Quasi-Z-source Inverter with Coupled Inductor

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Abstract— A novel quasi-Z-source inverter with Coupled Inductor technique is proposed in this paper. The boost ability of the proposed topology is enhanced compared to classic Z-source/quasi-Z-source topologies. The proposed topology provides a common mode ground with dc source, continuous input current, reduced voltage and current stress on active and passive components. Better energy utilization and suppressed inrush current are achieved by employing coupled inductor cells. Analytical and simulation results are presented to explore the advantages of the proposed design.

Keywords— Quasi-Z-source Inverter (qZSI), Boost Ability, Coupled Inductor

I. INTRODUCTION

The Z-source inverter (ZSI) has attracted much attention due to its promising advantages compared to the traditional voltage source inverter (VSI). Firstly, single-stage buck-boost inversion of Z-source inverter improves system efficiency compared to the classic double-stage inverter. Secondly, the utilization of shoot-through between upper switches and lower switches of inverter bridges avoids the existence of deadtime which causes distortion in the output current waveform. It also performs better in terms of immunity against EMI noise [1]. However, the main drawbacks of ZSI are high inrush input current, high voltage stress across switches and capacitors and relatively weak boost capability. In addition, the application of ZSI in DC sources is limited due to the discontinuous input current [2]. To overcome these drawbacks, several pulse width modulation (PWM) methods have been proposed to obtain higher boost ability and lower voltage stress on switches, such as maximum boost control [3] and constant maximum boost control [4]. However, the improvement of boost ability is limited, and the discontinuous input current cannot be avoided.

A class of quasi-Z source inverter (qZSI) shown in Fig. 1(a) was first proposed by Peng [5, 6], which has a continuous input current, fewer power devices with lower ratings and common ground between DC source and inverter. However, its practical boost ability is still limited. For this converter, higher voltage gain requires higher shoot-through duty ratio and lower modulation index, which leads to low output power quality and high voltage stress across switches and passive components [1].

To increase the boost capability of qZSI, many attempts have been made in recent years. A class of transformer-based

qZSI, such as trans-qZSI [7] and T-source inverter [8], has been proposed by replacing two separate inductors with magnetically coupled inductors, and the DC-link voltage can be boosted by changing the turns ratio. However, without any snubber circuit, the large switch-voltage spikes are produced by the leakage inductance of the coupled inductor. Improved trans-qZSI, proposed in [9], further upgraded the boost inversion ability and improved the input current profile, like suppression of inrush current at start-up. Nonetheless, more resonances and spikes of DC-link voltage are produced due to the additional inductors and capacitors.

Switched inductor [10, 11], switched capacitor [12, 13], voltage multiplier [14, 15], coupled inductor [16,17] and voltage lift [18,19] techniques are commonly used in transformerless dc-dc converter to obtain high step-up conversion ratio. Some of the techniques have been employed to qZSI [20-24] while maintaining the single-stage inversion. Fig. 1(b) shows a combination of ZSI and switched-inductor (SL) technique (simplified as SL-ZSIs) which has been presented in [20, 21], in order to break the boost limitation of the classic ZSI. The switched-inductor quasi-Z source inverter (SL-qZSI) was also proposed in [22], which has a continuous input current and fewer passive components as well as lower voltage stress on them compared to SL-ZSI. An improved SL-qZSI was proposed in [23] by replacing a diode in the switched-inductor with a bootstrap capacitor, which further increased the boost ability, and decreased the voltage stress on components as well as the current ripple in the input inductor. A combination of a three-winding switched-coupled inductor and switched capacitor technique was employed with the qZSI in the SCL-qZSI, which was proposed in [24]. It obtains the higher boost ability B than that of the existing high voltage gain qZSI and trans-ZSI and suppresses the startup inrush current while retaining all of the advantages of classic qZSI topology. However, the high boost ability of the SCL-qZSI is at the cost of the large count of passive components and smaller operation range, where the shoot-through duty ratio cannot exceed 0.25.

