A Novel Quadratic Boost Converter With Low Inductor Currents

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Abstract—In this paper, a novel quadratic boost converter is presented. The input current of the converter is non-pulsating and the currents of the inductors are relatively low. The operating principle and the mathematical model of the converter in continuous conduction mode (CCM) are given. The steady-state of the converter is calculated and the power loss of the converter is analyzed. Then the small-signal model of the converter is derived based on the state space averaging method. The inductor current and the voltage stress of the proposed converter are low, which can improve the efficiency of the converter. The simulations and the circuit experiments are presented to verify the good performance of the converter.

Index Terms—Inductor current, quadratic boost converter, steady state.

I. INTRODUCTION

RECENT years, more attention has been paid to the renewable energy for the increasing cost of fossil fuel and the regulations of CO₂ emissions [1]. The output voltage of these energy sources, such as solar panel and fuel cells, is low and variable. Thus, it needs high-ratio DC-DC converter to get the appropriate output voltage [2]. In order to obtain a high output voltage, the conventional boost converter must operate at extremely high duty-cycle ratios, that is quite difficult to be obtained in practical application for the limitation of semiconductors. Small off times cause high switching voltage stress with reverse recovery issues and the low efficiency [3].

A number of high step-up converter topologies have been proposed to get the high voltage gain with suitable duty cycle. Converters with coupled inductors can achieve a high step up voltage gain [4], but their efficiency is low for the losses of leakage inductors. The active-clamp circuit has been applied and can recycle the leakage energy [5], [6]. Also, the switched capacitor is a well-known voltage boosting technique based a charge pump circuit used in many converters [2], [7]. Another method for increasing voltage gain of a DC-DC converter is to employ several stages of converter modules in some ways, such as the quadratic boost converter [8].

The voltage gain of the quadratic converter is the quadratic function of its duty ratio [9]. A quadratic cascade boost converter is proposed by F. L. Luo in [10]. A multistage cascade quadratic converter with several voltage-boosting modules and only one active switch is presented in [11]. In [12], a quadratic boost converter with low capacitor voltage stress is proposed. However, the input current of this converter is pulsating, thus it is necessary and difficult to design input filter. Another quadratic boost converter with two active switches is proposed by [13], where the active switches are controlled by the same PWM signal. The input current of the converter in [13] is continuous, and the voltage stress on power switch is low. In [14], a novel quadratic boost converter with low inductor currents is proposed. Its voltage gain is the same as the conventional quadratic boost converter. The input current of the proposed converter is non-pulsating, the voltage stress on power switch is low, and the inductors' current of the proposed converter is low, that can decrease the power losses and increase the efficiency of the converter.

In this paper, the novel quadratic boost converter in [14] is analyzed in detail, and the circuit experiments are presented to verify the good performance of the proposed converter. This paper is organized as follows. In Section II, the topology and operating principle in CCM are given. The steady state analysis and the loss analysis are presented in Sections III and V, respectively. The comparisons among the proposed converter and the other quadratic boost converters in [10], [12] and [13] are given in Sections IV. The small-signal model is described in Section VI. In Sections VII and VIII, the simulations and the circuit experiments are presented for verification. Some conclusions and future work are concluded in Section IX.

II. OPERATING PRINCIPLE OF THE PROPOSED CONVERTER

The proposed converter is shown in Fig. 1(a). This converter consists of two power switches S_1 and S_2 , two diodes D_1 and D_2 , two inductors L_1 and L_2 , two capacitors C_1 and C_2 , and the load R_L . Two power switches are controlled synchronously and driven by the same PWM signal with the period being T and the duty cycle being D .

The proposed converter working in CCM contains two modes. In the first mode, $t \in (NT, NT + DT)$ the power switch S_1 and S_2 are turned on and the diodes D_1 and D_2 are not conducted for the inverse biased voltage, as shown in Fig. 1(b). The second mode $t \in (NT + DT, NT + T)$ is shown in Fig. 1(c). The power switches S_1 and S_2 are turned off in this mode. The diodes are conducted for its forward biased voltage. The related equations of the two modes can be written as follows:

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Fig. 1. (a) Proposed quadratic boost converter. (b) Mode 1. (c) Mode 2.

$$\begin{cases}
 L_1 \frac{di_{L1}}{dt} = v_{in} \\
 L_2 \frac{di_{L2}}{dt} = v_{C1} + v_{in} \\
 C_1 \frac{dv_{C1}}{dt} = -i_{L2} \\
 C_2 \frac{dv_o}{dt} = -\frac{v_o}{R_L}
 \end{cases} \quad t \in (NT, NT + DT) \quad (1)$$

$$\begin{cases}
 L_1 \frac{di_{L1}}{dt} = v_{in} - v_{C1} \\
 L_2 \frac{di_{L2}}{dt} = v_{in} - v_o \\
 C_1 \frac{dv_{C1}}{dt} = i_{L1} \\
 C_2 \frac{dv_o}{dt} = i_{L2} - \frac{v_o}{R_L}
 \end{cases} \quad t \in (NT + DT, NT + T) \quad (2)$$

III. STEADY STATE ANALYSIS

For steady-state theoretical analysis, it is assumed that all components are ideal and the proposed converter operates in CCM. It is assumed that V_{in} , V_{C1} , I_{L1} , I_{L2} , I_o and V_o are corresponding DC values of v_{in} , v_{C1} , i_{L1} , i_{L2} , i_o and v_o .

