A Novel Clamp-Mode Coupled-Inductor Boost Converter with High Step-Up Voltage Gain

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Abstract – In this paper, a new coupled inductor DC-DC converter with a high step-up voltage gain is proposed. It is developed from a clamp-mode coupled-inductor boost converter by incorporating an additional capacitor and diode. The proposed converter is able to achieve the higher voltage gain, while still retaining the switch voltage clamp property of its predecessor. In the paper, operation and analysis of the proposed converter are described. Experimental results from a prototype converter are presented to verify the validity of the analysis. The prototype circuit attains the highest efficiency of 92.8%.

Keywords: High step-up gain DC-DC converters, Clamp-mode coupled-inductor boost converter, DC-DC converter analysis

1. Introduction

Nowadays, a DC-DC converter with a high step-up voltage gain has become an essential component in modern power electronics applications [1-3]. For example, in automobile head lamps, a DC-DC converter is required to increase a battery voltage of 12V into 100V to supply a high-frequency inverter that drives a High Intensity Discharge (HID) lamp. In telecommunication centers, a power supply for a server computer is capable of working with dual input sources from an AC main and a 48V back-up battery. The battery voltage is fed to the power supply via a front-end DC-DC converter. When the main power outage occurs, the front-end converter provides a back-up power by boosting the battery voltage into a 380V bus voltage to continually supply downstream converters in the power supply system. Therefore, the server computer can continue to function despite the main power failure. Lastly, in power generation by solar PV or fuel cells, a DC-DC converter is often used to boost a small DC voltage generated from these renewable sources to a suitable value for a grid connected inverter. The DC-DC converters used in the above applications have the following common characteristics: (1) high step-up voltage gain, (2) high efficiency, and (3) no electrical isolation between input and output. The conventional power electronic circuit used for stepping up a DC voltage is a boost converter. In theory, the boost converter can produce a high voltage gain when operated at an extreme duty cycle (i.e. the duty cycle approaching one). However, in practice, due to component non-idealities, such as the inductor resistance, the diode

forward voltage drop, the switch on-state resistance etc., losses and voltage drops in the circuit increase with the increasing duty cycle. As a result, the efficiency and voltage gain of the boost converter are degraded, when the duty cycle is large [4, 5]. Besides, operating the boost converter at the extreme duty cycle means its rectifier diode has less time to conduct. The diode current thus becomes a narrow pulse with high current amplitude, which is responsible for a severe reversed recovery loss and EMI problems.

A simple solution to increase the voltage gain without extending the duty cycle is to use a cascade boost converter, in which two or more boost converters are connected in series [6]. This method, however, requires more power switches, i.e. one power switch per one boost stage, and suffers from low overall efficiency when a number of the boost stage is increased. In [7-10], an N-stage cascade boost converter with one active switch is proposed. It uses only one power switch and has the same voltage gain as the N boost converters connected in series. The main drawbacks of this topology, nonetheless, are high current and voltage stresses associated with the power switch, and the low efficiency due to multi-stage power processing. High gain DC-DC converters based on a coupled inductor [11-21] can overcome the efficiency and voltage stress problems of the aforementioned cascade converter. The coupled inductor converters achieve the high voltage gain by utilizing the coupled inductor's turn ratio, instead of cascading multiple boost stages, hence leading to the improved efficiency. Moreover, unlike the cascade converter which subjects its power switch to the output voltage during turn-off, a power switch in the coupled inductor converters operates with less voltage stress as the turn-off voltage is usually considerably less than the output voltage. In this respect, the advantage of the coupled inductor converters over the cascade converter becomes

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distinct when the output voltage is high. Among many coupled inductor converters, a clamp-mode coupled-inductor boost converter [12] is a very attractive topology. Its salient features include: (1) a simple circuit structure using only one power switch, (2) high voltage gain controlled by the coupled inductor's turn ratio, (3) a switch voltage clamp reducing the device voltage stress, and (4) high efficiency.