In this paper, a combination of coupled-inductor technique and qZSI, termed as CI-qZSI, is proposed. Compared with the classic qZSI and the SL-qZSI shown in Fig. 1, the proposed inverter achieves a higher boost factor of $2/(1-3D)$, with reduced sizes and number of passive components. It also has a greater current handling capability with a lower input current ripple and a better suppression of startup inrush current. Furthermore, it also achieves lower

voltage and current stresses across components for the same input and output voltage. The operation principle and analysis of the proposed topology is presented. Afterward, the topology is theoretically compared with the classic qZSI and SL-qZSI. The possibility of combining the basic structure of the proposed converter with multiple coupled inductor cells is discussed. Finally, the feasibility of the proposed CI-qZSI is validated by simulations.

Fig. 1. Classic Quasi-Z source inverter topologies.

II. PROPOSED CONVERTER

Fig. 2 shows the circuit of the proposed coupled inductor qZSI (CI-qZSI), which is obtained by the standard qZSI in combination with a coupled inductor based voltage multiplier cell. The proposed inverter consists of two diodes (D_1 and D_2), three capacitors (C_1 , C_2 and C_3), an input inductor L and coupled windings (N_1 and N_2). The turn ratio of windings N_2 to N_1 is n .

A. Circuit Analysis

The operation of the proposed topology is similar to that of classic qZSI, which contains a shoot-through state besides six traditional active states and two zero states. For the purpose of analysis, the operating states are simplified to two states: shoot-through state and nonshoot-through state. Fig. 3 shows the equivalent circuit including the magnetizing inductance L_m and two small leakage inductances L_{lk1} and L_{lk2} of windings N_1 and N_2 .

1) Shoot-Through State

As shown in Fig. 4 (a), both switches of each phase leg in the inverter bridge turn on simultaneously during the shoot-through state. Diodes D_1 is off and D_2 is on. The inductor L is charged by input DC voltage source and capacitor C_2 , while the winding N_1 is charged by capacitor C_1 through diode D_2 . The capacitor C_3 is charged by winding N_2 passing through diode D_2 .

2) Nonshoot-Through State

The nonshoot-through state shown in Fig. 4 (b) consists of traditional six states and two zero states. During this state, the diode D_1 is on and D_2 is off. The input inductor L_1 is discharged to C_1 through D_1 , while the capacitor C_2 is charged by winds N_1 , N_2 and capacitor C_3 . In addition, the energy stored in the leakage inductance of N_1 and N_2 is absorbed by C_2 , therefore switch-voltage spikes are avoided.

B. Step-up Ability of The Proposed Inverter

The boost factor B of the proposed inverter is the voltage gain between the DC-link voltage of the inverter V_{PN} and input dc voltage V_{in} . Similar to the derivation of the boost factor for SCL-qZSI [24] and trans-ZSIs [7], the small leakage inductance is ignored to derive its boost factor. The simplified circuits of the proposed topology during the two states are shown in Fig. 4 (a) and (b), respectively.

During the shoot-through state, the following equations can be derived by applying Kirchhoff's Voltage Law:

$$\begin{cases} V_L = V_{in} + V_{C2} \\ V_{N1} = V_{C1} \\ V_{N2} = V_{C3} \end{cases} \quad (1)$$

The voltage of secondary coupled winding N_2 is directly proportional to that of primary winding N_1 all the time:

$$V_{N2} = nV_{N1} \quad (4)$$

where n is the turn ratio of windings N_2 to N_1 . Therefore, the capacitor voltage V_{C3} is equal to n times V_{C1} .