TABLE I
DC VALUES OF THE PROPOSED CONVERTER

	Voltage	Current
Capacitors	$V_{C1} = \frac{1}{1-D} V_{in}$	
	$V_o = \frac{1}{(1-D)^2} V_{in}$	
Inductors		$I_{L1} = \frac{D}{(1-D)^2} I_o$
		$I_{L2} = \frac{1}{1-D} I_o$

TABLE II
VOLTAGE AND CURRENT STRESS OF SWITCHES AND DIODES

	Voltage	Current
Switches	$V_{S1} = \frac{V_{in}}{1-D}$	$I_{S1} = \frac{D}{(1-D)^2} I_o$
	$V_{S2} = \frac{V_{in}}{(1-D)^2}$	$I_{S2} = \frac{D}{1-D} I_o$
Diodes	$V_{D1} = \frac{V_{in}}{1-D}$	$I_{D1} = \frac{D}{1-D} I_o$
	$V_{D2} = \frac{(2-D)}{(1-D)^2} V_{in}$	$I_{D2} = I_o$

The DC values of the converter can be deduced as Table I by applying voltage-second balance and charge balance principle. Thus, the voltage conversion ratio of the proposed converter can be derived as (3).

$$M = \frac{V_o}{V_{in}} = \frac{1}{(1-D)^2} \quad (3)$$

From Fig. 1 and Table I, the voltage stress and current stress of the switches and diodes can be derived as Table II. These parameters are useful for selecting the specific model of the devices in subsequent experiments.

Some of the waveforms of the proposed converter in CCM mode are shown in Fig. 2.

From the time-domain waveforms shown in Fig. 2, the peak-to-peak ripple can be deduced as follows:

$$\begin{aligned}
 \Delta i_{L1} &= \frac{DTV_{in}}{L_1}; \quad \Delta i_{L2} = \frac{D(2-D)TV_{in}}{(1-D)L_2}; \\
 \Delta v_{C1} &= \frac{DTV_o}{(1-D)R_L C_1}; \quad \Delta v_o = \frac{DTV_o}{R_L C_2}.
 \end{aligned} \quad (4)$$

Current ripple ratios ($\rho_{L1} = \Delta i_{L1}/I_{L1}$ and $\rho_{L2} = \Delta i_{L2}/I_{L2}$) and voltage ripple ratios ($\rho_{C1} = \Delta v_{C1}/V_{C1}$ and $\rho_o = \Delta v_o/V_o$) which are useful for selecting the values of inductors and capacitors are calculated.

$$\begin{aligned}
 \rho_{L1} &= \frac{R_L T (1-D)^4}{L_1}; \quad \rho_{L2} = \frac{(1-D)^2 (2-D) D R_L T}{L_2}; \\
 \rho_{C1} &= \frac{DT}{(1-D)^2 R_L C_1}; \quad \rho_o = \frac{\Delta v_o}{V_o} = \frac{DT}{R_L C_2}.
 \end{aligned} \quad (5)$$

Fig. 2. Some typical waveforms of the proposed converter.

Fig. 3. Curves of boundary inductor time constants τ_{b1} and τ_{b2} .

The boundary between continuous conduction mode and discontinuous conduction mode is that the inductor current decrease to zero, that is $I_{L1} = \Delta i_{L1}/2$ and $I_{L2} = \Delta i_{L2}/2$. The condition for the proposed converter working in continuous conduction mode can be derived as follows:

$$\begin{aligned} \tau_{L1} &= \frac{2L_1}{R_L T} \geq \tau_{b1} = (1-D)^4; \\ \tau_{L2} &= \frac{2L_2}{R_L T} \geq \tau_{b2} = D(1-D)^2(2-D). \end{aligned} \quad (6)$$

τ_{b1} and τ_{b2} are normalized inductor time constants. The curves for τ_{b1} and τ_{b2} are shown in Fig. 3. From the proposed converter in the operating point, if $\tau_{L1} \geq \tau_{b1}$ and $\tau_{L2} \geq \tau_{b2}$, the converter will work in CCM. Otherwise, the converter will operate in DCM.