This paper proposes a new coupled inductor DC-DC converter with a high step-up voltage gain. It is developed from a clamp-mode coupled-inductor boost converter by incorporating an additional capacitor and diode. The proposed converter is able to achieve the higher voltage gain, while still retaining the switch voltage clamp property of its predecessor. Furthermore, the laboratory prototype demonstrates that the proposed converter operates with high efficiency. The reminder of this paper is organized as follows. Section 2 explains the evolution of the proposed converter. Section 3 describes the converter operation and analysis. Section 4 further simplifies the converter analysis via waveform approximation to facilitate the derivation of important current equations. Section 5 presents experimental results from the prototype converter. The results are used to verify validity of the analysis. Finally, conclusions are drawn in Section 6.

2. Evolution of Proposed Converter

The clamp-mode coupled-inductor boost converter is shown in Fig. 1(a). The coupled inductor has two windings, with N_1 and N_2 being a number of turns of primary and secondary windings. D_1 and C_1 constitute a clamp circuit. When SW is turned on, energy is stored in the coupled inductor, and the output capacitor C_3 is discharged to supply the load current. When SW is turned off, the stored energy is released to charge the output capacitor. Meanwhile, D_1 is turned on and the switch voltage is clamped to V_{C1} . During this turn-off interval, the output voltage is given by the sum of the input voltage and the induced voltages on the primary and secondary windings, which are additive. It is shown in [11, 12] that the voltage gain and clamp voltage of the clamp-mode coupled-inductor boost converter are given by

$$\frac{V_o}{V_{in}} = \frac{1+nD}{1-D}, \quad (1)$$

$$V_{C1} = \frac{V_{in}}{1-D}, \quad (2)$$

where $n=N_2/N_1$ is the coupled inductor's turn ratio and D is a duty cycle of SW . As seen in (1) and (2), the switch clamp voltage, V_{C1} , is lower than the output voltage, resulting in the low voltage stress on SW . Consequently, the switch can be implemented by an inexpensive low-voltage rated power MOSFET.

Fig. 1. (a) Clamp-mode coupled-inductor boost converter, and (b) the proposed converter

Fig. 2. Circuit variations of the proposed converter

The proposed converter in Fig. 1(b) is derived from the clamp-mode boost converter in Fig. 1(a) by the addition of C_2 - D_2 . Similar to its predecessor, the proposed converter stores energy in the coupled inductor during SW turn-on. When SW is turned off, the stored energy is released to charge the output capacitor and the switch voltage is clamped to V_{C1} via D_1 . During this turn-off interval, the output voltage is given by the sum of the input voltage, the induced voltages on the primary and secondary windings which are additive, and the voltage across C_2 (i.e. V_{C2}). The additional voltage provided by C_2 helps increase the

output voltage of the proposed converter, allowing it to achieve a higher voltage gain than the clamp-mode boost converter.

Other circuit variations of the proposed converter are depicted in Fig. 2. They are obtained by repositioning C_1 - D_1 , C_2 - D_2 , and C_3 of the circuit in Fig. 1(b). In these variant circuits, C_1 - D_1 serves to clamp the switch voltage and C_2 - D_2 serves to enhance the voltage gain. All the converters in Fig. 2 have the same voltage gain and switch clamp voltage as the proposed converter in Fig. 1(b). Operation and analysis of the proposed converter are described next.

3. Operation and Analysis of Proposed Converter

The converter operation is described and analyzed using the circuit shown in Fig. 3(a), where L_m is a magnetizing inductance of the coupled inductor. The primary and secondary voltages are related by: $v_2 = nv_1$, where $n = N_2/N_1$ is the coupled inductor's turn ratio. The primary and secondary currents are related by: $i_1 = ni_2$. To simplify the description and analysis, a leakage inductance of the coupled inductor is neglected. The input voltage, V_{in} , is constant. All the capacitor C_1 , C_2 and C_3 are assumed to be sufficiently large, so that V_{C1} , V_{C2} , and V_o are constant over one switching period. All the power semiconductor devices, i.e. SW , D_1 , D_2 , and D_3 , are assumed to be an ideal switch.

In Continuous Conduction Mode (CCM), where the magnetizing current, i_{Lm} , flows continuously and remains above zero throughout a switching period, the converter operation can be divided into three modes as follows.