Similarly, the equations in the nonshoot-through state are derived as:

$$V_L = V_{PN} - V_{in} - V_{C2} \quad (5)$$

$$V_{N1} + V_{N2} = V_{PN} - V_{C1} - V_{C3} \quad (6)$$

$$V_{N1} + V_{N2} = V_{C2} - V_{C3} \quad (7)$$

The shoot-through duty cycle is D and DT is the shoot-through time interval, then $(1 - D)T$ is the nonshoot-through time interval. Applying the voltage-second balance equation on input inductor L , the following equation can be derived according to (1) and (5):

$$V_{PN} = \frac{1}{1 - D} (V_{in} + V_{C2}) \quad (8)$$

Similarly, applying the voltage-second balance condition on winding N_1 , equation (9) can be derived from (2), (4) and (7):

$$V_{C1} = \frac{1 - D}{n + 1} V_{PN} \quad (9)$$

Fig. 2. Proposed qZSI with Coupled Inductor (CI) cell

Fig. 3. Equivalent circuit of the proposed inverter.

(b) Nonshoot-through state

Fig. 4. Equivalent circuit of the proposed topology.

From (6) and (7), the relation between DC-link voltage and capacitor voltage V_{C1} and V_{C2} can be acquired:

$$V_{PN} - V_{C1} = V_{C2} \quad (10)$$

Substituting (9) into (10), the capacitor voltage V_{C2} can be expressed as:

$$V_{C2} = \frac{n+D}{n+1} V_{PN} \quad (11)$$

From (8) and (11), the DC-link voltage V_{PN} across the inverter bridge is:

$$V_{PN} = \frac{n+1}{1-(n+2)D} V_{in} \quad (12)$$

Thus, the boost factor B of the proposed inverter can be expressed as:

$$B = \frac{n+1}{1-(n+2)D}, D < \frac{1}{n+2} \quad (13)$$

C. Generalization of the Basic Proposed Structure

To further increase the boost ability of the proposed structure, multiple secondary windings coupled-inductor cells are added to the classic qZSI as shown in Fig. 5. The relation between DC-link voltage V_{PN} and input DC source voltage V_{in} can be derived as:

$$V_{PN} = \frac{n \cdot N_{CI} + 1}{1-(n+N_{CI}+2)D} V_{in} \quad (14)$$

Where n is the turn ratio and N_{CI} is the number of coupled inductor cells. Therefore, the boost factor B is:

$$B = \frac{n \cdot N_{CI} + 1}{1-(n+N_{CI}+2)D}, D < \frac{1}{n+N_{CI}+2} \quad (15)$$

It can be seen that the range of shoot-through duty ratio D is further decreased with the increasing number of coupled-inductor cells N_{CI} and turns ratio n . To sum up, the generalized structure of the proposed structure can achieve a higher boost ability at the cost of limited operation range and a large number of passive components.

III. COMPARISON WITH PREVIOUS TOPOLOGIES

In this section, qZSI [6] and SL-qZSI [22] are selected to compare with the proposed inverter from several aspects, such as voltage gain, voltage stress and current stress on active switches and passive components and input current ripple.

A. Voltage Gain

When the turns ratio n is equal to 1, the boost factor is $2/(1-3D)$ and the range of duty ratio cannot exceed $1/3$ in order to yield high boost ability. Fig. 6 indicates the boost factor versus the duty ratio for three qZSI-based topologies in simple boost control method. The proposed inverter has a higher step-up ability than that of classic qZSI [6] and SL-qZSI [22].

According to simple boost control in [1], the relation between shoot-through duty ratio D and modulation index M is:

$$D = \frac{T_0}{T} = 1 - M \quad (15)$$

The boost factor B can be expressed by modulation index M as:

$$B = \frac{2}{3M-2} \quad (16)$$

The peak phase-to-phase voltage $\hat{V}_{ph}$ of the inverter is:

$$\hat{V}_{ph} = \frac{MV_{PN}}{2} = \frac{MBV_{in}}{2} \quad (17)$$

Substituting (16) into (17), the voltage gain G (MB) can be derived as:

$$G = \frac{\hat{V}_{ph}}{(\frac{V_{in}}{2})} = \frac{2M}{3M-2} \quad (18)$$

The voltage gain G versus the modulation index M for qZSI-based topologies is shown in Fig. 7. It can be seen that the higher modulation index M is required by the proposed topology for the same voltage gain G due to its high step-up ability. Therefore, compared to the other topologies, the proposed inverter can achieve better output performance by applying a high modulation index when it requires the same voltage conversion ratio.