IV. COMPARISONS OF QUADRATIC BOOST CONVERTERS

There are some quadratic boost converters in the literatures. The quadratic cascade boost converter is proposed by F. L. Luo

Fig. 4. Some existing quadratic boost converter. (a) Converter in [10]. (b) Converter in [12]. (c) Converter in [13].

in [10]. Then, a quadratic boost converter with low capacitor voltage stress is proposed by Y. M. Ye in [12]. F. Wang proposed another quadratic boost converter with two active switches in [13]. The respective circuit diagrams are shown in Fig. 4(a), (b) and (c).

The inductor currents, the inductor ripple currents, the currents flowing through the switches and the diodes, the output voltage, the voltage stress and the input current of the proposed converter (PCV) are compared with those terms in [10], [12] and [13], which are shown in Table III.

From Table III, we can conclude that under the same conditions, the average current of the first inductor of the proposed converter (I_{L1}) is lower than that in [10], [12], and the second average inductor current of the PCV (I_{L2}) is lower than that in [13], which can reduce the copper loss of the inductors and increase the efficiency of the converter. Moreover, the ripple current of the first inductor of the PCV (Δi_{L1}) is also lower than [13]. The average currents flowing through the switches of the PCV are same as that in [13], but average current flowing through S_1 is smaller than that in [10], [12]. The voltage stress of the proposed converter is same as that in [13], but voltage stress on S_1 is smaller than that in [10], [12]. The input current of PCV is non-pulsating since its value is equal to i_{L1} plus i_{L2} , that will make the design of the input filter much easier.

TABLE III
COMPARISONS AMONG DIFFERENT CONVERTERS

	C[10]	C[12]	C[13]	PCV
I_{L1}	$\frac{I_o}{(1-D)^2}$	$\frac{I_o}{(1-D)^2}$	$\frac{DI_o}{(1-D)^2}$	$\frac{DI_o}{(1-D)^2}$
I_{L2}	$\frac{I_o}{1-D}$	$\frac{I_o}{1-D}$	$\frac{I_o}{(1-D)^2}$	$\frac{I_o}{1-D}$
I_L				
ΔI_{L1}	$\frac{DTV_{in}}{L_1}$	$\frac{DTV_{in}}{L_1}$	$\frac{DTV_{in}}{(1-D)L_1}$	$\frac{DTV_{in}}{L_1}$
ΔI_{L2}	$\frac{DTV_{in}}{(1-D)L_2}$	$\frac{DTV_{in}}{(1-D)L_2}$	$\frac{D(2-D)TV_{in}}{(1-D)L_2}$	$\frac{D(2-D)TV_{in}}{(1-D)L_2}$
I_S				
S_1	$\frac{(2-D)DI_o}{(1-D)^2}$	$\frac{(2-D)DI_o}{(1-D)^2}$	$\frac{DI_o}{(1-D)^2}$	$\frac{DI_o}{(1-D)^2}$
S_2			$\frac{DI_o}{(1-D)}$	$\frac{DI_o}{1-D}$
D_1	I_o	I_o	$\frac{D}{1-D} I_o$	$\frac{D}{1-D} I_o$
I_D				
D_2	$\frac{DI_o}{(1-D)^2}$	$\frac{I_o}{1-D}$	I_o	I_o
D_3	$\frac{I_o}{1-D}$	$\frac{DI_o}{(1-D)^2}$		
V_S				
S_1	$\frac{V_{in}}{(1-D)^2}$	$\frac{V_{in}}{(1-D)^2}$	$\frac{V_{in}}{1-D}$	$\frac{V_{in}}{1-D}$
S_2			$\frac{V_{in}}{(1-D)^2}$	$\frac{V_{in}}{(1-D)^2}$
V_D				
D_1	$\frac{V_{in}}{(1-D)^2}$	$\frac{V_{in}}{(1-D)^2}$	$\frac{V_{in}}{1-D}$	$\frac{V_{in}}{1-D}$
D_2	$\frac{DV_{in}}{(1-D)^2}$	$\frac{V_{in}}{1-D}$	$\frac{(2-D)V_{in}}{(1-D)^2}$	$\frac{(2-D)V_{in}}{(1-D)^2}$
D_3	$\frac{V_{in}}{1-D}$	$\frac{DV_{in}}{(1-D)^2}$		
V_o	$\frac{V_{in}}{(1-D)^2}$	$\frac{V_{in}}{(1-D)^2}$	$\frac{V_{in}}{(1-D)^2}$	$\frac{V_{in}}{(1-D)^2}$
Input current	Non-pulsating	Pulsating	Non-pulsating	Non-pulsating

V. ANALYSIS OF THE POWER LOSS

Power loss of the proposed converter is analyzed in this part. The total power losses of the converter include the losses of power switches, diodes, inductors and capacitors. The parasitic resistances are defined as follows: r_{ds1} and r_{ds2} are the switch on-state resistances; r_{F1} and r_{F2} are the forward resistances of diodes D_1 and D_2 ; r_{L1} and r_{L2} are the equivalent series resistance (ESR) of inductors L_1 and L_2 ; r_{C1} and r_{C2} are the ESR of capacitors C_1 and C_2 ; forward voltages V_{F1} and V_{F2} are the forward bias voltages of diodes D_1 and D_2 . The voltage ripple across the capacitors and the inductors is ignored.

