Also, the input and output have common ground. Decrease in transformer turn ratio increases the voltage gain, which results in more practical topology in some voltage gains. It makes the transformer design easier and reduces the construction cost. A new safe-commutation strategy is provided for the proposed converter, which eliminates voltage and current spikes on the switches without using a snubber circuit. The operating principle, designing analyses, and experimental results are also presented to verify the validity of the proposed structure.

In conventional single-phase ZSAC [6], a small snubber circuit is needed for each switch in order to limit the voltage overshoots and provide commutation paths during the dead times. This causes inefficiency and unreliability. In order to share the ground of input and output sides in the conventional converter, another topology is proposed in [7]. The boost factor in both converters is expressed as follows:

$$B = \frac{1-D}{1-2D}. \tag{1}$$

Fig. 1 shows the proposed single-phase Γ-Z-source ac–ac converter. This topology uses an *LC* filter at the output, one capacitor, and a transformer with turn ratio of $\gamma_\Gamma = \frac{w_1}{w_2}$. The

Manuscript received June 17, 2014; revised August 18, 2014, October 28, 2014, and January 3, 2015; accepted March 11, 2015. Date of publication April 2, 2015; date of current version September 29, 2015. Recommended for publication by Associate Editor R. Burgos.

The authors are with the Electrical Engineering Department, Azarbaijan Shahid Madani University, Tabriz, Iran (e-mail: m.banaei@azaruniv.edu; ranaalizadeh@azaruniv.edu; n.jahanyari@azaruniv.edu; e.seifi@azaruniv.edu).

Color versions of one or more of the figures in this paper are available online at http://ieeexplore.ieee.org.

Digital Object Identifier 10.1109/TPEL.2015.2415735

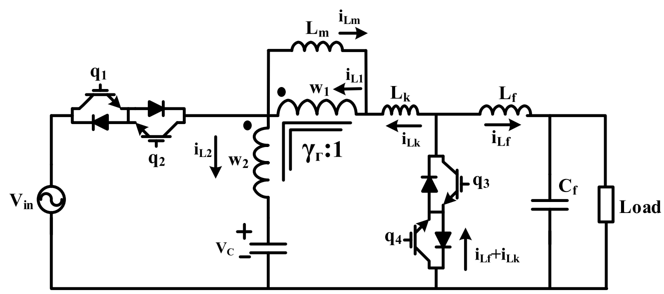

Fig. 1. Proposed Γ-Z-source ac–ac converter.

boost coefficient of the Γ-Z-source converter is expressed as

$$B = \frac{1-D}{1-D\left(1+\frac{1}{\gamma_\Gamma - 1}\right)}. \tag{2}$$

The boundary of $1 < \gamma_\Gamma < 2$ can obtain greater boost factor (B) in comparison with conventional ZSAC. Also, considering the boost factor (B) more than zero, new limits for shoot through duty cycle, D, will be $0 < D < \frac{\gamma_\Gamma - 1}{\gamma_\Gamma}$. It is obvious from (2) that smaller turns ratio results in higher boost factor.

II. Study of the Proposed Γ-Z-Source Converter

A. Circuit Analysis

The following assumptions are assumed for the circuit analysis of the proposed single-phase Γ-Z-source —ac–ac converter: 1) all capacitors and switches are ideal and lossless; and 2) the converter is operating in the continuous conduction mode.

The proposed converter contains two states in each switching period: 1) non-shoot-through and 2) shoot-through. The time interval of $(1-D)T_s$ is taken for the non-shoot-through state, while DT_s is the time interval of shoot-through mode.

The coupled inductor can be modeled as an ideal transformer that has magnetizing inductor (L_m) and leakage inductor (L_k). The turns ratio (γ_Γ) and coupling coefficients (k) of this transformer are defined as

$$\gamma_\Gamma = \frac{v_{w_1}}{v_{w_2}} \tag{3}$$

$$k = \frac{L_m}{L_m + L_k}. \tag{4}$$

Using these definitions, the equations for non-shoot-through mode are as follows:

$$\begin{pmatrix} L_m & 0 & 0 & 0 \\ 0 & L_f & 0 & 0 \\ 0 & 0 & C & 0 \\ 0 & 0 & 0 & C_f \end{pmatrix} \frac{d}{dt} \begin{pmatrix} i_{L_m}(t) \\ i_{L_f}(t) \\ v_C(t) \\ v_O(t) \end{pmatrix} =$$

$$\begin{pmatrix} 0 & 0 & -\gamma_\Gamma & 0 \\ 0 & 0 & \frac{\gamma_\Gamma}{k} & -1 \\ \gamma_\Gamma & -\gamma_\Gamma & 0 & 0 \\ 0 & 1 & 0 & \frac{-1}{R} \end{pmatrix} \begin{pmatrix} i_{L_m}(t) \\ i_{L_f}(t) \\ v_C(t) \\ v_O(t) \end{pmatrix} +$$

$$\begin{pmatrix} \gamma_\Gamma v_i(t) \\ \left(1-\frac{\gamma_\Gamma}{k}\right) v_i(t) \\ 0 \\ 0 \end{pmatrix}. \tag{5}$$

The shoot-through analyses are calculated as follows:

$$\begin{pmatrix} L_m & 0 & 0 & 0 \\ 0 & L_f & 0 & 0 \\ 0 & 0 & C & 0 \\ 0 & 0 & 0 & C_f \end{pmatrix} \frac{d}{dt} \begin{pmatrix} i_{L_m}(t) \\ i_{L_f}(t) \\ v_C(t) \\ v_O(t) \end{pmatrix} =$$

$$\begin{pmatrix} 0 & 0 & \frac{k\gamma_\Gamma}{\gamma_\Gamma - k} & 0 \\ 0 & 0 & 0 & -1 \\ \frac{\gamma_\Gamma}{1-\gamma_\Gamma} & 0 & 0 & 0 \\ 0 & 1 & 0 & \frac{-1}{R} \end{pmatrix} \begin{pmatrix} i_{L_m}(t) \\ i_{L_f}(t) \\ v_C(t) \\ v_O(t) \end{pmatrix}. \tag{6}$$

