Zero-current-switching switched-capacitor bidirectional DC–DC converter

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Abstract: The proposed zero-current-switching switched-capacitor quasi-resonant DC–DC converter is a new type of bidirectional power flow control conversion scheme. It possesses the conventional features of resonant switched-capacitor converters: low weight, small volume, high efficiency, low EMI emission and current stress. A zero-current-switching switched-capacitor step-up/step-down bidirectional converter is presented that can improve the current stress problem during bidirectional power flow control processing. It can provide a high voltage conversion ratio using four power MOSFET main switches, a set of switched capacitors and a small resonant inductor. The converter operating principle of the proposed bidirectional power conversion scheme is described in detail with circuit model analysis. Simulation and experiment are carried out to verify the concept and performance of the proposed bidirectional DC–DC converter.

1 Introduction

The switched-capacitor DC–DC converter is a non-magnetic converter that requires only capacitor and MOSFET switches in the power stage. This converter has the following features: light weight, smaller size and fabrication on a semiconductor integrated circuit chip. However, the larger switching loss and current stress are the essential drawbacks in the conventional high-frequency switching DC–DC converter [1–3]. Quasi-resonant converters that are able to operate at constant switching frequency with zero-current or zero-voltage switching (ZCS or ZVS) have been used to reduce the switching loss in the converter to overcome the aforementioned problems. This paper presents a new zero-current-switching switched-capacitor quasi-resonant (ZCS SC QR) converter that can operate at high switching frequency with less switching loss for increased converter efficiency with fewer switches [2, 3]. Although the ZCS SC QR converter has numerous advantages, its power flow control moves only in the unidirectional direction.

Bidirectional DC–DC power conversion is of great interest in systems fed by DC power, including electric vehicle hybrid energy systems, fuel-cell systems and aerospace systems [4, 5]. The bidirectionality in these applications involves current flow while the polarity of the DC voltage at either end remains uncharged. Application in battery equalisation schemes where the stronger energy of this system is transferred into the weaker energy subsystem using the bidirectional power flow control scheme [6]. A class of soft-switching bidirectional DC–DC converters is the expected candidate for applications such as uninterruptible power supplies (UPS), battery charging and discharging systems, auxiliary power supplies in HEV, and dual voltage automotive systems [4, 6]. A switched-capacitor-based bidirectional DC–DC converter providing the capability of step-down/step-up voltage is proposed in [7]. The input/output current pulsating peaks of the bidirectional SC converter are mitigated, but the switching losses and the overall converter efficiency are not significantly improved [8, 9].

This paper presents a new switched-capacitor step-up/step-down DC–DC quasi-resonant bidirectional converter designed to operate at noninverting mode and constant frequency. The proposed converter scheme can achieve zero-current soft switching and reduce the MOSFET switching loss to increase the converter efficiency. The high switching current stresses can also be reduced under the bidirectional power flow control scheme [10]. The proposed converter can be applied to many other topologies for varying voltage conversion ratios under suitable arrangement and designed control strategy of the switched-capacitor networks. A design example for the triple-mode/tri-section-mode (three-mode/4-mode) ZCS SC QR converter was conducted to verify and validate the predicted performance under bidirectional power flow control. The EMI emissions in the power line source can also be suppressed because of the soft switching and the source current waveforms along the approximate sinusoidal function.

2 Topologies description

Figure 1 shows the proposed noninverting type triple-mode/tri-section-mode ZCS SC QR bidirectional converter that was developed based on the ZCS SC QR converter. The circuit is composed of four MOSFET main switches paralleled with Schottky diodes. Only a very small inductor series connected with a set of switched capacitors is needed to construct the resonant tanks in the converter. A resonant inductor \( L_r \) is connected in series with the switching capacitor comprised of \( C_1 \) and \( C_2 \) to achieve a resonant cycle when each of the switches \( Q_{1S} \) (contains \( Q_{1P} \) and \( Q_{1N} \)) or \( Q_{2S} \) (contains \( Q_{2P} \) and \( Q_{2N} \)) are switched on during the operating interval. The switches can be designed to switch...
on and off at zero-current state with the \( L_r - C \) resonant current rising and falling to zero to achieve zero current switching for reducing the MOSFET switching loss. The switches \( Q_a, Q_b \) and \( Q_{ab} \) are used to control the switched capacitors for parallel or series connection during the charging or discharging states. Figure 2 shows a noninverting type step-up ZCS SC QR DC–DC converter under various operation stages. Turn-on/off switches \( Q_{1P} \) or \( Q_{1N} \) and the switched-capacitor switches \( Q_a, Q_b \) or \( Q_{ab} \) can control the forward power flow from the source \( V_1 \) to the other source \( V_2 \) as a triple-mode converter (i.e. \( V_2 = 3 \times V_1 \)) shown as Fig. 2. The turn-on/off switches \( Q_{2P} \) or \( Q_{2N} \) and the switched-capacitor switching diodes \( D_a, D_b \) or \( D_{ab} \) can also control the reverse power flow from \( V_2 \) to \( V_1 \) as a trisection-mode converter (i.e. \( V_1 = V_2/3 \)), shown in Fig. 3. When the resonant current \( I_{Lr} \) increases to a peak value and decreases to zero current, it cannot reverse into negative current because there is a diode in the resonant circuit loop of the converter that ceases current reversal. Figures 2a–2d and Figs. 3a–3d show the equivalent circuit of the proposed bidirectional converter under the forward and reverse power flow control scheme.