A. Power Losses of Power Switches

MOSFET is chosen as the power switch of the proposed converter. The losses of the power switches include the conduction loss and the switching loss. The conduction power losses of the switches can be calculated as follows:

$$P_{SWC} = r_{ds1} I_{S1,rms}^2 + r_{ds2} I_{S2,rms}^2 = r_{ds1} D(I_{L1} + I_{L2})^2 + r_{ds2} DI_{L2}^2 \quad (7)$$

$$= \frac{r_{ds1} D}{(1-D)^4} \frac{V_o^2}{R_L^2} + \frac{r_{ds2} D}{(1-D)^2} \frac{V_o^2}{R_L^2}$$

The switching losses are related to the rise time t_r and the fall time t_f of the switch ($t_{off} = t_r + t_f$), the voltage across the power switch V_s and the averaged current I_s across it.

$$P_{SWoff} = \frac{1}{2} I_{S1} V_{S1} t_{off1} f_s + \frac{1}{2} I_{S2} V_{S2} t_{off2} f_s \quad (8)$$

$$= \frac{1}{2} \frac{DV_o^2}{(1-D)R_L} t_{off1} f_s + \frac{1}{2} \frac{DV_o^2}{(1-D)R_L} t_{off2} f_s$$

where f_s is the switching frequency, $f_s = 1/T$; t_{off1} and t_{off2} are the turn off across time of the switches S_1 and S_2 , respectively.

The total losses of the switch S ($P_{Switch,tot}$) can be earned as follows:

$$P_{Switch,tot} = P_{SWC} + P_{SWoff} \quad (9)$$

B. Power Losses of Diodes

The power loss of diode is mainly caused by the forward voltage drop V_{F1} and V_{F2} , and the series resistance r_{F1} and r_{F2} .

The diodes D_1 and D_2 forward resistance losses can be get as follows:

$$P_{RFD} = r_{F1} I_{D1,rms}^2 + r_{F2} I_{D2,rms}^2 = \frac{r_{F1} D^2 V_o^2}{(1-D)^3 R_L^2} + \frac{r_{F2} V_o^2}{(1-D) R_L^2} \quad (10)$$

The diodes D_1 and D_2 forward voltage losses can be achieved as follows:

$$P_{VFD} = V_{F1} I_{D1} + V_{F2} I_{D2} = V_{F1} \frac{DV_o}{(1-D)R_L} + V_{F2} \frac{V_o}{R_L} \quad (11)$$

The total losses of the diodes D_1 and D_2 (P_D) can be deduced as follows:

$$P_D = P_{RFD} + P_{VFD} \quad (12)$$

C. Power Losses of Inductors

The power loss of inductor has copper loss caused by winding resistance (r_{L1} and r_{L2}) and core loss caused by hysteresis and eddy current in the magnetic core.

The copper loss places an important role in the total power loss of inductor. The conduction losses of the inductors L_1 and L_2 can be get as follows:

$$P_{Lr} = r_{L1} I_{L1,rms}^2 + r_{L2} I_{L2,rms}^2 = \frac{r_{L1} D^2 V_o^2}{(1-D)^4 R_L^2} + \frac{r_{L2} V_o^2}{(1-D)^2 R_L^2} \quad (13)$$

Obviously, decreasing the RMS of the currents flowing through L_1 and L_2 can reduce the copper losses of the inductors.

The iron losses P_{Lfe} of the inductors L_1 and L_2 mainly consist of two parts: the magnetic hysteresis loss (P_{Lh1} , P_{Lh2}) and the eddy current loss (P_{Lc1} , P_{Lc2}).

$$\begin{aligned} P_{Lfe} &= P_{Lh1} + P_{Lc1} + P_{Lh2} + P_{Lc2} \\ &= K_{h1} f B_{m1}^2 + K_{c1} (f B_{m1})^2 + K_{h2} f B_{m2}^2 + K_{c2} (f B_{m2})^2 \end{aligned} \quad (14)$$

where K_{h1} and K_{h2} are the loss coefficients related to the hysteresis loss, K_{c1} and K_{c2} are the loss coefficients related to the eddy current loss, f is the frequency of the voltages across the inductors, B_{m1} and B_{m2} are the magnetic flux density, which is related to the current flowing through the inductors L_1 and L_2 , respectively.