In the steady state, the averaging equation is deduced as

$$\begin{pmatrix} L_m & 0 & 0 & 0 \\ 0 & L_f & 0 & 0 \\ 0 & 0 & C & 0 \\ 0 & 0 & 0 & C_f \end{pmatrix} \frac{d}{dt} \begin{pmatrix} i_{L_m}(t) \\ i_{L_f}(t) \\ v_C(t) \\ v_O(t) \end{pmatrix} = \begin{pmatrix} 0 \\ 0 \\ 0 \\ 0 \end{pmatrix}. \tag{7}$$

Hence

$$\begin{cases} v_C(t) = \dfrac{(1-D)v_i(t)}{1-D\left(1+\dfrac{k}{\gamma_\Gamma - k}\right)} \\ v_O(t) = \dfrac{(1-D)v_i(t)}{1-D\left(1+\dfrac{k}{\gamma_\Gamma - k}\right)} \\ i_{L_m}(t) = \dfrac{-(1-D)i_{L_f}(t)}{1-D\left(1+\dfrac{1}{\gamma_\Gamma - 1}\right)} \\ i_{L_f}(t) = \dfrac{v_O(t)}{R}. \end{cases} \tag{8}$$

Hence, the voltage gain of the proposed converter is deduced as

$$B = \frac{(1-D)}{1-D\left(1+\frac{k}{\gamma_\Gamma - k}\right)}. \tag{9}$$

Considering $k = 1$, the voltage gain can be obtained as

$$B = \frac{(1-D)}{1-D\left(1+\frac{1}{\gamma_\Gamma - 1}\right)}. \tag{10}$$

It can be deduced from (9) and (10) that the boost factor will increase with reducing the transformer turn ratio. As mentioned above, this advantage results in more practical topology in some required voltage gains. It also reduces transformer construction and wounding cost and simplifies the transformer design. Hence, a practical ac–ac converter would be achieved.

B. Commutation Study

As it is shown in Fig. 1, two switches with body diodes are connected in common emitter back-to-back manner to

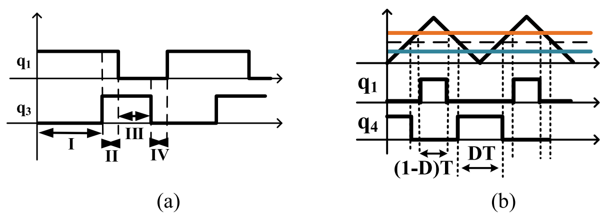

Fig. 2. Modulation of switches (a) considering practical dead-time for switches and (b) considering the limitation for buck out-of-phase mode ($v_i > 0$).

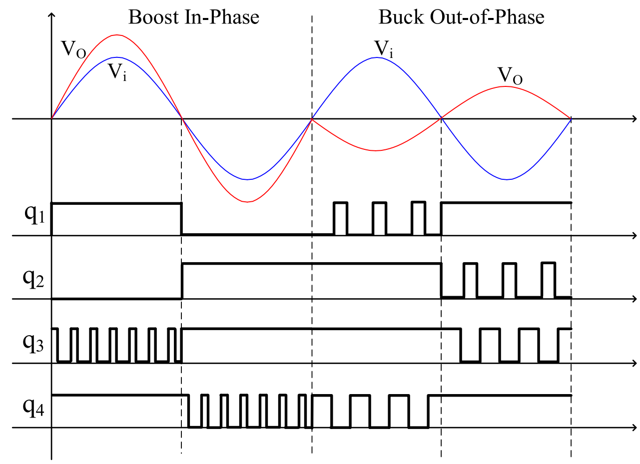

Fig. 3. Proposed safe commutation strategy for boost in-phase and buck out-of-phase modes.

implement as a bidirectional switch. In order to prevent the damages on implemented switches, a suitable safe commutation strategy is proposed to solve the problem of current and voltage spikes. It is worth mentioning that current spikes are generated because of sudden changes in capacitor voltages. Also, voltage spikes are due to the sudden changes in current flowing through inductors [7], [28].

If $v_i > 0$, q_1 and q_3 should be modulated in complement as shown in Fig. 2(a). Also, during the time interval in which $v_i < 0$, q_2 and q_4 should be modulated in complement. Considering the fact that the switches are not ideal and have dead time in both turning ON and OFF modes, four modes will occur during switching period as shown in Fig. 2(a):

For $v_i > 0$, at the time interval I, q_1 is turned ON and q_3 is turned OFF. During this mode, the capacitor C is charged.

At time interval II, q_3 is turned ON, while q_1 has not still been OFF due to nonideal characteristics of switches. Hence, two voltage loops will be formed: 1) the first loop contains the input voltage, v_i, the secondary side of coupled transformer and capacitor, and 2) the second loop consists of the coupled transformer and capacitor. Therefore, a sudden change in capacitor voltage occurs, which results in current spikes in the bidirectional switches.

At time interval III, q_1 is turned OFF and q_3 is turned ON. In this mode, the capacitor C is discharged.

At time interval IV, q_3 is turned OFF before q_1 is turned ON due to nonideal characteristics of the switches. In this time interval, the series connection of output filter inductor and primary side of the coupled transformer results in a sudden change in inductor current. Therefore, a voltage spike on bidirectional switches will occur.

