A multiplicator ZCS turn on boost converter with high-efficiency and high-voltage-gain

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Abstract This paper presents a multiplicator ZCS turn on boost converter with high-efficiency and high-voltage-gain. A resonant multiplicator structure combined with two-phase technique is provided. Based on this structure, the problem of switches turn-on loss and inductor current ripple is minimized. The voltage resonant multiplicator also shows a good scalability. And the presented prototype is well adapted for multiphase dc–dc converters. The converter is tested for an application requiring the output power of 100W~500W, operating with 12V input voltage and 220V output voltage. The measured peak efficiency equals to 95.4% with the gain of 18.333.

Keywords: resonant multiplicator, high gain, high efficiency, good scalability;
Classification: Electron devices, circuits and modules (silicon, compound semiconductor, organic and novel materials)

1. Introduction

Nowadays, global warming resulting from carbon emissions is a severe challenge faced by humanity. The importance of renewable energy has increased many folds [1,2]. A high step-up gain and high efficiency converter is one of the critical points of energy conversion. Mostly conventional boost converter is used to increase the output voltage from low input voltage systems. But the main disadvantage of boost converters is that it cannot work in high ration duty cycle. Therefore, many solutions are proposed.

The voltage multiplicator composed of traditional diodes and capacitors [3, 4, 5], combined with traditional boost [6, 7, 8, 9], forward, flyback and buck-boost [10] circuits, can reduce the voltage stress of the devices and increase the voltage gain. Adding a transformer [11, 12, 13, 14, 15, 16, 17, 18, 19, 20, 21, 22, 23] to the classical circuit achieves the same effect. But the commutation loss has not been reduced. Ying [24, 25] and Kerui Li [26] used boost resonant inverter cascaded voltage doubler rectifier to achieve boost. But its zero-voltage turn-on (ZVS) reduces switching losses, its device voltage and current are far greater than traditional circuits and its actual efficiency is low. Lin [27] used a dual-phase boost circuit to reduce the inductor current stress. But its commutation loss is not reduced.

In order to reduce switching losses, while satisfying high gain and low voltage and current stress, a multiplicator composed of inductors and diodes and capacitor ZCS boost converter is presented in the paper. The converter realizes the dual-tube ZCS turn-on at the same time.

2. Operating principle of proposed converter

Fig. 1 represents a use of the classical interleaved boost converter connected with the inverting VM cells [3].

Fig. 1 The classical interleaved boost converter connected with the VM cells

In order to reduce the converter commutation losses, allowing the operation with low voltage and current stress, maintaining high efficiency, the converter is designed is shown in Fig.2. The structure of the provided resonant multiplicator cell is composed by the diodes (D_b1, D_a11, D_a12), the capacitors (C_b1, C_a12, C_a11) and the resonant inductor L_r. The inclusion of this inductance L_r allows the switch to operate with ZCS turn-on.

Fig. 2 The new resonant multiplicator cell integrated with two-phase boost converter

The working process of the circuit is composed of

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six stages. Better operation characteristics are obtained when the converter operates in continuous conduction mode (CCM). Thus, the operation stages (Figs.3) and the theoretical waveforms (Fig.4) are presented for CCM operation.

![Operating circuit diagrams of proposed converter.](image)

Fig. 3 Operating circuit diagrams of proposed converter.
(a) First Stage t_1-t_2; (b) Second Stage t_2-t_3; (c)Third Stage t_3-t_4; (d) Fourth Stage t_4-t_5; (e) Fifth Stage t_5-t_6; (f) Sixth Stage t_6-t_11

![The main theoretical waveforms of the proposed converter.](image)

Fig. 4 The main theoretical waveforms of the proposed converter. During t_1-t_2, the tube S_6 has just turned on, and S_6 is in the conduction state before. According to the law of conservation of charge \(dU=\Delta Q/C\), the voltage of the capacitor \(C_{b1}\) and \(C_{a12}\) is expressed by (1) and (2). At the instant \(t_0\), the resonant inductor current \(i_{L_r}\) is equal to the inductor \(L_b\) current. The current in the tube \(S_b\) is zero and the tube \(S_b\) realizes ZCS turn-on.

