Topology Design and Sliding Mode Controller Design for a Novel High Step-Up DC-DC Converter

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Abstract. This paper discusses a high voltage gain boost converter with enhanced conversion efficiency. A passive clamped circuit is applied to the switch of quadratic boost converter. A step-up capacitor and resonant inductor is adopted to connected in series with the clamp capacitor through diodes. Those constitute a voltage multiplier cell that can absorb stray energy to clamp capacity, reduce the stresses of switch and diodes, and promote voltage gain. Furthermore, a sliding mode control (SMC) scheme with constant frequency is addressed for this converter. It is shown that the tracking error can converge to the equilibrium point against input voltage disturbance and load variation. Numerical simulations are given to verified the theoretical considerations.

1. Introduction

Recently, many industrial applications, for example, renewable energy sources, uninterruptible power supply(UPS), electric vehicles [1-3] develop rapidly. Those are increasing the demand of high performance dc-dc converters. In these applications, the static voltage gain and efficiency of the converter need to be high respectively. However, basic boost converter is hard to provide such a high static voltage gain even for an extreme duty cycle which probably leads to high power losses, and a severe reverse-recovery problem [4].

Many researches have devoted to increase static voltage gain and conversion efficiency. Such as, switched inductor, switched capacitor, coupled-inductor, cascaded structure, etc [5-9]. Those can present higher voltage gain than the conventional boost converter. But more switched-inductor or switched capacitor stages are adopted to obtain a large voltage gain, resulting additional cost and complex circuit. The quadratic boost converter using a single active switch is another topology. That can provide a higher voltage gain than the previously mentioned converters and less components are needed. However, the overlarge voltage stress still exists which reduces efficiency and causes additional cost. In some isolated converters and the converters with coupled-inductor [10], the ultra voltage gain is easily obtained with a proper duty cycle, but the leakage inductor of the coupled-inductors and transformers may cause high voltage spikes and induce energy losses. Moreover, the power transformer and coupled inductor’s volume and weight are obstacles for the development of a compact converter. Clamp techniques are adopted to reduce voltage stress of power devices in high voltage level applications [11]. A voltage multiplier cell is also alternative choice [12]. This cell can not only improve voltage gain but also alleviate voltage stress.

Due to the variational input voltage of dc-dc converters, a control strategy is necessary to stabilize it against load variations, input voltage disturbances and parameter uncertainties. Many classical linear methods can’t guarantee the stabilization or achieve good performance due to the strong nonlinear property of DC-DC converters. Thus, many nonlinear control algorithms have been implemented in DC-DC converters, for example, neural network control, adaptive control, sliding mode control(SMC) [13-15]. SMC is inherently suitable to DC-DC converters because the two switching states of DC-DC converters are similar to switch operation of SMC. Moreover, SMC is characterized by strong robustness, good stability and fast transient response and ease of implementation. Thus, SMC has been widely applied to DC-DC converters. In earlier research, a boundary layer was adopted to alleviate
chattering with a proper frequency [16]. However, the switching frequency was still variational. This can cause excessive power loss, EMI problem and filter design complication. To overcome this drawback, a fixed frequency SMC with linear sliding mode was proposed [17-19]. This controller can ensure a constant operation frequency against external disturbance.

This paper aims to improve the output voltage gain and efficiency of boost converter. First, A passive clamped circuit is applied to the switch of quadratic boost converter in [8] to release voltage stress. Second, a step-up capacitor and resonant inductor is connected in series with the clamp capacitor through diodes to improve voltage gain and promote efficiency. Moreover, a fixed frequency SMC scheme with reaching law is designed for this converter.

2. Operational principles of the proposed converter

The simplified equivalent circuit of the proposed converter can be shown by Figure 1. In this figure, Q is switch; L₁, L₂, L₃ are input inductors and snubber inductor respectively; C₁, Cₒ, C₇, C₉ are capacitors respectively; D₁, D₂, D₃, D₄, D₅ are diodes respectively; R is load. To simplify steady state analysis, many assumptions are as follows:

- All components are ideal.
- L₁, L₂ are large enough so that the converter works under current continues mode (CCM).
- C₁, Cₒ are also sufficiently large such that the voltage across it is considered to be a constant.