Mode 1 (t_0 - t_1): The equivalent circuit is shown in Fig. 3(b). SW is turned on, D_2 is turned on, and D_1 and D_3 are turned off. The primary winding is supplied by V_{in} , causing i_{Lm} to increase linearly and energy being stored in L_m . Since D_3 is turned off, both primary and secondary currents, i_1 and i_2 , are zero. Meanwhile, C_1 is discharged by C_2 via D_2 and SW . At the output, C_3 is discharged by R , providing the load current, I_o .

From the equivalent circuit in Fig. 3(b), the following current relationships can be written:

$$i_{in} = i_{Lm}, \quad (3)$$

$$i_2 = 0, \quad (4)$$

$$i_{D1} = 0, \quad (5)$$

$$i_{SW} = i_{in} + i_{D2}, \quad (6)$$

where i_{in} is the input current, i_{SW} is the switch current, i_{D1} is a charging current from L_m to C_1 , and i_{D2} is a discharging current from C_1 to C_2 . This mode is ended at time t_1 when SW is turned off.

Mode 2 (t_1 - t_2): The equivalent circuit is shown in Fig. 3(c). SW is turned off, D_1 and D_3 are turned on, and D_2 is turned off. The primary voltage, v_1 , is equal to $V_{in} - V_{C1}$,

Fig. 3. Proposed converter: (a) circuit diagram; (b) operational mode 1; (c) operational mode 2, and (d) operational mode 3

which is negative because V_{C1} is greater than V_{in} . Alternatively, considering the outer loop of Fig. 3(c), v_1 can also be expressed as

$$v_1 = V_{in} + V_{C2} - v_2 - V_o = V_{in} + V_{C2} - nv_1 - V_o. \quad (7)$$

That is,

$$v_1 = \frac{V_{in} + V_{C2} - V_o}{(1+n)} = V_{in} - V_{C1}. \quad (8)$$

Since v_1 is negative, i_{Lm} decreases linearly and the energy stored in L_m is released to charge C_1 via D_1 and to charge C_3 via D_3 . The voltage across SW is clamped to V_{C1} . Meanwhile, C_2 is discharged by the current i_2 . Notice that C_2 is connected in series with the input voltage source and the primary and secondary windings of the coupled inductor; the sum of the voltages across these components is equal to the output voltage.

From the equivalent circuit in Fig. 2(c), the following current relationships can be written:

$$i_{in} = i_{Lm} - i_1 = i_{Lm} - ni_2, \quad (9)$$

$$i_{D1} = i_{in} - i_2 = i_{Lm} - (n+1)i_2, \quad (10)$$

$$i_{SW} = 0. \quad (11)$$

This mode is ended at time t_2 when $i_{D1} = 0$ (i.e. C_1 is fully charged).

Mode 3 (t_2 - t_3): The equivalent circuit is shown in Fig. 3(d). SW is turned off, D_1 and D_2 are turned off, and D_3 is turned on. The primary voltage, v_1 , is the same as in mode 2 given by (8). Since v_1 remains negative, i_{Lm} continues to decrease and the energy stored in L_m is released to charge C_3 via D_3 . Meanwhile, C_1 is floating. C_2 continues to be discharged by the current i_2 . Still, in this mode, C_2 is connected in series with the input voltage source and the primary and secondary windings of the coupled inductor; the sum of the voltages across these components is equal to the output voltage.

From the equivalent circuit in Fig. 3(d), the following current relationships can be written:

$$i_{in} = i_2, \quad (12)$$

$$i_2 = \frac{i_{Lm}}{n+1}, \quad (13)$$

$$i_{D1} = 0, \quad (14)$$

$$i_{SW} = 0. \quad (15)$$

This mode is ended at time t_3 when SW is turned on, thereby repeating mode 1 again.