$$B_{m1} = \mu H_1 = \mu k_1 I_{L1}, \quad B_{m2} = \mu H_2 = \mu k_2 I_{L2},$$

where μ is the magnetic permeability of the iron core, H_1 and H_2 are the magnetic field intensity, k_1 and k_2 are the constant coefficients. Thus, the iron losses P_{Lfe} of the inductors L_1 and L_2 can be deduced as follows:

$$\begin{aligned} P_{Lfe} &= P_{Lh1} + P_{Lc1} + P_{Lh2} + P_{Lc2} \\ &= \frac{(K_{h1} f + K_{c1} f^2) k_1^2 \mu^2 D^2 V_o^2}{(1-D)^4 R_L^2} + \frac{(K_{h2} f + K_{c2} f^2) k_2^2 \mu^2 V_o^2}{(1-D)^2 R_L^2} \end{aligned}$$

Due to the magnetic saturation, μ decreases as the current flowing through the inductor increases. The iron losses P_{Lfe} of the given inductors with certain frequency can be considered as a fixed value when the load R_L varies.

The total power losses of inductors are the sum of core loss and copper loss, that is,

$$P_L = P_{Lfe} + P_{Lr} \quad (15)$$

D. Power Losses of Capacitors

The power losses of capacitors C_1 and C_2 due to the ESR can be obtained as follows:

$$P_C = r_{C1} \hat{I}_{C1,rms}^2 + r_{C2} \hat{I}_{C2,rms}^2 = \frac{r_{C1} D V_o^2}{(1-D)^3 R_L^2} + \frac{r_{C2} D V_o^2}{(1-D) R_L^2} \quad (16)$$

E. Efficiency

The total power dissipation of the proposed converter is the sum of the power losses of power switches, diodes, inductors and capacitors, that is,

$$P_{Loss} = P_{Switch,tot} + P_D + P_L + P_C \quad (17)$$

Thus, the efficiency η of the converter can be obtained as follows:

$$\eta = \frac{P_o}{P_o + P_{Loss}} = \frac{1}{1 + \frac{P_{Loss} R_L}{V_o^2}} \quad (18)$$

VI. SMALL-SIGNAL MODELING

Based on state averaging method, the average model of the proposed converter operating in CCM mode can be described as follows:

$$\begin{cases} L_1 \frac{d\langle i_{L1} \rangle}{dt} = \langle v_{in} \rangle - (1-d)\langle v_{C1} \rangle \\ L_1 \frac{d\langle i_{L2} \rangle}{dt} = \langle v_{in} \rangle + d\langle v_{C1} \rangle - (1-d)\langle v_o \rangle \\ C_1 \frac{d\langle v_{C1} \rangle}{dt} = -d\langle i_{L2} \rangle + (1-d)\langle i_{L1} \rangle \\ C_2 \frac{d\langle v_o \rangle}{dt} = (1-d)\langle i_{L2} \rangle - \frac{\langle v_o \rangle}{R_L} \end{cases} \quad (19)$$

where $\langle i_{L1} \rangle$, $\langle i_{L2} \rangle$, $\langle v_{in} \rangle$, $\langle v_{C1} \rangle$ and $\langle v_o \rangle$ are the average values of i_{L1} , i_{L2} , v_{in} , v_{C1} and v_o respectively. d is the duty cycle.

The small ac perturbations are given as follows:

$$\begin{cases} \langle i_{L1} \rangle = I_{L1} + \hat{i}_{L1} \\ \langle i_{L2} \rangle = I_{L2} + \hat{i}_{L2} \\ \langle v_{C1} \rangle = V_{C1} + \hat{v}_{C1} \\ \langle v_o \rangle = V_o + \hat{v}_o \\ \langle v_{in} \rangle = V_{in} + \hat{v}_{in} \\ \langle d \rangle = D + \hat{d} \end{cases} \quad (20)$$

where, the small ac signals and the DC values satisfy the following conditions, that is,

$$\begin{cases} \hat{i}_{L1} \ll I_{L1} \\ \hat{i}_{L2} \ll I_{L2} \\ \hat{v}_{C1} \ll V_{C1} \\ \hat{v}_o \ll V_o \\ \hat{v}_{in} \ll V_{in} \\ \hat{d} \ll D \end{cases} \quad (21)$$