In order to solve the aforementioned problems, safe commutation strategy should be utilized. A modulation limitation should be considered in order to prevent current spikes in q_1 and q_2. Four switches q_1, q_2, q_3, and q_4 should never be ON at the same time. For this reason, according to the dead time characteristics of utilized switches, the pulsewidth of DT_s and $(1-D)T_s$ is reduced as shown in Fig. 2(b).

In boost in-phase operation mode, if $v_i > 0$ and $I_{\rm in} < 0$, q_2 can be modulated considering the limitation that the four switches should not be turned ON simultaneously. Therefore, q_2 can provide the input current path in this situation. On the contrary, if $v_i < 0$ and $I_{\rm in} > 0$, q_1 can be modulated considering the aforementioned limitation. However, if this condition occurs in a negligible time, it is preferable to turn q_2 and q_1 fully OFF when $v_i > 0$ and $v_i < 0$, respectively. This strategy is shown in Fig. 3. This is because, every turn-on and turn-off operation of the switches include losses. Especially, because q_1 and q_2 are at input side, have high stress, and are at hard-switching situation, the low frequency modulation of these switches can reduce the losses and increase the switch lifetime. Also, in this modulation, the aforementioned limitation is not needed to be considered, because one of the switches is OFF at each time period.

In this commutation strategy, two switches are fully turned ON considering the input voltage direction. q_1 and q_4 are fully turned ON and q_3 is modulated in shoot-through duty cycle, DT_s. If $i_{Lf} + i_{Lk} > 0$, q_4 and the body diode of q_3 are the current path; otherwise, the current will pass through q_1 and the body diode of q_2. When $-v_{w2} + v_C$ becomes larger than input voltage source (i.e., $-v_{w2} + v_C > v_{\rm in}$), the body diode of q_2 is turned off and the capacitor C is discharged. The proposed safe commutation strategy is depicted in Fig. 3.

For buck out of phase mode, if $v_i > 0$, q_1 and q_4, otherwise q_2 and q_3, are modulated. For $v_i > 0$, q_1 and q_4 are modulated in this structure, while q_2 and q_3 are fully on. If $i_{Lf} + i_{Lk} > 0$, then q_2 and the body diode of q_1 will provide the current path. But if $i_{Lf} + i_{Lk} < 0$, the current will pass through q_3 and the body diode of q_4. For $v_i < 0$, the same method should be used as shown in Fig. 3.

The safe commutation strategy for boost out-of-phase mode is on contrary with Fig. 3, where q_1 and q_4 are fully ON when the input voltage is negative. Also, q_2 and q_3 are fully ON when the input voltage is positive.

C. Parameter Design

1) Γ-Z-Source AC–AC Converter Network Design: The magnetizing inductor parameter of the proposed Γ-Z-source ac–ac converter is selected according to the current ripple in state II. During this state, the voltage of magnetizing inductor, L_m, equals $\frac{k\gamma_\Gamma}{\gamma_\Gamma - k} v_C$. The worst case for the parameter design is when $k = 1$. Hence, the parameter of L_m could be calculated as

$$L_m = \frac{\gamma_\Gamma}{\gamma_\Gamma - 1} \frac{DT.\,|v_C|}{\Delta i_{L_m}}. \tag{11}$$

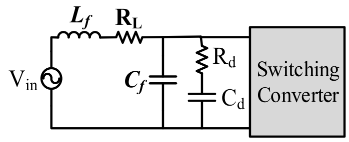

Fig. 4. Input *LC* filter.

Considering the current ripple to be $\Delta i_m \leq x\% \cdot I_{Lm}$, the following equation could be deduced:

$$L_m \geq \frac{\gamma_\Gamma}{\gamma_\Gamma - 1}\frac{DT.\,|v_C|}{x\% \cdot I_{L_m}} = \sqrt{2}\frac{\gamma_\Gamma}{\gamma_\Gamma - 1}\frac{DT}{x\% \cdot I_{L_m}}\frac{V_i V_o}{P_o}. \quad (12)$$

In order to select the size of capacitor C, the voltage ripple of capacitor in state II is used:

$$C = \frac{\gamma_\Gamma}{\gamma_\Gamma - 1}\frac{DT \cdot |i_{L_m}|}{\Delta v_C}. \quad (13)$$

Assuming the voltage ripple of capacitor to be $\Delta v_C \leq m\% \cdot V_C$, the equation of capacitor C can be written as

$$C \geq \frac{\gamma_\Gamma}{\gamma_\Gamma - 1}\frac{DT \cdot |i_{L_m}|}{m\% \cdot V_C} = \sqrt{2}\frac{\gamma_\Gamma}{\gamma_\Gamma - 1}\frac{DT \cdot P_o}{m\% \cdot V_i \cdot V_o}. \quad (14)$$

2) Parameter Design of Output Filter: The output filter inductance is calculated as follows:

$$L_f = \frac{\varepsilon R_L}{2\pi f_C}. \quad (15)$$

Moreover, the output filter capacitor is deduced as

$$C_f = \frac{1}{2\pi f_C \varepsilon R_L} \quad (16)$$

where f_c and R_L are cutoff frequency and load resistance, respectively. ε is a constant between 0.5 and 0.8 [7].

3) Input Filter Design: In some applications such as voltage sag/swell compensators, the switches q_1 and q_2 at input side can be used for bypass mode of compensator [29]. However, in order to attenuate the switching harmonics in the converter input current, an input filter is of great importance [30]. The input *LC* filter has become very critical in its design and must be designed not only for conducted electromagnetic interface but also for system stability. The equivalent circuit of the input filter is also shown in Fig. 4. The voltage Δv_C is the sum of the voltage across the equivalent series resistance (ESR) and the reactance of the capacitor.