Based on (3) to (15), key waveforms of the proposed converter can be drawn as shown in Fig. 4. The time duration when SW is turned on (i.e. mode 1) is DT , where D is the duty cycle of SW and T a switching period. The time duration when SW is turned off (i.e. mode 2 and 3) is $(1-D)T$. The current i_{D1} , which is a charging current from L_m to C_1 , has a triangle shape similar to that in the clamp-mode boost converter [12]. The current i_{D2} , which is a discharging current from C_1 to C_2 , is approximated by a pulse current with an amplitude I_p . The bottom two waveforms in Fig. 4 depict the currents i_{C1} and i_{C2} . C_1 is discharged by i_{D2} during the time duration DT and charged by i_{D1} during the time duration $(1-D)T$. C_2 is charged by i_{D2} during DT and discharged by i_2 during $(1-D)T$.

In steady state, the voltage v_1 in Fig. 4 averaged over one switching period must be zero, that is

$$\frac{V_{in}DT + (V_{in} - V_{C1})(1-D)T}{T} = 0. \quad (16)$$

The above equation yields

$$V_{C1} = V_{C2} = \frac{V_{in}}{1-D}. \quad (17)$$

Note that $V_{C1} = V_{C2}$ because C_1 and C_2 are connected in

Fig. 4. Key waveforms of the proposed converter

parallel during the operational mode 1 (Fig. 3(b)) and, in this analysis, the capacitor voltages are assumed to be constant over one switching period. It can be seen from (17) that the switch clamp voltage, V_{Cl} , of the proposed converter is identical to that in (2). This indicates that the proposed converter preserves the voltage clamping property of the clamp-mode coupled-inductor boost converter, despite the addition of C_2 - D_2 .

According to (8), Eq. (16) can be rewritten as

$$\frac{V_{in}DT + \frac{(V_{in} - V_o + V_{C2})}{1+n}(1-D)T}{T} = 0. \quad (18)$$

Solving (18), the voltage gain of the proposed converter is obtained as

$$\frac{V_o}{V_{in}} = \frac{2+nD}{1-D}. \quad (19)$$

Comparing (19) to (1), it is evident that the voltage gain of the proposed converter is greater than the clamp-mode

coupled-inductor boost converter. For example, if $n = 5$ and $D = 0.5$, the voltage gain of the proposed converter would be nine, whereas the clamp-mode boost converter would be seven. Fig. 5 plots the voltage gain in (19) as a function of the duty cycle, with the different coupled inductor's turn ratios. From the graph, it can be seen that, for a given duty cycle, the gain becomes larger as n is increased. Hence, by appropriately selecting n , the proposed converter can provide the high voltage gain, without operating at the extreme duty cycle.

By inspecting each operational mode in Fig. 3, the blocking voltages of power semiconductor devices in the proposed converter circuit can be determined as given below

$$V_{SW,max} = V_{C1} = V_o + nV_{in} - (n+1)V_{C1}, \quad (20)$$

$$V_{D1,max} = V_{C2}, \quad (21)$$

$$V_{D2,max} = V_{C2}, \quad (22)$$

$$V_{D3,max} = V_o + nV_{in} - V_{C1}, \quad (23)$$

where $V_{SW,max}$, $V_{D1,max}$, $V_{D2,max}$, and $V_{D3,max}$ are the blocking voltages of SW , D_1 , D_2 , and D_3 respectively. These equations provide a basis for determining voltage ratings of the power semiconductor devices used in the proposed converter.

4. Waveform Approximation and Analysis

To derive the expressions for $I_{Lm,max}$, $I_{sw,max}$, $I_{D1,max}$, $I_{2,max}$, and I_p in Fig. 4, the proposed converter is assumed to be operating with the time interval in mode 2 relatively short compared with the time interval in mode 1 and 3. As a result, the operating mode 2 can be omitted from the converter operation and the waveforms in Fig. 4 can be approximated as shown in Fig. 6, where the time interval t_1-t_2 becomes zero and the operating mode 3 takes up the entire $(1-D)T$ interval. It should be noted that as the time interval t_1-t_2 approaches zero, i_{D1} becomes an impulse current which is represented by the vertical arrow in Fig. 6. In figure, I_{Lm} , $I_{Lm,max}$, and $I_{Lm,min}$ are the average, maximum, and minimum magnetizing currents respectively. $I_{sw,max}$ is the maximum switch current. I_2 , $I_{2,max}$, and $I_{2,min}$ are the average, maximum, and minimum secondary currents respectively. $I_{2,nom}$, which is an average value between $I_{2,max}$ and $I_{2,min}$, is referred to as the nominal secondary current. The expressions for these current quantities are derived as follows.