B. Voltage Stress On Switches

The voltage stress across the switches of inverter bridges V_{SW} is equal to the DC-link voltage V_{PN} , which is in relation with boost factor B , accordingly with modulation index M and voltage gain G . Based on the derivation above, the voltage stress on switches can be expressed as:

$$V_{SW} = BV_{in} = \left(\frac{3G-2}{2}\right) V_{in} \quad (19)$$

Fig. 8 shows the Normalized voltage stress/ Input voltage versus the voltage gain G for three qZSI-based topologies. It can be observed that the proposed topology requires the lowest voltage stress on switches for the same voltage gain G due to its highest step-up ratio.

Fig. 5. Proposed inverter with multiple secondary windings coupled-inductor cells.

C. Voltage Current Stresses On Passive Components

If the duty ratio D and the input voltage V_{in} are fixed, the summary of stresses on components of three different topologies are presented in Table I. In this case, it can be observed that the passive components in the proposed inverter suffer higher voltage current stress than those of the other topologies. It is because of the higher boost ability of the proposed inverter. Specifically, S_d represents the shoot-through state with the value of 1, whereas $\bar{S}_d$ is equal to 1 during the nonshoot-through state.

At the same input voltage V_{in} and the same output phase voltage V_{ph} with fixed voltage gain G , the voltage and current stresses comparison of three topologies are shown in Table II. In this case, lower voltage and current stress on components are achieved by the proposed inverter due to its strong boost ability.

Fig. 6. The boost factor (B) versus the duty ratio (D).

Fig. 7. Voltage gain G versus modulation index M.

Fig. 8. Normalized voltage stress/ Input voltage versus the voltage gain G.

D. Input Current Ripple

Due to the common ground designs of three qZSI-based topologies, all of them have continuous input current. In order to compare the input current ripple in those topologies, the following equations are derived by [24]:

$$\Delta i_{L1} = \frac{V_{L1}}{L} DT \quad (20)$$

Where L is the value of inductance, T is the shoot-through period, Δi_{L1} is the input current ripple, V_{L1} is the voltage on the input inductor L during the shoot-through

state. For the same input voltage V_{in} and fixed voltage gain G, the relation between the value of inductance L_1 and input current ripple Δi_{L1} can be derived by replacing duty ratio D with voltage gain G in Table II:

$$\Delta i_{L1} = \frac{V_{L1}T(G-2)}{L_1(3G-2)} \quad (21)$$

Fig. 9 shows the normalized input current ripples versus inductance in the same input voltage V_{in} and voltage gain G. It is obvious that the lower input current ripple across the input inductor by the same inductance in the proposed inverter. Alternatively, for the same current ripple, the proposed topology requires the lower value and smaller size of the input inductor.

Fig. 9. Normalized input current ripples versus inductance in the same input voltage V_{in} and voltage gain G.

Table I. Summary of stresses with the same input voltage and duty ratio.