The voltage developed across the capacitance is

$$\Delta v_{CC} = I_C \frac{D \cdot (1 - D)}{C_1 \cdot T}. \quad (17)$$

Also, the voltage across the ESR is developed as

$$\Delta v_{CR} = I_C \cdot (\mathrm{ESR}). \quad (18)$$

Hence

$$\Delta v_C = \Delta v_{CC} + \Delta v_{CR}. \quad (19)$$

The input filter can affect the stability of the switching converter. In order to avoid this to happen, the output impedance of the input filter should be lower than the open-loop input impedance of the converter, i.e.,

$$Z_{\mathrm{in}} \gg Z_{\mathrm{of}} \quad (20)$$

$$\frac{\eta \cdot (V_{\mathrm{in}})^2}{P_O} > \frac{L}{\dfrac{C + (R_L + R_S) \cdot (\mathrm{ESR})}{(R_L + R_S) + (\mathrm{ESR})}} \quad (21)$$

where η is the converter efficiency, V_{in} is the maximum input voltage, P_O is the output power in watts, L is the input inductor in henries, C is the filter capacitor in farads, R_L is the inductor series resistance in ohms, R_S is the source resistance in ohms, and R_d is ESR in ohms.

If additional damping is required, it can be done by increasing R_d or R_L [see Fig. 4].

If parallel damping was necessarily required, R_d and C_d can be calculated as follows:

$$R_d = \sqrt{\frac{L}{C}}, \quad 4C \leq C_d. \quad (22)$$

The input filter L is designed as follows. Considering the inductor current ripple developed by Δv_{CR} to be ΔI_{LR}

$$\Delta I_{LR} = \left(\frac{\Delta v_{CR}}{L}\right) D.(1 - D)T. \quad (23)$$

Also, the inductor ripple current developed by ΔI_{LC} is as follows:

$$\Delta I_{LC} = \left(\frac{\Delta v_{CC}}{2L}\right)\left(\frac{T}{4}\right). \quad (24)$$

Due to the capacitor, ESR, ΔI_{LR} dominates ΔI_{LC}. Hence

$$L = \frac{\Delta v_{CR}}{\Delta I_{LR}}(D) \cdot (1 - D)T. \quad (25)$$

4) Voltage and Current Stress of Switches: The maximum voltage of bidirectional switches is as follows:

$$V_{s_1-\max} = V_{s_2-\max} = \sqrt{2}(V_i - V_{L_m}) = \frac{\sqrt{2}V_i}{\gamma_\Gamma(D - 1) + 1}. \quad (26)$$

The RMS current values of two switches are calculated as follows:

$$I_{s_1-\mathrm{rms}} = \sqrt{1 - D}I_i = \frac{1}{\sqrt{1 - D}}\frac{P_O}{V_i} \quad (27)$$

$$I_{s_2-\mathrm{rms}} = \sqrt{D}(i_2 + i_{L_f}) = \sqrt{D}\left(\frac{2\gamma_\Gamma(1 - D) - 1}{(1 - D)(\gamma_\Gamma - 1)}\right)\frac{P_O}{V_i}. \quad (28)$$

D. Proposed Γ-Z-Source Converter Versus Other Z-Source Topologies

1) Γ-Z-Source Converter in Comparison With Previous Topologies: Table I indicates the boost factor, currents, and voltages of utilized components in the proposed topology, conventional converter [6], and the quasi-ZSAC presented in [17]. I_i, I_{L1}, I_{L2}, I_{Lm}, and V_i are the RMS values of input, inductor,

TABLE I
Γ-Z-SOURCE CONVERTER VERSUS OTHER Z-SOURCE TOPOLOGIES [6], [17]

	Conventional converter in [6]	Quasi-Z-source converter in [17]	Proposed Γ-Z-source converter
$\frac{V_o}{V_i}$	$\frac{1-D}{1-2D}$	$\frac{1-D}{1-2D}$	$B = \frac{1-D}{1-D\left(1+\frac{1}{\gamma_\Gamma - 1}\right)}$
$\frac{V_{C1}}{V_i}$	$\frac{1-D}{1-2D}$	$\frac{D}{1-2D}$	
$\frac{V_{C2}}{V_i}$	$\frac{1-D}{1-2D}$	$\frac{1-D}{1-2D}$	$B = \frac{1-D}{1-D\left(1+\frac{1}{\gamma_\Gamma - 1}\right)}$
$I_{L1} \& I_{L2}$	$\frac{P_o}{V_i}$	$\frac{P_o}{V_i}$	-
I_{Lm}	-	-	$\frac{P_o}{V_i}$
I_i	$\frac{P_o}{\sqrt{1-D}V_i}$	$\frac{P_o}{V_i}$	$\frac{P_o}{\sqrt{1-D}V_i}$
$I_{s-\max}$	$\frac{\sqrt{2}P_o}{(1-D)V_i}$	$\frac{\sqrt{2}P_o}{(1-D)V_i}$	$I_{s1-\max} = \sqrt{2}\frac{P_0}{V_i(1-D)}$ $I_{s2-\max} = \sqrt{2}\frac{2\gamma_\Gamma(1-D)-1}{(1-D)(\gamma_\Gamma-1)} \cdot \frac{P_O}{V_i}$
$V_{s-\max}$	$\frac{\sqrt{2}V_i}{1-2D}$	$\frac{\sqrt{2}V_i}{1-2D}$	$\frac{\sqrt{2}V_i}{\gamma_\Gamma(D-1)+1}$

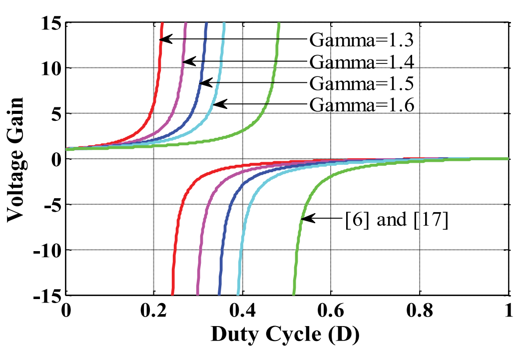

Fig. 5. Output voltage gain versus duty cycle with variable ratio of γ_Γ.