Beginning with the secondary current, since i_2 is feeding C_3 in parallel with R , the average secondary current, I_2 , is equal to the output current, I_o :

$$I_2 = I_o. \quad (24)$$

Given I_2 in (24), $I_{2,nom}$ can be determined from

Fig. 5. Voltage gain vs duty cycle

Fig. 6. Approximated waveforms of the proposed converter

$$I_{2,nom}(1-D)T = I_2T. \quad (25)$$

The above equation gives

$$I_{2,nom} = \frac{I_o}{1-D}. \quad (26)$$

The average input current, I_{in} , in Fig. 6 can be expressed as

$$I_{in} = \frac{I_{Lm}DT + I_{2,nom}(1-D)T}{T} = I_{Lm}D + I_o. \quad (27)$$

For the lossless converter, the input power is equal to the output power, i.e.

$$I_{in} = \left(\frac{V_o}{V_{in}} \right) I_o. \quad (28)$$

Substituting V_o/V_{in} in (19) and I_{in} in (27) into (28), I_{Lm} is obtained as

$$I_{Lm} = \left(\frac{2+nD}{1-D} - 1 \right) \frac{I_o}{D}. \quad (29)$$

Given I_{Lm} in (29), the maximum and minimum magnetizing currents can be found as

$$I_{Lm,max} = I_{Lm} + \frac{\Delta I_{Lm}}{2} = I_{Lm} + \frac{V_{in}DT}{2L_m}, \quad (30)$$

$$I_{Lm,min} = I_{Lm} - \frac{\Delta I_{Lm}}{2} = I_{Lm} - \frac{V_{in}DT}{2L_m}. \quad (31)$$

Since $i_2 = i_{Lm}/(n+1)$ during the time duration $(1-D)T$, the maximum and minimum secondary current can be expressed in terms of $I_{Lm,max}$ and $I_{Lm,min}$ as

$$I_{2,max} = \frac{I_{Lm,max}}{(n+1)}, \quad (32)$$

$$I_{2,min} = \frac{I_{Lm,min}}{(n+1)}. \quad (33)$$

In steady state, the positive and negative areas under the i_{C2} waveform in Fig. 6 must be equal, i.e.

$$I_pDT = I_{2,nom}(1-D)T. \quad (34)$$

Substituting $I_{2,nom}$ in (26) into (34), I_p is obtained as

$$I_p = \frac{I_o}{D}. \quad (35)$$

The maximum switch current, $I_{SW,max}$, in Fig. 6 is given by

$$I_{SW,max} = I_{Lm,max} + I_p. \quad (36)$$

Finally, in steady state, the positive and negative areas under the i_{C1} waveform in Fig. 4 must be equal, i.e.

$$I_pDT = \frac{1}{2}t_{12}I_{D1,max}. \quad (37)$$

The above equation yields

$$I_{D1,max} = \frac{2I_pDT}{t_{12}} \quad (38)$$

where t_{12} is the time interval t_1-t_2 . If t_{12} is known, $I_{D1,max}$ can be calculated. It should be emphasized that $I_{Lm,max}$ in (30), $I_{2,max}$ in (32), I_p in (35), $I_{SW,max}$ in (36), and $I_{D1,max}$ in (38) provide a basis for determining current ratings of the power semiconductor devices used in the proposed converter. The validity of these derived current expressions is verified in the next section.

5. Experimental Results

5.1 Step-up Gain of 10

The proposed converter in Fig. 3 is prototyped as shown in Fig. 7. The input voltage, V_{in} , is 12V, the output voltage, V_o , is 120V, and the switching frequency is 100kHz (the switching period $T = 10\mu s$). The load current, I_o , can vary from 0.3A to 0.9A, which is equivalent to the output power between 36W and 108W. Table 1 lists the devices and components used in the prototype converter.