The proposed converter topology can be extended to operate as a noninverting type \( n \)-mode/\( \frac{1}{n} \)-mode and an inverting type double-mode/half-mode zero-current-switching switched-capacitor quasi-resonant DC–DC converter for reconstructing the switched-capacitor configuration in the capacitor bank. The following Sections of this paper analyse the proposed noninverting type triple-mode/trisection-mode ZCS SC QR bidirectional DC–DC converter.

### 3 Circuit analysis

The equivalent circuit in Fig. 1 for the proposed noninverting type ZCS SC QR bidirectional converter is shown in Figs. 2a–2d and Figs. 3a–3d under the forward and reverse power flow control scheme for the various operating time intervals. Some assumptions are necessary to analyse the dynamic circuit behaviour: (i) the resonant inductor \( L_r \) has a small inductance, neglecting internal resistance; (ii) the semiconductor switches and diodes are ideal; and (iii) the capacitors in the switched-capacitor bank are ideal. Therefore, the resonant current \( I_{Lr} \) is controlled by the MOSFET switches \( Q_{1S} \) and \( Q_{2S} \) for the forward and reverse power flow control modes, respectively. The direction of the controlled power flow is determined according to the

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**Fig. 1** Triple-mode/trisection-mode ZCS SC QR bidirectional DC–DC converter

**Fig. 2** Equivalence circuit of various operation stages for ZCS SC QR bidirectional DC–DC converter with forward power flow:

- **a** \( t_0 < t < t_1 \)
- **b** \( t_1 < t < t_2 \)
- **c** \( t_2 < t < t_3 \)
- **d** \( t_3 < t < t_4 \)
switched-capacitor charging/discharging situation and/or the desired power flow control orientation for the converter system. The corresponding switching waveforms of the ZCS SC QR bidirectional converter are also shown in Figs. 4a and b under various power flow control schemes. The detailed equivalent circuits for the proposed converter scheme are shown in Figs. 2a–2d at different time intervals under forward power flow control mode. The state space dynamic of all the operation modes are easily analysed in the following stages:

Stage 1 (Fig. 2a; \(t_0<t_1\)): The main switch Q1P and switched-capacitor switches Qa, Qb are turned on at \(t=t_0\), the source \(V_1\) provides current though MOSFET switch Q1P, D1P, the paralleled \(C_1\) and \(C_2\), and the resonant reactance \(L_r\) stores the electric energy in the capacitors \(C_p = C_1 + C_2 = 2C\). The dynamic state equation in this interval is given by

\[
\begin{bmatrix}
\frac{dI_{L_r}(t)}{dt} \\
\frac{dV_C(t)}{dt}
\end{bmatrix} = \begin{bmatrix}
0 & -\frac{1}{L_r} \\
\frac{1}{C_p} & 0
\end{bmatrix} \begin{bmatrix}
I_{L_r}(t) \\
V_C(t)
\end{bmatrix} + \begin{bmatrix}
\frac{1}{L_r} \\
0
\end{bmatrix} V_1
\]

(1)

with the initial conditions \(I_{L_r}(t_0) = 0\) and \(V_{C1}(t_0) = V_{C2}(t_0) = V_{C01}\). The solutions of (1) can be obtained as

\[
I_{L_r}(t) = \frac{V_1 - V_{C01}}{Z_r} \sin \omega_1 t
\]

(2)

\[
V_C(t) = V_1 + (V_{C01} - V_1) \cos \omega_1 t
\]

(3)

where the resonant angular frequency in stage 1 is \(\omega_1 = 1/\sqrt{L_r C_p}\), the normalised impedance is \(Z_1 = \sqrt{L_r / C_p}\). The inductor current \(I_{L_r}(t)\) will increase to reach the peak value, and substantially decrease to zero at \(t=t_1\).