Substituting (20) and (21) into (19), separating the small-signal ac terms from (19), omitting the higher-order small signal terms, and the small ac equations of the proposed converter can be obtained as follows:

$$\begin{cases}
L_1 \frac{d\hat{i}_{L1}}{dt} = \hat{v}_{in} + V_{C1} \hat{d} - (1-D)\hat{v}_{C1} \\
L_2 \frac{d\hat{i}_{L2}}{dt} = \hat{v}_{in} + (V_{C1} + V_o) \hat{d} + D\hat{v}_{C1} - (1-D)\hat{v}_o \\
C_1 \frac{d\hat{v}_{C1}}{dt} = -(I_{L1} + I_{L2}) \hat{d} + (1-D)\hat{i}_{L1} - D\hat{i}_{L2} \\
C_2 \frac{d\hat{v}_o}{dt} = -I_{L2} \hat{d} + (1-D)\hat{i}_{L2} - \frac{\hat{v}_o}{R_L}
\end{cases} \quad (22)$$

Using Laplace transform, the transfer functions of the output voltage to the duty cycle $G_{vd}(s)$ and the first inductor current to the duty cycle $G_{ild}(s)$ can be deduced as follows:

$$\begin{aligned}
G_{vd}(s) &= \frac{\hat{v}_o(s)}{\hat{d}(s)} \Big|_{\hat{v}_{in}(s)=0} \\
&= \frac{b_3 s^3 + b_2 s^2 + b_1 s + b_0}{(1-D)^3 (a_4 s^4 + a_3 s^3 + a_2 s^2 + a_1 s + a_0)} \quad (23)
\end{aligned}$$

$$\begin{aligned}
G_{ild}(s) &= \frac{\hat{i}_{L1}(s)}{\hat{d}(s)} \Big|_{\hat{v}_{in}(s)=0} \\
&= \frac{m_3 s^3 + m_2 s^2 + m_1 s + m_0}{(1-D)^3 (a_4 s^4 + a_3 s^3 + a_2 s^2 + a_1 s + a_0)} \quad (24)
\end{aligned}$$

where,

$$\begin{cases}
a_4 = R_L L_1 L_2 C_1 C_2 \\
a_3 = L_1 L_2 C_1 \\
a_2 = (1-D)^2 R_L L_1 C_1 + (1-D)^2 R_L L_2 C_2 + D^2 R_L L_1 C_2 \\
a_1 = (1-D)^2 L_2 + D^2 L_1 \\
a_0 = R_L (1-D)^4 \\
b_3 = -V_{in} L_1 L_2 C_1 \\
b_2 = (1-D)^2 (2-D) V_{in} R_L L_1 C_1 \\
b_1 = -V_{in} [(1-D)^2 L_2 + (D+D^2) L_1] \\
b_0 = 2V_{in} (1-D)^4 R_L
\end{cases} ,$$

and

$$\begin{cases}
m_3 = (1-D)^2 V_{in} R_L L_2 C_1 C_2 \\
m_2 = V_{in} [(1-D)^2 L_2 C_1 + L_2 C_2] \\
m_1 = V_{in} \left[\frac{L_2}{R_L} + (1-D)^4 R_L C_1 + 2D(1-D)^2 R_L C_2 \right] \\
m_0 = V_{in} [(3D+1)(1-D)^2]
\end{cases} \quad (25)$$

$G_{ild}(s)$ will affect the current loop transfer function, and $G_{vd}(s)$ will affect the voltage loop transfer function. According to $G_{ild}(s)$ and $G_{vd}(s)$, the controller should be designed to make the open loop transfer function of the entire system

TABLE IV
SPECIFICATIONS OF THE MAIN CIRCUIT

Parameter	Value
Input voltage V_{in}	20 V
Output voltage V_o	70 V
Switching frequency f	50 kHz
Output load R_L	200 Ω
Output power P_o	24.5 W
Current ripple ratio L_1	< 30%
Current ripple ratio L_2	< 30%
Voltage ripple ratio C_1	< 10%
Voltage ripple ratio C_2	< 0.1%

having enough stability margin.

VII. DESIGN PROCEDURE AND SIMULATION RESULTS

The specifications of the main circuit are shown in Table IV. The duty cycle is designed according to the voltage gain of the converter.

$$D = 1 - \frac{1}{\sqrt{M}} = 1 - \sqrt{\frac{V_{in}}{V_o}} = 0.4655. \quad (26)$$

The inductors and capacitors are designed according to the Table IV and ripple ratios analyzed in Section III. The conditions for the inductance value and capacitance value are deduced as follows:

$$L_1 > \frac{(1-D)^4 R_L T}{0.3} \quad (27)$$

$$L_2 > \frac{(2-D)(1-D)^2 D R_L T}{0.3} \quad (28)$$

$$C_1 > \frac{DT}{0.1 R_L (1-D)^2} \quad (29)$$

$$C_2 > \frac{DT}{0.001 R_L}. \quad (30)$$

Therefore, the parameters of the inductors and capacitors are selected as follows: $L_1 = 3$ mH, $L_2 = 3$ mH, $C_1 = 10$ μ F and $C_2 = 100$ μ F.