Fig. 8 shows the output voltage and current waveforms measured from the prototype for the three loading conditions, i.e. $I_o = 0.3A$, $I_o = 0.6A$, and $I_o = 0.9A$. In each case, the input voltage is fixed at 12V and the duty cycle of the MOSFET gate drive signal is adjusted to give the constant

Table 1. Devices and components used in the prototype converter

Component/Devices	Description
Coupled inductor	$L_m = 60\mu H$, $n=6$ ETD59 ferrite core, gap = 0.3mm $N_1 = 11$ turns, copper strip (width: 22mm) $N_2 = 66$ turns, 4xAWG20 copper wire
Capacitors	Electrolytic capacitor $C_1 = 180\mu F$ (100V) $C_2 = 180\mu F$ (100V) $C_3 = 180\mu F$ (100V)
D_1 , D_2 , and D_3	MUR840 (400V, 8A) ultrafast diode
SW	IRF3710 (100V, 57A) power MOSFET

Fig. 7. Photo of the prototype converter

output voltage of 120V. The duty cycles used to achieve this tenfold voltage gain are recorded and shown in Table 2. The duty cycle is 0.5 when $I_o = 0.3\text{A}$, is 0.53 when $I_o = 0.6\text{A}$, and is 0.56 when $I_o = 0.9\text{A}$. The duty cycle increases

(a) x-scale: 2.5μs/div

Top trace V_o : 50V/div, Bottom trace I_o : 0.2A/div

(b) x-scale: 2.5μs/div

Top trace V_o : 50V/div, Bottom trace I_o : 0.5A/div

(c) x-scale: 2.5μs/div

Top trace V_o : 50V/div, Bottom trace I_o : 0.5A/div

Fig. 8. Output voltage and current waveforms at: (a) $I_o = 0.3\text{A}$, (b) $I_o = 0.6\text{A}$, and (c) $I_o = 0.9\text{A}$

Table 2. Duty cycle and efficiency measurement for step-up gain of 10

I_o	V_o	I_{in}	V_{in}	Efficiency	D
0.30	120.22	3.24	12	92.8%	0.50
0.62	120.22	7.28	12	85.3%	0.53
0.92	120.22	11.55	12	79.8%	0.56

at high load currents to compensate for the increased voltage drop caused by circuit non-idealities, such as the leakage inductance and winding resistance of the coupled inductor, the MOSFET's on-state resistance ($R_{ds,on}$), the forward voltage drop of the diodes, etc. By substituting $V_{in} = 12\text{V}$, $V_o = 120\text{V}$, and $n = 6$ into (19), the theoretical duty cycle is calculated as $D = 0.5$. The actual duty cycles shown in Table 2 closely agree with the theoretical value.

Also shown in Table 2 is the efficiency measurement result. The converter efficiency, defined as a product of the average output voltage (V_o) and current (I_o) divided by a product of the average input voltage (V_{in}) and current (I_{in}), has the maximum value of 92.8% at $I_o = 0.3\text{A}$. The efficiency is reduced as the load current is further increased.

The decrease in efficiency at high load currents is caused by the increase in conduction losses of SW , D_1 , D_2 and D_3 .

Fig. 9 shows the key current waveforms measured from the prototype, when $I_o = 0.9\text{A}$. It can be seen that the measured current waveforms are similar to the theoretical waveforms in Fig. 4. The important current quantities, which include $I_{Lm,max}$, $I_{2,max}$, I_p , and $I_{SW,max}$ are measured and compared with the calculation based on (30), (32), (35), and (36) respectively. The comparison results are given in Table 3. It is seen that the calculated currents closely

(a) x-scale: 2.5μs/div

Top trace i_{in} : 10A/div, Bottom trace i_{SW} : 10A/div

(b) x-scale: 2.5μs/div

Top trace i_{D1} : 5A/div, Middle trace i_{D2} : 2A/div

Bottom trace i_{D3} : 2A/div

Fig. 9. Key current waveforms at $I_o = 0.9\text{A}$ for step-up gain of 10

Table 3. Comparison between calculated and measured currents

Quantity	$I_o = 0.3A$		$I_o = 0.6A$		$I_o = 0.9A$	
	Cal.	Mea.	Cal.	Mea.	Cal.	Mea.
$I_{Lm,max}$	5.9	5.8	11.9	11.8	18.5	18
$I_{2,max}$	0.8	0.7	1.7	1.6	2.6	2.6
I_P	0.6	0.6	1.1	1.2	1.6	2.0
$I_{sw,max}$	6.5	6	13	13	20.1	20