Based on the mentioned assumptions, the relative operating modes are presented by Figure 2-3, and the key theoretical waveforms in one period are shown by Figure 4. This converter mainly includes five operation modes. It restarts the whole cycle and repeats the operation in mode 1 after mode 5 ended.

**Figure 1.** The equivalent circuit of the proposed converter

**Figure 2.** Mode 1

**Figure 3.** Mode 2

**Figure 4.** Key theoretical waveforms
1) Mode 1 ([t0-t1], Figure 2): at the instant t0, the switch is turned off, D1, D4 are reverse biased, D2, D3, Do are conducted. The energy stored in the input inductor L1, L2, is transferred to load R and C0 through Do, and to Cc through D2. Meantime, the inductor L1 releases energy to C1 through D2. \(i_{i1}, i_{i2}\) decrease linearly. The relative functions are given as follows:

\[
\frac{di_{i1}}{dt} = \frac{V_{in} - V_{c1}}{L1}, \quad \frac{di_{i2}}{dt} = \frac{V_{c1} - V_{c2}}{L2}
\]

(1)

\[
\frac{dV_{c1}}{dt} = \frac{i_{i1} - i_{i2}}{C1}, \quad \frac{dV_{c2}}{dt} = \frac{i_{i2} - V_{o} / R}{C0}
\]

(2)

where \(i_{i1}, i_{i2}\) denote the current flowing through \(L1, L2\); \(V_{in}, V_{c1}, V_{c2}\) denote averaged input voltage, output voltage and the voltage across \(C1, C2\), respectively.

2) Mode 2([t1-t2], Figure 3): at the instant t1, the switch is turned on. D3, Do, D4 are reverse biased, D1, D2 are conducted. The resonant circuit stops working. The DC source and C1 transfer energy to L1 and L2 respectively so that \(i_{i1}, i_{i2}\) increase linearly until turned off the switch at instant t2. The relative functions are given as follows:

\[
\frac{di_{i1}}{dt} = \frac{V_{in}}{L1}, \quad \frac{di_{i2}}{dt} = \frac{V_{c1}}{L2}
\]

(3)

\[
\frac{dV_{c1}}{dt} = \frac{-i_{i1}}{C1}, \quad \frac{dV_{c2}}{dt} = \frac{-V_{o} / R}{C0}
\]

(4)

Clearly, Cc can be considered as an output of the quadratic boost converter. Cc is charged by Cc in resonant way in a much shorter period than the switch cycle and the current flowing into Cc is also neglected. Thus, \(V_{c1} = V_{c2} = V_{o} / 2\) can be obtained. The averaged state-space model of proposed converter operating in CCM can be written as (5):

\[
\frac{di_{i1}}{dt} = \frac{V_{in} - V_{c1} (1-u)}{L1}, \quad \frac{di_{i2}}{dt} = \frac{V_{c1} - V_{o} (1-u) / 2}{L2}
\]

\[
\frac{dV_{c1}}{dt} = \frac{i_{i1} (1-u) - i_{i2}}{C1}, \quad \frac{dV_{c2}}{dt} = \frac{i_{i2} (1-u) - V_{o} / R}{C2}
\]

(5)

where \(u\) is the control input which takes ‘0’ for the turned off state and ‘1’ for the turned on state of the switch.

According to steady state of this converter, the DC output voltage gain can be calculated by

\[
V_{o} = \frac{2}{(1-D)^2} V_{in}
\]

(6)

From (6), it can be seen that the voltage gain of the proposed converter is much higher than that of basic boost converter. Figure 5 shows a voltage gain comparison of the proposed converter with converter in [9] and [10].