Stage 2 (Fig. 2b; \(t_1<t_2\)): The main switch Q1P and switched-capacitor switches Qa, Qb are still turned on, and the diode D1P cannot reverse into negative current, which ceases the current reversing in this interval. The inductor current will still cease at zero state until \(t=t_2\). The states in this stage are \(I_{L_r}(t) = 0\) and \(V_{C1}(t) = V_{C2}(t) = V_C(t) = V_{C01}\).

Stage 3 (Fig. 2c; \(t_2<t_3\)): When Q1N and Qa are turned on at \(t=t_2\), \(V_1\), \(C_1\) and \(C_2\) are series connected with the switches, Q1N and Qa, and the resonant inductor \(L_r\). Therefore, the stored energy is transferred into the source \(V_2\) through the \(L_r - C_1\) resonant tank circuit. The dynamic state equation in this interval can be expressed by

\[
\begin{bmatrix}
\frac{dI_{L_r}(t)}{dt} \\
\frac{dV_C(t)}{dt}
\end{bmatrix} = \begin{bmatrix}
0 & \frac{2}{L_r} \\
-\frac{1}{C} & 0
\end{bmatrix} \begin{bmatrix}
I_{L_r}(t) \\
V_C(t)
\end{bmatrix} + \begin{bmatrix}
\frac{1}{L_r} \\
0
\end{bmatrix} V_1
\]

(4)

where \(C = C_1 = C_2\), the resonant angular frequency in stage 3 is \(\omega_3 = 1/\sqrt{L_r C_3}\), \(Z_2 = C_3/\sqrt{L_r C_3}\). The initial conditions of the different equations are \(I_{L_r}(t_2) = 0\), and \(V_{C1}(t_2) = V_{C2}(t_2) = V_{C02}\). The solutions of (4) can be obtained as

\[
I_{L_r}(t) = \frac{V_1 - V_2 - V_{C02}}{Z_2} \sin \omega_3 t
\]

(5)

\[
V_C(t) = V_{C02} + (V_1 - V_2 - V_{C02})(\cos \omega_3 t - 1)
\]

(6)
The inductor current will decrease along the sinusoidal function to the negative peak value and continuously fall to zero at \( t = t_3 \). After \( I_{Lr} \) resonates back to zero, diodes \( D_{1N} \) and \( D_{1P} \) are biased in reverse and this operation stage is terminated.

Stage 4 (Fig. 2d; \( t_3 < t < t_4 \)): In this interval, the inductor current will stay in zero current state, the states are \( I_{Lr}(t) = 0 \) and \( V_{C1}(t) = V_{C2}(t) = V_C(t) = V_{C0} \). After a specified time, \( Q_{1N} \) and \( Q_a \) are turned off at zero current state when \( t = t_4 \). Let the duty ratio of \( Q_{1P} \) and \( Q_{1N} \) be denoted by \( l_{1P} \) and \( l_{1N} \), respectively. The zero current switching conditions of \( Q_{1P} \) and \( Q_{1N} \) are then obtained as \( l_{1P}T_S \geq \pi/a_1 \) and \( l_{1N}T_S \geq \pi/a_2 \), where \( T_S \) is the switching duty of the proposed bidirectional converter.

Figures 3a–3d show the alternating equivalent circuits of the proposed bidirectional converter under the reverse power flow control scheme. In stage 1 (Fig. 3a; \( t_0 < t < t_1 \)), when \( Q_{2P} \) is turned on at \( t = t_0 \), then \( D_{1N} \) and \( D_{ab} \) are forced to turn on, the sources \( V_1 \) and \( V_2 \), \( C_1 \) and \( C_2 \) are series connected with the switches, \( Q_{2P} \) and \( D_{ab} \), and the resonant inductor \( L_r \). Therefore, the energy is stored to the \( C_1 \) and \( C_2 \) through the \( L_r-C_S \) resonant tank circuit. The dynamic state equation of stage 1 is the same as (4) with the negative state variables \( I_{Lr}(t) \) and \( V_C(t) \). In stage 3 (Fig. 3c; \( t_2 < t < t_3 \)), the main switch \( Q_{2N} \) is turned on at \( t = t_2 \), then forced to turn on the diodes \( D_{2N}, D_a \) and \( D_b \). The stored electric energy in the parallel capacitors \( C_p = C_1 + C_2 = 2C \) is discharged to the source \( V_1 \) through the switch \( Q_{2N} \) and diodes \( D_{2N}, D_a \) and \( D_b \). The dynamic state equation in this interval is same as (1) with the negative states. The typical switching waveforms of the bidirectional converter are shown in Fig. 4b. The analytical procedure is similar to that for the bidirectional converter in the forward power flow control and is omitted in this paper.