To verify theoretical analysis, simulation is done by MATLAB/Simulink. Fig. 5 shows the time-domain simulation waveforms of the inductor currents i_{L1} and i_{L2} , output voltage v_o , driving voltage v_{G1} and diode voltage v_{D1} for the proposed converter. The simulation waveforms in time-domain show the converter operates in CCM, and the output voltage is about 70 V, which is equal to the theoretical value.

VIII. EXPERIMENTAL RESULTS

The experiment circuit is built to verify the theoretical and

Fig. 5. Simulated waveforms of the proposed converter.

simulation results. The parameters of the circuit experiment are same as those parameters presented in the simulations of Section VII. IRFP4668 is chosen as the power switch. MBR40250 is selected as the diode.

The average current mode control is applied in the experiment, and the controller is designed according to the transfer functions of $G_{vd}(s)$ and $G_{ild}(s)$ shown in (23) and (24).

The schematic diagram of average current mode control is shown in Fig. 6(a). The diagram of current loop and voltage loop are given in Fig. 6(b) and (c). The bode plots of $G_{vd}(s)$ and $G_{ild}(s)$ are shown in Fig. 7(a) and (b), respectively. Fig. 7(c) shows that when applying the average current mode control, the stability margin of the open loop transfer function

Fig. 6. Schematic diagram of control method and diagram of current loop and voltage loop. (a) Schematic diagram of average current mode control. (b) Diagram of current loop. (c) Diagram of voltage loop.

of entire system is about 90° .

Fig. 8 shows the time-domain experimental waveforms of the inductor currents i_{L1} and i_{L2} , input current i_{in} , output voltage v_o , driving voltage v_{G1} and diode voltage v_{D1} for the proposed converter. From Fig. 8(a), the proposed converter operates in CCM, and the input current is non-pulsating. The average input current of i_{in} is 1.23 A, and the average currents of i_{L1} and i_{L2} are 0.66 A and 0.57 A. According to the average model, the average currents of the inductors should be 0.57 A and 0.655 A, the average input current should be 1.225 A. The experimental results of the average inductor currents coincide with the average model well. From Fig. 8(b), the output voltage is 70 V, which is in accordance with the average model. Due to the leakage inductance of the inductor and the stray inductance in line, some voltage spikes appear in the waveform of the output voltage.

In order to compare the efficiency of the proposed converter with the converters in [10], [12] and [13], the experiments have been done. The circuit parameters of the above converters have been selected as follows: $L_1 = 3 \text{ mH}$, $L_2 = 3 \text{ mH}$, $C_1 = 10 \mu\text{F}$ and $C_2 = 100 \mu\text{F}$. The input voltage is set to be 20 V, the output voltage is 70 V. The experimental results of the converters with different loads, $R_L = 200 \Omega$, 300Ω , 400Ω , are given in the Table V. Fig. 9 shows the experimental efficiency of the above converters under different loads. Compared with the converter in [10], [12] and [13], efficiency of the proposed converter has been improved due to the reduced inductor currents and the

Fig. 7. Bode diagram. (a) Bode diagram of $G_{vd}(s)$. (b) Bode diagram of $G_{ild}(s)$. (c) Bode diagram of open loop transfer function of entire system when applying average current mode control.

lower voltage stress on power switch S_1 .

Fig. 10 is the transient response curves of the output voltage when the load and input voltage are disturbed. In Fig.10(a), switching the load from 500Ω to 150Ω at time t_1 , the output voltage suddenly drops and then returns to stable value. Switching the load from 150Ω to 500Ω at time t_2 , the output voltage suddenly rises and then returns to 70 V. In Fig. 10(b), switching the input voltage from 20 V to 25 V at time t_3 ,

Fig. 8. Experimental waveforms. (a) Current waveforms. (b) Voltage waveforms.

TABLE V
THE EXPERIMENTAL RESULTS OF THE CONVERTERS WITH DIFFERENT LOADS

		$R_L = 200 \Omega$	$R_L = 300 \Omega$	$R_L = 400 \Omega$
PCV	I_{in}	1.46 A	0.91 A	0.65 A
	η	83.9%	89.74%	94.23%
C[13]	I_{in}	1.53 A	0.92 A	0.65 A
	η	78.03%	88.76%	94.23%
C[12]	I_{in}	1.77 A	0.96 A	0.68 A
	η	69.21%	85.07%	90.07%
C[10]	I_{in}	2.12 A	1.04 A	0.73 A
	η	57.78%	78.53%	83.90%

Fig. 9. Experimental efficiency of the proposed converter, the converters in [10], [12] and [13].