\[
V_{C_{b1}}(t) = V_{C_{b1}}(t_0) - \int_{t_0}^{t} \frac{i_{L_r}(t)}{C_{b1}} dt \\
V_{C_{a12}}(t) = V_{C_{a12}}(t_0) + \int_{t_0}^{t} \frac{i_{L_r}(t)}{C_{a12}} dt
\]

where \(i_{L_r}\) is expressed by

\[
i_{L_r}(t) = i_{L_b}(t_0) \cdot \frac{(V_{C_{a12}}(t) - V_{C_{b1}}(t))}{L_r} \\
\]

In the t_1-t_3 stage, the tube \(S_b\) turns on, and \(S_a\) is also in the on state before. After the resonant current reaches zero, the current oscillation period is as shown in the following formula (4). The leakage inductance energy oscillates in the parasitic capacitance of \(L_r\) and \(D_{a11} (C_{T_{D11}})\) added the parasitic capacitance of \(D_{a12} (C_{T_{D12}})\). Since the oscillation amplitude is much smaller than the operating current, the voltage values of \(C_{a12}\) and \(C_{a11}\) are approximately unchanged.

\[
t_{on} = \frac{2\pi}{\sqrt{L_r \cdot C_{eq}}} \\
C_{eq} = C_{T_{D11}} + C_{T_{D12}}
\]
During t2-t3, the tube $S_b$ turns off and the tube $S_a$ turns on, and the resonant inductor current $i_{L_r}$ increases linearly in the reverse direction. In the meanwhile, output diode ($D_o$) is freewheeling.

$$V_{Ca3}(t) = V_{Ca3}(t_2) + \int_{t_2}^{t} \frac{i_{L_r}(t)}{C_{b1}} dt$$  \hspace{1cm} (7)

$$V_{Ca2}(t) = V_{Ca2}(t_2) + \int_{t_2}^{t} \frac{i_{L_r}(t)}{C_{a12}} dt$$ \hspace{1cm} (8)

where $i_{L_r}$ and $i_{Ca}$ in the formula (7)-(8) can be obtained from the following formula.

$$i_{L_r}(t) = -\frac{V_o - V_{Ca2}(t) + V_{Ca3}(t) - V_{Ca1}(t)}{L_r}$$ \hspace{1cm} (9)

In the t1-t3 stage, the tube $S_a$ has just turned on before $S_b$ is in the conduction state. The diode $D_a$ is reversely cut off and $D_{a12}$ linearly continues to flow to zero as defined by. The working status is similar to the t0-t1 stage.

$$V_{Ca3}(t) = V_{Ca3}(t_3) + \int_{t_3}^{t} \frac{i_{La}(t)}{C_{a1}} dt$$ \hspace{1cm} (10)

where $i_{La}$ in the formula (10) can be obtained from the following formula.

$$i_{La}(t) = i_{La}(t_3) - \frac{V_{Ca2}(t) - V_{Ca3}(t)}{L_a}$$ \hspace{1cm} (11)

In the t3-t4 stage, the working status is similar to the t1-t2 stage.

$$V_{Ca3}(t) = V_{Ca3}(t_4) + \int_{t_4}^{t} \frac{i_{La}(t)}{C_{b1}} dt$$ \hspace{1cm} (12)

$$V_{Ca2}(t) = V_{Ca2}(t_4) + \int_{t_4}^{t} \frac{i_{La}(t)}{C_{a12}} dt$$ \hspace{1cm} (13)

where $i_{La}$ and $i_{Ca}$ in the formula (13) can be obtained from the following formula.

$$i_{La}(t) = i_{La}(t_4) - \frac{D_{a}V_o / (1 - D_a)}{L_a}$$ \hspace{1cm} (14)

3. Steady state analysis of the proposed converter

3.1 DC conversion ratio and circuit stress

From the working state t3-t4, the voltage of the capacitor $C_{a1}$, $C_{a2}$ can be derived as

$$V_{Ca2} = V_{in} \frac{1}{1 - D_a}$$ \hspace{1cm} (23)

$$V_{Ca2} = V_{in} \frac{1}{1 - D_a} + V_{b1}$$ \hspace{1cm} (24)

Similarly from the working state t2-t3, the voltage of the capacitor $C_{b1}$ can be derived as

$$V_{Cb3} = V_{in} \frac{1}{1 - D_b} + V_{b1} = V_{in} \frac{1}{1 - D_a} + V_{b1} \frac{1}{1 - D_b}$$ \hspace{1cm} (25)

Hence the gain expression $G$ of the circuit is as follows.

$$G = \frac{V_{out}}{V_{in}} = \frac{2}{1 - D_a} + M \frac{1}{1 - D_b}$$ \hspace{1cm} (26)

where $D_a$, $D_b$ are switch duty-cycles.