L_m currents, and input voltage, respectively. P_o is the output power. $I_{s,\max}$ and $V_{s,\max}$ are RMS values of current and voltage stresses of switches, respectively.

It is obvious from this table that the proposed topology uses less components in comparison with the conventional ac–ac Z-source converter [6] and the same number of elements in comparison with the topology presented in [15], considering the input LC filter for the proposed Γ-Z-source converter. Also, the boost factor of the proposed converter is higher and has wider range of control due to the use of coupled transformer. Voltage and current stress in all three topologies are nearly the same considering both buck and boost modes of operation.

Fig. 5 shows the output voltage gains versus the duty cycle D with variable turn ratios of γ_Γ. This figure depicts that there are three operation regions: 1) when duty cycle is smaller than $\frac{\gamma_\Gamma - 1}{\gamma_\Gamma}$, the output voltage is boosted and in phase with the input voltage; 2) when duty cycle is $\frac{\gamma_r - 1}{\gamma_r} < D < \frac{2(\gamma_r - 1)}{2\gamma_r - 1}$, the output voltage is boosted out of phase with the input voltage; and 3) when duty cycle is $\frac{2(\gamma_\Gamma - 1)}{2\gamma_\Gamma - 1} < D < 1$, the output voltage is bucked and out of phase with the input voltage. Also, the efficiency

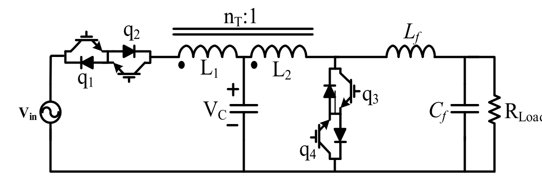

Fig. 6. Proposed single-phase ac–ac Z-source converter with the T-structure.

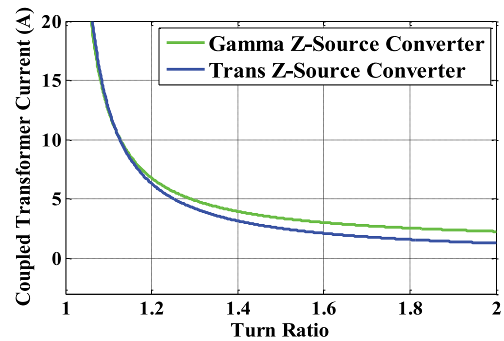

Fig. 7. Current passing through coupled transformer in both T and Γ structures.

comparison in boost in-phase operation mode of the proposed converter with previous topologies [6], [17] , [28] is obtained within experimental results in Section III. As it is shown in this section, the efficiency of the proposed converter is higher than previous topologies.

2) Γ-Structure of Coupled Transformer in Comparison With the T-Structure: Considering the quasi-ac–ac Z-source converter [17] and replacing the utilized inductors with coupled transformer, the voltage across L_1 is reflected to the inductor L_2 through magnetic coupling. Then, one of the two capacitors can be removed from the circuit. Fig. 6 shows the T-structures of coupled transformer in an ac–ac Z-source converter.

The boost factor for trans-source converter is obtained as

$$B = \frac{1-D}{1-D(1+n_T)}. \tag{29}$$

Also, the boost factor for Γ-source converter is achieved in (10). For better comparison of the two proposed topologies, the boost factor and shoot-through time intervals are considered equal. Hence, we have

$$\gamma_\Gamma = 1 + \frac{1}{n_T}. \tag{30}$$

From equalization of (30), for $1 \leq n_T < \infty$, the turn ratio of the Γ-structure will be $1 < \gamma_\Gamma < 2$. It is obvious from (29) that increasing n_T results in higher voltage gain. Although this feature is natural for most of the transformer-based power electronic converters, but it is a drawback for some required voltage gains, making the converter impracticable. Equation (10) shows that the voltage gain of Γ-Z-source converters increases while decreasing the transformer turn ratio. γ_Γ is within the narrow range of (1,2]. Fig. 7 shows that the current passing through both transformer structures are nearly the same. Although smaller transformer turns results in higher nH/t2 of the transformer core, this feature is more practical in larger voltage gains [27].

TABLE II
LIST OF EXPERIMENTAL PARAMETERS

$D_{\text{Shootthrough}}$ (boost and buck)	0.3 and 0.7
Switching frequency (f_s)	30 KHz
C	25 μF
γ_r	94/60
L_m	0.6 mH
Coupling coefficient factor (k)	0.999
Output LC filter (L_f and C_f)	50 mH and 40 μF
$R_{\text{Load}} - L_{\text{Load}}$	60 Ω–100 mH

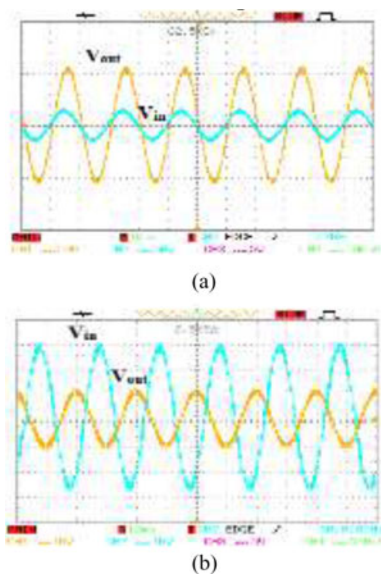

Fig. 8. Experimental results. (a) Input and output voltages, $(1v/\text{div})$, boost in-phase mode $D = 0.3$. (b) Input and output voltages, $(1 \times 1v/\text{div})$, buck out-of-phase mode $D = 0.7$.