Table 4. Capacitor voltage measurement

Quantity	$I_o = 0.3A$		$I_o = 0.6A$		$I_o = 0.9A$	
	Cal.	Mea.	Cal.	Mea.	Cal.	Mea.
V_{C1}	24.2	24.8	25.4	26.7	27.0	28.8
V_{C2}	24.2	23.6	25.4	25	27.0	26.8

x-scale: 2.5μs/div
Top trace V_{C1} : 20V/div, Bottom trace V_{C2} : 20V/div

Fig. 10. Voltage across capacitor C_1 and C_2 at $I_o = 0.9A$

agree with the measured currents for all loading conditions. Furthermore, from the i_{D1} waveform in Fig. 9(b), the time interval t_1-t_2 is found to be $t_{12} = 2\mu s$. Given this time interval, $I_{D1,max}$ can be calculated from (38):

$$I_{D1,max} = \frac{2I_P DT}{t_{12}} = \frac{2 \times 1.6 \times 0.56 \times 10 \times 10^{-6}}{2 \times 10^{-6}} = 9A$$

The calculated $I_{D1,max}$ is consistent with the actual value in Fig. 9(b), where $I_{D1,max} \approx 10A$. The consistency between the calculated and measured results in all the above cases confirms the validity of the current expressions derived via waveform approximation in Section 4.

Fig. 10 shows the waveforms of V_{C1} and V_{C2} measured from the prototype, when $I_o = 0.9A$. Both capacitor voltages are relatively constant, with the amplitude $V_{C1} = 28.8V$ and $V_{C2} = 26.8V$. The capacitor voltage measurement for all loading conditions is summarized in Table 4. As seen in the table, the measured capacitor voltages conform with the calculation by (17). Note that V_{C1} is slightly higher than V_{C2} because, when C_1 is discharged to C_2 in mode 1 (Fig. 3b), there are some voltage drops across D_2 and SW .

Fig. 11 shows the voltage waveforms across the power semiconductor devices measured from the prototype, when $I_o = 0.9A$. As seen in the v_{sw} waveforms, there is an initial

(a) x-scale: 2.5μs/div

Top trace v_{SW} : 20V/div, Bottom trace v_{D1} : 20V/div

(b) x-scale: 2.5μs/div

Top trace v_{D2} : 20V/div, Bottom trace v_{D3} : 100V/div

Fig. 11. Voltage waveforms across power semiconductor devices at $I_o = 0.9A$

voltage spike when SW is turned off caused by the leakage inductance of the coupled inductor. After the voltage spike has died out, the switch voltage is clamped to $V_{C1} = 28.8V$, corresponding with (20). Subsequently, as the converter operation enters mode 3, SW is no longer clamped by C_1 and the switch voltage is reduced to about 26V imposed by the input voltage and primary voltage of the couple inductor (see Fig. 3(d)). Similarly, the initial voltage spike and the subsequent voltage transition are also observed in the v_{D1} and v_{D2} waveforms. As seen in the bottom trace of Fig. 11, v_{D3} exhibits a high frequency ringing, when D_3 is turned off. This is caused by the resonance between a leakage inductance of the coupled inductor and a parasitic capacitance of D_3 . It should be noted that in the experimental prototype, there is an RC snubber ($R=150\Omega$ and $C=3nF$) connected across D_3 ; therefore the effect of the resonance has been somewhat attenuated. Without the snubber, the ringing would be more pronounced. After the ringing has subsided, v_{D3} in Fig. 11 is subjected to the blocking voltage of 150V, which is lower than $V_{D3,max} = 165V$ calculated by (23).