The maximum voltage stresses across power devices of the proposed converter and other converters are given in Table 1. This table shows that the voltage stresses of all power devices are relatively lower than others at a certain voltage. Specially, the voltage stresses of the output diode and switch are half of the converter in [9] and [10]. The above analysis shows that the power loss of output diode and switch is reduced and many low-cost components can be applied.
3. Sliding mode controller design

In this section, a sliding mode control scheme with fixed frequency is addressed. The output voltage $V_o$ is selected to be the desired control variable. The tracking error can be expressed by:

$$e_i = V_r - \beta V_o$$  \hspace{1cm} (7)

where $e_i$ is the voltage error; $V_r$ is the reference voltage; $\beta$ is the voltage feedback network ratio.

The state-space descriptions of the converter operating under CCM is given as follows:

$$
\begin{bmatrix}
\dot{e}_1 \\
\dot{e}_2 \\
\dot{e}_3
\end{bmatrix} =
\begin{bmatrix}
0 & 1 & 0 \\
0 & -\frac{1}{RC_2} & 0 \\
1 & 0 & 0
\end{bmatrix}
\begin{bmatrix}
e_1 \\
e_2 \\
e_3
\end{bmatrix} +
\begin{bmatrix}
0 \\
\beta (V_{C1} - V_o / 2) \\
L/C_2
\end{bmatrix} u$$  \hspace{1cm} (8)

where $e_1$, $e_2$, $e_3$ are the voltage error dynamics, and the integral of voltage error, respectively; $V_{C1}$ is the voltage across $C_1$; $u$ is the inverse logic of $u$.

A general SMC law adopts a switching function, as follows:

$$u = \begin{cases} 
1, & S > 0 \\
0, & S < 0 
\end{cases}$$  \hspace{1cm} (9)

where $S$ is the sliding mode and can be chosen as follows:

$$S = \alpha_1 e + \alpha_2 \dot{e} + \alpha_3 \ddot{e}$$  \hspace{1cm} (10)

$\alpha_1$, $\alpha_2$, $\alpha_3$ represent the sliding coefficients.

To ensure that the state trajectory is maintained within the vicinity of the sliding line, the existence condition derived from Lyapunov’s second method must be obeyed:

$$\lim_{S \to 0} S \dot{S} < 0 = \begin{cases} 
S \to 0^+, & \dot{S} < 0 \\
S \to 0^-, & \dot{S} > 0 
\end{cases}$$  \hspace{1cm} (11)

where $\dot{S}$ is the time derivative of $S$ and is calculated as follows:

$$\dot{S} = -\alpha_1 \beta \frac{\dot{i}_C}{C_o} + \alpha_2 \beta - \frac{\dot{i}_C}{RC_o} + \alpha_3 e_i$$  \hspace{1cm} (12)

The existence condition is given by:

---

Table 1. Power devices voltage stress comparison

| Power devices | The proposed converter in [9] | The converter in [10] |
|---------------|-----------------|-----------------|
| Q             | $\frac{2V_o}{2}$ | $\frac{2V_o}{2+nk+nDk}$ |
| D$_o$         | $\frac{V_o}{2}$  | $\frac{DV_o}{2(1-D)}$   |
| D$_1$         | $V_{in} / (1-D)$ | $D_{in} / (1-D)$ |
| D$_2$         | $V_{in} / (1-D)$ | $D_{in} / (1-D)$ |
| D$_3$         | $V_o / 2$       | $V_o / 2$ |
| D$_4$         | $V_o / 2$       | $V_o / 2$ |

---

Figure 5. Voltage gain comparison of converters.
\[ 0 < \beta L_2 \left( \frac{\alpha_1}{\alpha_2} - \frac{1}{RC_o} \right) I_{C_o} - L_o C_o \frac{\alpha_3}{\alpha_2} e_1 < \beta \left( \frac{V_o}{2} - V_{C_i} \right) \]  

(13)

the \( \alpha_1, \alpha_2, \alpha_3 \) must be satisfied by (13) with considering the steady value of state variables. Hence employed the control input \( u \), tracking error can converge to zero. However, SM controlled converters generally suffer from significant switching frequency variation when the input voltage or output load is varied [17]. To solve the drawback, an equivalent SMC with constant switching frequency is proposed.