This is because, for tight coupling of the transformer, especially at low voltage side, the number of turns should be large enough. This results in even larger turns in the transformer high voltage side. For example, instead of $n_T = 3$ in the T-structure, a turn ratio of $n_{\text{gamma}} = 1.33$ can be used. Hence, less turn ratio in gamma results in more practical topology. It makes the transformer design easier and reduces construction cost remarkably [31], [32].

III. Experimental Results

Experimental results are presented to verify the validity of the proposed ac–ac Γ-Z-source converter. The list of experimental parameters is given in Table II. As calculated in previous section, for the mentioned parameters, an input LC filter of 0.625 mH and 19 μF can be used. The input peak voltage of 25 V with transformer turn ratio of 94/60 is used for both buck and boost modes in experiment. Fig. 8(a) and (b) shows the input and output voltages of the proposed converter for boost in-phase and buck out-of-phase modes of operation. It is clear from this figure that the modulation change with zero crossing does not affect the output voltage.

Figs. 9(a), (b) and 10(a), (b) show the voltage across the switches q_1 and q_2, q_3 and q_4 in boost mode, respectively. The

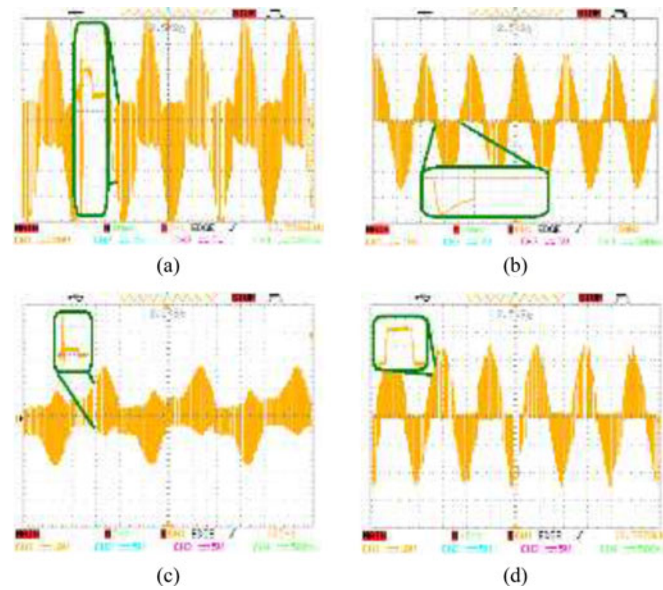

Fig. 9. Voltage and current of q_1 and q_2 before and after safe commutation strategy for boost mode. (a) v_{q1-q2} before safe commutation strategy $(1v/\text{div})$. (b) v_{q1-q2} after safe commutation strategy $(10 \times 1v/\text{div})$. (c) i_{q1-q2} before safe commutation strategy $(10 \times 1A/\text{div})$. (d) i_{q1-q2} after safe commutation strategy $(1A/\text{div})$.

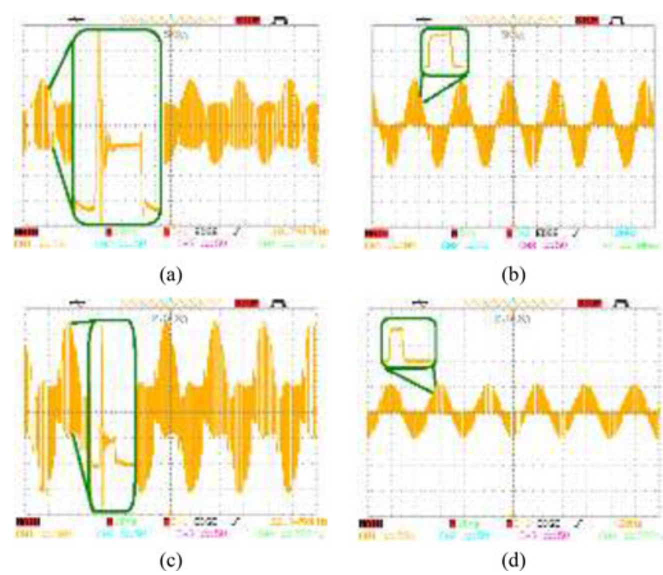

Fig. 10. Voltage and current of q_3 and q_4 before and after safe commutation strategy for boost mode. (a) v_{q3-q4} before safe commutation strategy, $(1v/\text{div})$. (b) v_{q3-q4} after safe commutation strategy $(10 \times 1\,v/\text{div})$. (c) i_{q3-q4} before safe commutation strategy $(10 \times 1\,A/\text{div})$. (d) i_{q3-q4} after safe commutation strategy $(1\,A/\text{div})$.

problem of voltage spikes has been solved properly after using safe commutation strategy. Figs. 9(c), (d) and 10(c), (d) show the input current flowing through q_1, q_2 and q_3, q_4 switches in boost mode, before and after using safe commutation strategy, respectively. It is clear from the figures that safe commutation strategy has prevented the current spikes thoroughly.