Voltage stress: the maximum voltage applied across the power switches ($S_a$ and $S_b$) is equal to the output diodes ($D_{a1}$, $D_{a12}$, $D_{a2}$, $D_{b}$) reverse voltage (assuming duty cycle $D_{a1}=D_{a2}=D$).

$$V_{Sa} = V_{Sb} = V_{b1} = V_{b12} = V_{b0} = V_{in} \frac{1}{1 - D}$$ \hspace{1cm} (27)

Table I shows that the voltage gains and the voltage stresses of the proposed converter. It is nearly to other structures.

| Ref. | Normalized voltage stress across the power | Voltage gain $G$ |
|------|------------------------------------------|-----------------|
| 28   | $(G+1)/(4G)$                             | $(3+D)/(1-D)$   |
| 29   | $(G+1)/(4G)$                             | $(1+3D)/(1-D)$  |
| 30   | $(G+1)/(4G)$                             | $(3+D)/(1-D)$   |
|      | work                                     | 0.333           |

Fig.5(a) illustrates the voltage gain curves versus duty cycle. Considering Table I and Fig. 5(a), the voltage gain of the proposed converter is higher than the boost structures and nearly equal to [28]-[30].
The normalized voltage stresses across the power switches are plotted in Fig.5(b). The voltage stresses across the power switches are almost equal to the related structures.

3.2 Component selection

Passive components: The design of the input inductance is based on its current ripple, considering the efficiency equal to 92%.

\[
I_w = \frac{P_o}{V_{in} \cdot \eta} = 45.29 \text{A}
\]

\[
L_o = L_a = \frac{P \cdot V_o \cdot D \cdot \Delta t_{L_a} \cdot f}{V_{C1}} = 7.38 \mu H
\]

The minimum capacitance of the voltage multiplier capacitor depends on the maximum output power, the multiplier capacitor drop voltage \(\Delta V\), (0.1 times output voltage) and the switching frequency \(f_s\).

\[
C \geq \frac{P_{\text{out}}}{0.5AV_c \cdot f_s} = 27.8 \mu F
\]

Resonant inductor: The resonant inductor can be defined by the maximum current variation (di/dt) at the turn-on commutation. In the operation stage \((t_{a1} - t_{b1})\), the reduction of the resonant inductor current occurs at the switch turn-on. The current variation is limited by the presence of the resonant inductor, defined by

\[
\frac{di}{dt} = \frac{V_{\text{Cap2}} - V_{\text{Cap1}}}{L_r}
\]

Considering the maximum resonant inductor variety at the \(S_o\), turn-on commutation equal to 100A/\mu s, \(L_r\) value is defined by

\[
L_r = \frac{V_{\text{Cap2}} - V_{\text{Cap1}}}{\frac{di}{dt}} = 0.6 \mu H
\]

Table II shows the parameters of this work according to the formula (23)-(32).

| Parameters and Symbols | This work |
|------------------------|-----------|
| Input Voltage: \(V_{in}/\text{V}\) | 12 |
| Output Voltage: \(V_{C1}/\text{V}\) | 220 |
| Output power: \(P_{out}/\text{W}\) | 500 |
| Switching Frequency: \(f_s/\text{kHz}\) | 300 |
| Number of multiplier stages: \(M\) | 1 |
| Parallel stages: \(P\) | 2 |
| \(L_a\), \(L_o\) | L=10\mu H, DCR=0.88m |
| \(S_a\), \(S_b\) | IXFA130N15X3 (\(V_{\text{fem}} = 150\text{V}, R_{\text{on}} = 9\text{m}\)) |
| \(L_r\) | L=0.68\mu H, DCR=1.4m |
| \(D_{a1}, D_{a2}, D_{a3}, D_o\) | DSS 6-015AS (\(V=150\text{V}, V_f=0.62\text{V}\)) |
| \(C_{a11}\) | CKG57NX7S2A226M500JH (\(C=22\mu F\times5\)) |
| \(C_{b11}, C_{a12}\) | C4AJLBW5900M30K (\(C=90\mu F\)) |