4 Experimental results

To validate the performance of the proposed ZCS SC QR bidirectional DC–DC converter, a PSpice simulation and experiment were carried out for the proposed triple-mode/trisection-mode noninverting ZCS SC QR...
DC–DC converter. The designed parameters are as follows: MOSFET switches are IRF3710, Schottky diodes are SBL1060, $L_r = 1 \mu\text{H}$, $C_1 = C_2 = C = 0.33 \mu\text{F}$, $f_s = 200\ \text{kHz}$, the duty ratios are $\lambda_{1P} = 0.56$ and $\lambda_{1N} = 0.43$ and $P_0 = 60\ \text{W}$. The input voltages are $V_1 = 14\ \text{V}$ and $V_2 = 42\ \text{V}$ for forward and reverse power flow control, respectively. Figures 4a and b show the simulation results of the proposed triple-mode and trisection-mode ZCS SC QR converter, respectively. The output voltage of the proposed converters are slightly less than the designed values due to the ESR drop of the active and passive devices in the converter loop. Figures 5a and b show the experimental results of the corresponding waveforms and converter efficiency of the proposed converter under forward and reverse power flow control mode. The measured output voltages of the proposed converter $V_o$ are 36.3 and 12.4 V under the various directional modes, respectively. The measurements are slightly less than the designed values owing to the practical voltage drop of the active and passive devices. The average efficiency of the forward and reverse power flow control converter under various load conditions is high—more than 90%.

Observations on the proposed ZCS SC QR bidirectional converter can be summarised as follows:

- The power MOSFETs of the proposed bidirectional converter are turned on and off in the zero-current state. The total switching loss and EMI emission can be significantly reduced, and the switching frequency can be further increased and the size of capacitor can be decreased.
- The current stresses of the MOSFETs are significantly reduced compared with the conventional switched-capacitor bidirectional converter.
- The maximum efficiency achieved can be about 95 and 93% for the forward and reverse power flow control schemes, respectively.
- For high-power application, only a small magnetic component is required in the bidirectional converter. Therefore, the core size and core losses are also be reduced.
- The proposed converter scheme design can be extended to a noninverting type $\frac{1}{2}$-mode bidirectional DC–DC converter for high voltage conversion ratio power supply applications, as shown in Fig. 6. And the concept

![Experimental waveforms and efficiency of ZCS SC QR bidirectional converter](image)

*Fig. 5 Experimental waveforms and efficiency of ZCS SC QR bidirectional converter*

a Forward power flow control
b Reverse power flow control
can be applied to many other topologies, such as an inverting type double-mode/half-mode bidirectional converter for battery equalisation applications [6] shown in Fig. 7.

- The proposed converter scheme is designed to operate as a non-isolation type triple-mode/trisection-mode bidirectional DC–DC converter for high voltage conversion ratio power supply applications. The inductor current is operated in an unsymmetrical alternating mode, the reversing current will automatically demagnetise the residual flux in the magnetic core of the inductor choke to avoid the risk of saturation due to cumulative magnetic flux in the core. There is a small DC current component flowing through the inductor that could produce a saturation problem in the choke, which will be considered in selecting high flux magnetic toroidal core (CH270060) for the inductor design procedure. The maximum output power and voltage conversion ratio of the proposed converter have some limitations, subject to the saturation constraint for the magnetic design of the resonant inductor.

5 Conclusions

A triple-mode/trisection-mode power conversion scheme based on quasi-resonant zero-current-switching converter theory for a bidirectional converter has been studied. It was found that the proposed unit possesses interesting properties such as reducing the switching loss, increasing the converter efficiency, and significantly reducing the MOSFET current stress of the converter. Simulations and experimental results were performed to verify the theoretical analysis. It was shown that the proposed ZCS SC QR converter is suitable for high-frequency and high-efficiency bidirectional power converter applications.

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7 References