The efficiency of the proposed converter is measured and compared with conventional Z-source [6], quasi-Z-source [17],

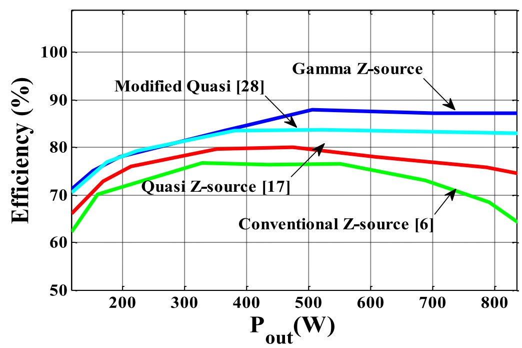

Fig. 11. Experimental results of the efficiency of the proposed converter versus output power, in comparison with conventional Z-source [6], Quasi Z-source [17], and modified quasi Z-source converters [28] in boost in-phase mode.

and modified quasi-ZSACs [28] in boost in phase mode, as shown in Fig. 11.

IV. Conclusion

A novel single-phase ac–ac Z-source converter based on the Γ-structure of coupled transformer is proposed in this paper. A novel safe commutation strategy is also employed. The safe commutation strategy is a significant improvement as it makes it possible to avoid voltage and current spikes on the switches. The operation principal, steady-state analyses, and parametric design are presented in detail. The output and input voltages share the same ground in the proposed converter. Transformer ratio is controlled in a small range of $1 < \gamma_\Gamma < 2$. The gain factor increases while reducing the transformer ratio. This exclusive feature results in a practical converter for some voltage gains. The input *LC* filter design is also provided to attenuate the switching harmonics that are presented in the converter input current. Experimental results are also provided to verify the validity of the proposed structure. The boosted in-phase and bucked out-of-phase output voltages, the current passing through the switches, and the voltage across them are shown before and after using safe commutation strategy within experimental results

References

[1] T. Friedli, J. W. Kolar, J. Rodriguez, and P. W. Wheeler, "Comparative evaluation of three-phase AC–AC matrix converter and voltage DC-link back-to-back converter systems," *IEEE Trans. Ind. Electron.*, vol. 59, no. 12, pp. 4487–4510, Dec. 2012.

[2] L. Empringham, J. W. Kolar, J. Rodriguez, and P. W. Wheeler, "Technological issues and industrial application of matrix converters: A review," *IEEE Trans. Ind. Electron.*, vol. 60, no. 10, pp. 4260–4271, May 2013.

[3] O. Ellabban, H. Abu-Rub, and Ge Baoming, "Field oriented control of an induction motor fed by a quasi-Z-source direct matrix converter," in *Proc. IEEE 39th Ann. Conf. Ind. Electron. Soc.*, Vienna, Austria, 2013, pp. 4850–4855.

[4] O. Abdel-Rahim, H. Abu-Rub, A. Iqbal, and A. Kouzou, "Five-to-three phase direct matrix converter with model predictive control," in *Proc. IEEE Conf. Power Eng.*, 2013, pp. 204–208.

[5] B. H. Kwon, G. Y. Jeong, S. H. Han, and D. H. Lee, "Novel line conditioner with voltage up/down capability," *IEEE Trans. Ind. Appl.*, vol. 49, no. 5, pp. 1110–1119, Oct. 2002.

[6] X. P. Fang, Z. M. Qian, and F. Z. Peng, "Single-phase Z-source PWM AC-AC converters," *IEEE Power Electron. Letter*, vol. 3, no. 4, pp. 121–124, Dec. 2005.

[7] Y. Tang, S. Xie, and C. Zhang, "Z-source AC–AC converters solving commutation problem," *IEEE Trans. Power Electron.*, vol. 22, no. 6, pp. 2146–2154, Nov. 2007.

[8] F. L. Luo and H. Ye, "Research on dc-modulated power factor correction AC/AC converters," in *Proc. IEEE 33rd Ann. Conf. Ind. Electron.*, Taipei, Taiwan, 2007, pp. 1478–1483.

[9] S. Subramanian and M. K. Mishra, "Interphase AC–AC topology for voltage sag supporter," *IEEE Trans. Power Electron.*, vol. 25, no. 2, pp. 514–518, Feb. 2009.

[10] D. M. Divan and J. Sastry, "Voltage synthesis using dual virtual quadrature sources—A new concept in ac power conversion," *IEEE Trans. Power Electron.*, vol. 23, no. 6, pp. 3004–3013, Nov. 2008.

[11] M. K. Nguyen, Y. G. Jung, Y. C. Lim, and Y. M. Kim, "A single-phase Z-source buck-boost matrix converter," *IEEE Trans. Power Electron.*, vol. 25, no. 2, pp. 453–462, Feb. 2010.

[12] F. Z. Peng, L. Chen, and F. Zhang, "Simple topologies of PWM AC–AC converters," *IEEE Power Electron. Lett.*, vol. 1, no. 1, pp. 10–13, Mar. 2003.

[13] R. Stala, S. Pirog, A. Mondzik, M. Baszynski, A. Penczek, J. Czekonski, and S. Gasiorek, "Results of investigation of multicell converters with balancing circuit. Part II," *IEEE Trans. Ind. Electron.*, vol. 56, no. 7, pp. 2620–2628, Jul. 2009.

[14] R. H. Wilkinson, T. A. Meynard, and H. T. Mouton, "Natural balance of multicell converters: The general case," *IEEE Trans. Power Electron.*, vol. 21, no. 6, pp. 1658–1666, Nov. 2006.

[15] L. Li, J. Yang, and Q. Zhong, "Novel family of single-stage three-level ac choppers," *IEEE Trans. Power Electron.*, vol. 26, no. 2, pp. 504–511, Feb. 2011.

[16] D. Chen and J. Liu, "The uni-polarity phase-shifted controlled voltage mode AC–AC converters with high frequency AC link," *IEEE Trans. Power Electron.*, vol. 21, no. 4, pp. 899–905, Jul. 2006.

[17] M. K. Nguyen, Y. G. Jung, and Y. C. Lim, "Single-phase AC–AC converter based on quasi-Z-source topology," *IEEE Trans. Power Electron.*, vol. 25, no. 8, pp. 2200–2210, Aug. 2010.