	qZSI [6]	SL-qZSI [22]	CI-qZSI (n=1)
V_{PN}	$\frac{1}{1-2D} V_{in}$	$\frac{1+D}{1-2D-D^2} V_{in}$	$\frac{2}{1-3D} V_{in}$
V_C	$V_{C1} = \frac{1-D}{1-2D} V_{in}$ $V_{C2} = \frac{D}{1-2D} V_{in}$	$V_{C1} = \frac{1-D}{1-2D-D^2} V_{in}$ $V_{C2} = \frac{2D}{1-2D-D^2} V_{in}$	$V_{C1} = \frac{1-D}{1-3D} V_{in}$ $V_{C2} = \frac{1+D}{1-3D} V_{in}$ $V_{C3} = V_{C1}$
V_D	$V_D = V_{PN}$	$V_{D1} = V_{PN}, V_{D2} = V_{C1},$ $V_{D3,4} = \frac{D}{1-D} V_{C1}$	$V_{D1} = V_{PN}$ $V_{D2} = \frac{V_{PN}}{2}$
$\widehat{V_{ph}}$	$(1-D) \frac{V_{PN}}{2}$	$(1-D) \frac{V_{PN}}{2}$	$(1-D) \frac{V_{PN}}{2}$
I_{D1}	$\frac{I_{in}}{1-D}$	$\frac{I_{in}}{1-D}$	$\frac{I_{in}}{1-D}$
I_{sh}	$2I_{in}$	$\frac{3+D}{1+D} I_{in}$	$\frac{1+3D}{2D} I_{in}$
I_L	$I_{L1,2} = I_{in}$	$I_{L1} = I_{in}$ $I_{L2} = \frac{I_{in}}{1+D}$	$I_{L1} = I_{in}, I_{m} = I_{in}$ $I_{N1} = S_d \cdot (\frac{1+D}{2D}) I_{in}$ $I_{N2} = S_d \cdot (I_{N1} - I_{in})$ $I_{N1,2} = S_d \cdot (I_{m}/2)$

Table II Summary of stresses with the same input voltage V_{in} and output phase voltage V_{ph} .

	qZSI [6]	SL-qZSI [22]	CI-qZSI (n=1)
D	$\frac{G-1}{2G-1}$	$\frac{G - \sqrt{2G^2 - 2G + 1}}{1-G}$	$\frac{G-2}{3G-2}$
V_{PN}	$(2G-1)V_{in}$	$\frac{(1-G)(1-\sqrt{2G^2-2G+1})}{-2G+2\sqrt{2G^2-2G+1}} V_{in}$	$\frac{3G-2}{2} V_{in}$
V_C	$V_{C1} = GV_{in}, V_{C2} = (G-1)V_{in}$	$V_{C1} = \frac{(1-G)(1-2G+\sqrt{2G^2-2G+1})}{-2G+2\sqrt{2G^2-2G+1}} V_{in}, V_{C2} = (G-1)V_{in}$	$V_{C1} = \frac{G}{2} V_{in}, V_{C2} = (G-1)V_{in}, V_{C3} = V_{C1}$
$\widehat{V_{ph}}$	$GV_{in}/2$	$GV_{in}/2$	$GV_{in}/2$
I_{D1}	$\frac{(2G-1)}{G} I_{in}$	$\frac{1-G}{1-G+\sqrt{2G^2-2G+1}} I_{in}$	$\frac{3G-2}{2G} I_{in}$
I_{sh}	$2I_{in}$	$\frac{3-2G-\sqrt{2G^2-2G+1}}{1-\sqrt{2G^2-2G+1}} I_{in}$	$\frac{3G-4}{G-1} I_{in}$
I_L	$I_{L1,2} = I_{in}$	$I_{L1} = I_{in}$ $I_{L2} = \frac{(1-G)I_{in}}{1-\sqrt{2G^2-2G+1}}$	$I_{L1} = I_{in}, I_{m} = I_{in}$ $I_{N1} = S_d \cdot (\frac{2G-2}{G-2}) I_{in}$ $I_{N2} = S_d \cdot (I_{N1} - I_{in}), I_{N1,2} = S_d \cdot (I_{m}/2)$