Fig. 10. Transient response of output voltage. (a) Load disturbance. (b) Input disturbance.

switching the input voltage from 25 V to 20 V at time t_4 , the output voltage can stabilize after a short time. The proposed converter has a good dynamic response.

IX. CONCLUSIONS AND FUTURE WORK

A novel quadratic boost converter is presented in this paper. The advantages of the proposed converter are listed as follows: the inductors' average currents are lower than that in [10], [12] and [13], that is helpful for reducing the copper losses of the inductors and increasing the efficiency of the converter; the input current is continuous; the voltage stress on power switch S_1 is relatively low. Steady state analysis and small signal model have been discussed under CCM. The average current mode control was applied in the experiment to achieve an output power of 24.5 W. The efficiency of the proposed converter can reach 95% when the load is 500 Ω . In the future work, two power switches will be soft-switched to improve the efficiency of the converter.

REFERENCES

- [1] F. Blaabjerg, F. Iov, T. Kerekes, and R. Teodorescu, "Trends in power electronics and control of renewable energy systems," in *Proceedings of 14th International Power Electronics and Motion Control Conference EPE-PEMC 2010*, Ohrid, 2010, pp. K-1-K-19.
- [2] A. A. Fardoun and E. H. Ismail, "Ultra step-up DC-DC converter with reduced switch stress," in *IEEE Transactions on Industry Applications*, vol. 46, no. 5, pp. 2025–2034, Sept. 2010.
- [3] S. Ben-Yaakov and I. Zeltser, "The dynamics of a PWM boost converter with resistive input," in *IEEE Transactions on Industry Applications*, vol. 46, no. 3, pp. 613–619, 1996.
- [4] Q. Zhao, F. Tao, and F. C. Lee, "A front-end DC/DC converter for network server applications," in *2001 IEEE 32nd Annual Power*

Electronics Specialists Conference, Vancouver, BC, 2001, pp. 1535–1539, vol. 3.

- [5] X. Liu, X. Zhang, X. Hu, H. Chen, L. Chen, and Y. Zhang, "Interleaved high step-up converter with coupled inductor and voltage multiplier for renewable energy system," in *CPSS Transactions on Power Electronics and Applications*, vol. 4, no. 4, pp. 299–309, Dec. 2019.
- [6] A. A. A. Freitas, F. L. Tofoli, E. M. S. Júnior, S. Daher, and F. L. M. Antunes, "High-voltage gain DC–DC boost converter with coupled inductors for photovoltaic systems," in *IET Power Electronics*, vol. 8, no. 10, pp. 1885–1892, 2015.
- [7] D. Sivaraj, M. Arounassalame, "High gain quadratic boost switched capacitor converter for photovoltaic applications," in *Power Control Signals and Instrumentation Engineering (ICPCSI) 2017 IEEE International Conference*, 2017, pp. 1234–1239.
- [8] M. Forouzesh, Yam P. Siwakoti, S. A. Gorji, F. Blaabjerg, and B. Lehman, "Step-Up DC–DC Converters: A Comprehensive Review of Voltage- Boosting Techniques Topologies and Applications," in *IEEE Transactions on Power Electronics*, vol. 32, no. 12, pp. 9143–9178, 2017.
- [9] D. S. Wijeratne and G. Moschopoulos, "Quadratic power conversion for power electronics: principles and circuits," in *IEEE Transactions on Circuits and Systems I: Regular Papers*, vol. 59, no. 2, pp. 426–438, Feb. 2012.
- [10] F. L. Luo and H. Ye, "Positive output cascade boost converters," in *IEE Proceedings–Electric Power Applications*, vol. 151, no. 5, pp. 590–606, 9 Sept. 2004.
- [11] M. G. Ortiz-Lopez, J. Leyva-Ramos, E. E. Carbajal-Gutierrez, and J. A. Morales-Saldana, "Modelling and analysis of switch-mode cascade converters with a single active switch," in *IET Power Electronics*, vol. 1, no. 4, pp. 478–487, 2008.
- [12] Y. M. Ye and K. W. E. Cheng, "Quadratic boost converter with low buffer capacitor stress," in *IET Power Electronics*, vol. 7, no. 5, pp. 1162–1170, 2014.
- [13] F. Wang, "A novel quadratic boost converter with low current and voltage stress on power switch for fuel-cell system applications," in *Renewable Energy*, vol. 115, pp. 836–845, 2017.
- [14] G. Li, X. Jin, X. Chen and X. Mu, "Analysis and Circuit Implementation of a Novel Quadratic Boost Converter with Low Inductor Current," in *2019 IEEE 10th International Symposium on Power Electronics for Distributed Generation Systems (PEDG)*, Xi'an, China, 2019, pp. 542–546.

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