[18] R. Strzelecki, M. Adamowicz, N. Strzelecka, and W. Bury, "New type T-source inverter," in *Proc. Compat. Power Electron.*, Badajoz, 2009, pp. 191–195.

[19] W. Qian, F. Z. Peng, and H. Cha, "Trans-Z-source inverters," *IEEE Trans. Power Electron.*, vol. 26, no. 12, pp. 3453–3463, Dec. 2011.

[20] P. C. Loh, F. Blaabjerg, and C. P. Wong, "Comparative evaluation of pulse width modulation strategies for Z-source neutral-point-clamped inverter," *IEEE Trans. Power Electron.*, vol. 22, no. 3, pp. 1005–1013, May 2007.

[21] Y. Tang, S. Xie, C. Zhang, and Z. Xu, "Improved Z-source inverter with reduced Z-source capacitor voltage stress and soft-start capability," *IEEE Trans. Power Electron.*, vol. 24, no. 2, pp. 409–415, Feb. 2009.

[22] J. Anderson and F. Z. Peng, "Four quasi-Z-source inverters," in *Proc. IEEE Power Electron. Spec. Conf.*, Rhodes, 2008, pp. 2743–2749.

[23] H. Cha, F. Z. Peng, and D. W. Yoo, "Distributed impedance network (Z-network) DC–DC converter," *IEEE Trans. Power Electron.*, vol. 25, no. 11, pp. 2722–2733, Nov. 2010.

[24] Yu Tang, J. Ding, and S. Xie, "An optimal PWM strategy of Z-source inverters," in *Proc. IEEE Energy Convers. Cong. Exp.*, Denver, CO, USA, 2013, pp. 4221–4226.

[25] L. He, S. Duan, and F. Peng, "Safe-commutation strategy for the novel family of quasi-Z-source AC–AC converter," *IEEE Trans. Ind. Informat.*, vol. 9, no. 3, pp. 1538–1547, Aug. 2013.

[26] X. Fang and F. Z. Peng, "Novel three-phase current-fed Z-source AC–AC converter," in *Proc. IEEE Rec. Power Electron. Spec. Conf.*, Orlando, FL, USA, 2007, pp. 2993–2996.

[27] P. C. Loh, D. Li, and F. Blaabjerg, "Γ-Z-source inverters," *IEEE Trans. Power Electron.*, vol. 28, no. 11, pp. 4880–4884, Nov. 2013.

[28] M.-K. Nguyen, Y.-C. Lim, and Y.-J. Kim, "A modified single-phase quasi-Z-source AC–AC converter," *IEEE Trans. Power Electron.*, vol. 27, no. 1, pp. 201–210, Jan. 2012.

[29] M.-K. Nguyen, Y.-G. Jung, and Y.-C. Lim, "Single-phase Z-source voltage sag/swell compensator," in *Proc. IEEE Inter. Symp. Ind. Electron.*, Seoul, Korea, 2009, pp. 24–28.

[30] C. William and T. McLyman, *Transformer and Inductor Design Handbook*, 3rd ed., Boca Raton, FL, USA: CRC Press, 2004.

[31] W. Mo, P. C. Loh, and F. Blaabjerg, "Asymmetrical Γ-source inverters," *IEEE Trans. Ind. Electron.*, vol. 61, no. 2, pp. 637–647, Feb. 2014.

[32] D. Li, P. C. Loh, M. Zhu, F. Gao, and F. Blaabjerg, "Generalized multicell switched-inductor and switched-capacitor Z-source inverters," *IEEE Trans. Power Electron.*, vol. 28, no. 2, pp. 837–848, Feb. 2013.

Mohamad Reza Banaei was born in Tabriz, Iran. He received the M.Sc. degree in control engineering from the Polytechnic University of Tehran, Tehran, Iran, in 1999, and the Ph.D. degree in power engineering from the Electrical Engineering Faculty, Tabriz University, Tabriz, Iran, in 2005.

He is currently an Associate Professor in the Electrical Engineering Department, Azarbaijan Shahid Madani University, Tabriz, Iran, which he joined in 2005. His main research interests include the designing and controlling of power electronic converters, renewable energy systems, modeling and controlling of FACTS and custom power devices, and power systems dynamics.

Rana Alizadeh was born in Iran, in 1990. She received the B.Sc.Eng. degree in electrical engineering from the Azarbaijan Shahid Madani University, Tabriz, Iran, in 2012. She is currently working toward the M.Sc.Eng. degree at the Power Electronics and Renewable Energy Generation Research Lab, Electrical Engineering Department, Azarbaijan Shahid Madani University.

Her current research interests include Z-source, dc–dc, ac–dc power electronic converters, and renewable energy generation systems.

Nazanin Jahanyari was born in 1991, in Iran. She received the B.Sc. degree in electrical engineering from Tabriz University, Tabriz, Iran, in September 2012. She is currently working toward the M.Sc. degree at Smart Distribution Grid Research Lab, Electrical Engineering Department, Azarbaijan Shahid Madani University, Tabriz.

Her research interests include smart distribution networks, micrograds optimal planning, probabilistic modeling in power systems, unit commitment, and ac–ac converter.

Ebrahim Seifi Najmi was born in Tabriz, Iran, in 1986. He received the B.S. degree in electrical engineering from Shahid Madani University, Tabriz, Iran, in 2009, and the M.S. degree in electrical engineering from Tarbiat Modares University, Tehran, Iran, in 2012. He is currently working toward the Ph.D. degree at the Faculty of Electrical and Computer Engineering, Shahid Madani University.

His current research interests include adjustable motor drives, multilevel inverters, switch reduced converters, and FACTS devices.