3.3 Component losses and theoretical efficiency

The switching loss is divided into turn-on loss and turn-off loss. Turn-on loss is the most important factor affecting the efficiency of the circuit. The commutation loss is reduced in the proposed converter because the converter is in ZCS state. The switch current \(I_{\text{Sat(off)}}\) is 20.833A at turn off time \((t_{\text{off}}=62\text{ns})\) and the conduction resistance \(R_{\text{on}}\) of the tube \(S\), \(S_o\) is equal to 10m\Omega.

\[
P_{\text{on}} = 2 \cdot I_{\text{Sat(off)}} \cdot R_{\text{on}} + 2 \cdot \left(\frac{1}{2} \cdot \frac{V_o}{1 - D} \cdot I_{\text{Sat(off)}} \cdot t_{\text{s}} \cdot f_s\right) = 35.604W
\]

The average current \((P_o/V_o)\) in four diodes \((D_{\text{a1}}, D_{\text{a2}}, D_{\text{a3}}, D_o)\) is equal to the output current in the two-phase structure. The conduction voltage \(V_f\) is 0.6V. The conduction losses of all diodes is calculated by

\[
P_o = 4 \cdot \frac{P_o}{V_o} \cdot V_f = 6.364W
\]

The dc resistance \(R_{\text{f}}\) of the inductor is 0.7m\Omega. The loss of the filter inductor is presented in

\[
P_{\text{L}} = 2 \left(\frac{I_{\text{on}}}{2}\right)^2 \cdot R_{\text{f}} = 0.903W
\]

The final value of the efficiency calculation is as follows.

\[
\eta = \left(1 - \frac{P_o}{P_o + P_{\text{Sat}}}\right) \cdot 100\% = 92.1\%
\]
4. Experimental results

A prototype is also designed to verify the feasibility of the system as shown in Fig.8. The test system employs UCC27523 as the integrated driving circuit of the DC-DC converter. The PID compensator and DPWM are realized through DSP.

Fig. 8 Prototype of proposed converter

Fig. 9(a) shows the case with resonant multiplicator. when the drain-to-source voltage of the tube $S_a$ and $S_b$ returns to zero, the current $I_{Sa}$ and $I_{Sb}$ flowing through the MOS tube is almost zero. The turn-on loss of the two switches is zero.

Fig. 9(b) shows the current waveforms of the multiphase and resonant inductors. The phase difference of the current $I_{La}$ and $I_{Lb}$ is 180 degrees. When the current of the resonant inductor linearly increases to the maximum, the value of the current $I_{Lb}$ is nearly equal to $I_{La}$. According to Kirchhoff's law, the current flowing through the tube $S_b$ is zero. It realizes ZCS turn-on.

Fig.10 shows the measured efficiency of the proposed converter. The peak efficiency is 95.4% at the power of 250W. The efficiency is 91.6% at the maximum power.

Table III. Performance comparison between the circuit in this work and other reported high-gain power supplies with the same type

| structure               | This work [31] | [32] | [33] |
|-------------------------|----------------|------|------|
| $V_{in}$/V              | 12             | 20   | 35   |
| $V_{out}$/V             | 220            | 300  | 380  | 400 |
| $f$/kHz                 | 300            | 50   | 500  | 50  |
| gain                    | 18.333         | 15   | 10.85| 13.333 |
| $P_{out}$/W             | 100–500        | 300  | 280  | 100–1000 |
| efficiency              | peak; 95.4% at 250W; 91.6% at max power | 93.07% | 91.8% |
|                         | 95% at 300W; 93.6% at max power | 95%   | 93.6% |

5. Conclusion

A multiplicator ZCS turn on boost converter is presented in this paper, which is capable to provide a large voltage gain to boost up a source voltage of 12V to a voltage of 220V at the output side. The presented converter contains an interleaved boost stage and a resonant multiplier circuit for the voltage boosting purpose. The experimental evaluation is conducted to confirm the performance feasibility of the converter. The measured peak efficiency equals to 95.4% with the gain of 18.333. This work can be well adapted for low input to high output voltage and high power applications such as photovoltaic and fuel cell system.
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