Equating

\[ \dot{S} = 0 \]  

(14)

yields the equivalent control input

\[ u_{eq} = -\frac{L_2}{(V_o/2 - V_{C_i})} \left( \frac{\alpha_1}{\alpha_2} - \frac{1}{RC_o} \right) I_{C_o} - \frac{\alpha_3}{\alpha_2} L_o C_o e_1 + \frac{1}{\beta \alpha_2 (V_o/2 - V_{C_i})} \]  

(15)

where \( u_{eq} \) is a smooth function of the discrete control input \( u \) and \( 0 < u_{eq} < 1 \).

An exponential reaching law is adopted to improve transient response and can be expressed:

\[ \dot{S} = -k_1 S - k_2 \text{sign}(S) \]  

(16)

where \( k_1, k_2 \) are positive parameters.

When (16) is solved for \( u \), the total smooth control input:

\[ u = -\frac{L_2}{(V_o/2 - V_{C_i})} \left( \frac{\alpha_1}{\alpha_2} - \frac{1}{RC_o} \right) I_{C_o} - \frac{L_o C_o (\alpha_3 e_1 + k_1 S + k_2 \text{sign}(S))}{\beta \alpha_2 (V_o/2 - V_{C_i})} + \frac{1}{(V_o/2 - V_{C_i})} \]  

(17)

Finally, the total smooth control input \( u \) and ramp signal \( V_{\text{ramp}} = 1 \) with a certain frequency are fed into a pulsewidth modulator to generate the discrete control input \( u \).

4. Simulation results and discussions

In this section, the simulation of the PWM-based SM controller is presented. Table 2 shows the specification of the proposed converter.

**Table 2.** main parameter of proposed converter

| \( L_1 \) | \( L_2 \) | \( L_r \) | \( C_1 \) | \( C_c \) | \( C_r \) | \( C_o \) | \( R \) | \( f_s \) |
|---|---|---|---|---|---|---|---|---|
| 100 \( \mu \)H | 150 \( \mu \)H | 40 \( \mu \)F | 250 \( \mu \)F | 15 \( \mu \)F | 40 \( \mu \)F | 470 \( \mu \)F | 500 \( \Omega \) | 40 kHz |

**Figure 6.** The transient response of the system.

**Figure 7.** The output voltage waveform versus load variation

**Figure 8.** The output voltage waveform versus input variation

Figure 6 illustrates the transient responses comparison of the different SM controllers for the proposed converter. In the figures, ‘a’ represents output trajectory with the proposed SM controller in (17), ‘b’ denotes the SM controller in (15), ‘c’ is the SM controller formulated by the sliding mode in
(10) without integral term respectively. Taking 120 V as the reference, the system transient responses for the proposed SM controller has a settling time of 1 ms and no overshoot, which is relatively better than other responses with 1.5 ms and 3 ms settling time respectively. Figure 7 presents the output response versus load disturbance. A sudden load change is introduced at 0.02 s, and 0.025 s. three responses possess similar settling time around 2 ms, but the smallest output voltage variation belongs to the proposed SM controller around 5 V. Similarly, figure 8 shows the output response versus input variation, which decreases from 24 V to 14 V, then return to 24 V. Three output voltages maintain at 120 V around 0.5 V variation and the system for the SM controller in (15) has the longest settling time about 2 ms. It can be seen that the proposed controller is relatively superior and robust compared with ‘b’ controller and ‘c’ controller for the new converter.

5. Conclusion
This paper has developed a high step-up boost converter with a voltage multiplier. The newly proposed converter has the following features. (1) the output voltage can highly heightened because of the voltage multiplier and serial structure of boost converter. (2) the stray energy also can be recycled into the clamped capacity ($C_c$) to improve the efficiency. (3) the voltage stresses of the MOSFET and diodes ($D_1$, $D_3$, $D_4$) are decreased to half of the output voltage so that more economical components can be applied to it. Moreover, a sliding mode control scheme with fixed frequency is designed to stabilized the converter and track reference voltage. The simulation results show that the controlled system can track the reference voltage with short settling time and no overshoot. The highly robustness of the proposed controller against input voltage and load disturbances is also verified.

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