IV. SIMULATION RESULTS

To verify the theoretical analysis, the simulation of the proposed inverter is employed in Simulink/ Matlab. Table III presents the specifications of the proposed topology. The input dc voltage is selected as 150V and the output phase-to-phase voltage is $240V_{RMS}$ to meet the grid-tied requirement. The values of passive components are selected as $L=700\mu H$, $C_1=C_2=C_3=200\mu F$, and the magnetizing inductance is $100\mu H$ and leakage inductance is $0.2\mu H$. The output filter inductors and capacitors are $1mH$ and $20\mu F$ and a three-phase resistive load of $30\Omega/\text{phase}$ is selected. Specifically, the simple boost control method is employed in all of the qZSI-based topologies. Voltage gain G of 4.52 is needed in simple boost control, therefore the values of parameters are theoretically calculated from Table III, $D=0.22$, $B=5.78$, $V_{PN}=867V$, $V_{C1}=339V$, $V_{C2}=528V$, and $V_{C3}=339V$.

Table III Specifications of the proposed inverter.

Input voltage	150 Vdc
Output phase voltage	$240 V_{RMS}$
Z-source capacitor (C_1, C_2, C_3)	$200 \mu F$
Z-source inductor (L)	$700 \mu H$
Coupled inductor ($L_{N1}, L_{N2}, n=1$)	$L_m = 100 \mu H, L_{lk} = 0.2 \mu H$
Output filter inductor (L_f)	$1 mH$
Output filter capacitor (C_f)	$20 \mu F$
Switching frequency (f)	$10 kHz$
Three-phase resistive load (R_l)	$30 \Omega/\text{phase}$

Fig. 10. Simulation results of the proposed inverter. (a) Three-phase output phase-to-phase voltage and current. (b) Capacitor voltage and dc-link voltage.

Fig. 11. Current stress across components in the proposed inverter. (a) Input inductor current, magnetizing current, and shoot-through current. (b) Winding currents and diode currents.

Fig. 10 present the simulation results of the proposed inverter. It can be seen that capacitor voltage C_1, C_2, C_3 are 338V, 520V, 332V and DC-link voltage V_{PN} is 860V, which match the theoretical results. The current across components is shown in Fig. 11.

When the simple boost control is applied to produce the same input (200V) and output voltage ($240 V_{RMS}$), the comparison of voltage stress on inverter switches in three qZSI-based inverters is shown in Fig. 12. In this case, the modulation index M of the classic qZSI, SL-qZSI, and the proposed CI-qZSI are 0.6, 0.69 and 0.83 respectively. It can be observed that lower voltage stress across switches in the proposed inverter (796V) compared to those in the qZSI (1062V) and the SL-qZSI (889V). Furthermore, the spike voltage in the proposed inverter (1192V) is also much lower than those in the other qZSI-based topologies (qZSI with 2027V and SL-qZSI with 1613V).

Fig. 13 indicates the input current across the input inductor. The proposed inverter obtains the lowest input inrush current of 231A and the smallest input current ripple of 8.3A, while the input inrush current of 542A and input current ripple of 18.5A in the qZSI and the input inrush current of 418A and input current ripple of 13.0A in the SL-qZSI. In conclusion, the proposed inverter has better current handling capability and lower voltage stress on switches due to its higher boost ability.

Fig. 12 Simulation results of DC-link voltage in (a) CI-qZSI. (b) SL-qZSI and (c) qZSI.

Fig. 13 Simulation results of input current in (a) CI-qZSI. (b) SL-qZSI and (c) qZSI.

V. CONCLUSIONS

This paper proposed a CI-qZSI with the following characteristics: high boost voltage ability, common ground between DC source and inverter and continuous input current. For the same input and output voltage, the proposed CI-qZSI provides the lower voltage stress on capacitors and active switches, reduced current stress on the inductor. Furthermore, it offers better current handle capability, such as suppression of inrush input current and reduced current ripple.

The simulation results have been verified the proposed inverter has higher boost inversion ability and reduced voltage stress across switches compared to the classic qZSI [6] and SL-qZSI [22] for dc 200V and ac 240V_{rms} phase-to-phase output.

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