Finally, a comparison is made between the proposed converter and the Integrated Boost Flyback Converter (IBFC) [5, 15] with regard to the stress of a power switch and converter efficiency. The IBFC prototype used in this

study has the same circuit specifications and coupled inductor parameters as the proposed converter. Fig. 12 shows the voltage and current waveforms of the MOSFET switch measured from the two prototype circuits. It can be seen that the peak switch voltages and currents in both converters are similar; hence the devices are essentially subjected to the same level of stresses. Despite the equal voltage and current stresses on the power switches, the voltage gain of the proposed converter (19) nonetheless is higher than the IBFC, whose gain is identical to the clamp-mode boost converter in (1). Thus, as evident in Fig. 12, the MOSFET switch of the IBFC has to operate with the greater duty cycle (the longer conduction time) than the

(a) x-scale: 2.5μs/div

Top trace v_{SW} : 20V/div, Bottom trace i_{SW} : 10A/div

(b) x-scale: 2.5μs/div

Top trace v_{SW} : 20V/div, Bottom trace i_{SW} : 10A/div

Fig. 12. Voltage and current waveforms of the MOSFET switch at $I_o = 0.9A$: (a) proposed converter (b) IBFC

Fig. 13. Efficiency comparison between the proposed converter and IBFC

proposed converter in order to achieve the tenfold step-up gain. The greater duty cycle causes the average switch current of the IBFC to be higher than the proposed converter. The higher average currents yield the greater conduction losses, which are responsible for the efficiency of the IBFC being lower than the proposed converter for the entire load currents, as shown in Fig. 13.

5.2 Step-up Gain of 5

The prototype converter in Fig. 7 is also experimented at a step-up gain of 5 with $V_{in} = 24V$, $V_o = 120V$ and $I_o = 0.3A$ to $1.8A$ (i.e. $P_o = 36W$ to $216W$). The circuit parameters and components are the same as those listed in Table 1. The purpose of this experiment is to demonstrate that the

Table 5. Duty cycle and efficiency measurement for step-up gain of 5

I_o	V_o	I_{in}	V_{in}	Efficiency	D
0.32	120.22	1.73	24	92.7%	0.25
0.62	120.22	3.53	24	88.0%	0.27
0.92	120.22	5.24	24	87.9%	0.28
1.24	120.22	7.25	24	85.7%	0.29
1.54	120.22	9.21	24	83.8%	0.30
1.84	120.22	11.38	24	81.0%	0.32

(a) x-scale: 2.5μs/div

Top trace i_{in} : 10A/div, Bottom trace i_{SW} : 10A/div

(b) x-scale: 2.5μs/div

Top trace i_{D1} : 5A/div, Middle trace i_{D2} : 5A/div

Bottom trace i_{D3} : 2A/div

Fig. 14. Key current waveforms at $I_o = 1.8A$ for step-up gain of 5

proposed converter can work satisfactorily at a different step-up gain. The duty cycle and efficiency measurement are shown in Table 5. When the load current is increased from $I_o = 0.3\text{A}$ to 1.8A , the duty cycle required to maintain V_o at 120V has increased slightly from 0.25 to 0.32, and the converter efficiency drops from 92.7% to 81%. As explained before, the higher duty cycle is needed to compensate for the increased voltage drop in the circuit at higher load power. Meanwhile, the efficiency decreases at higher load power due to the increase in conduction losses.

Fig. 14. shows the key current waveforms measured at the maximum load power ($P_o = 216\text{W}$). The waveforms generally resemble their counterparts in Fig. 9. The duty cycle of the MOSFET switch is evidently lowered for the step-up gain of 5 as can be seen by comparing the width of the i_{sw} waveforms in Fig. 9 and Fig. 14.

6. Conclusion

This paper has proposed a novel clamp-mode coupled inductor boost converter with a high step-up voltage gain. It was developed from the original clamp-mode converter in Fig. 1(a) by incorporating an additional capacitor (C_2) and diode (D_2). During a turn-off period of the main switch, C_2 is discharged in series with the input voltage source and the primary and secondary windings of the coupled inductor. As a result, the voltages across these components are added together to produce the output voltage. Due to the contribution by C_2 , the voltage gain of the proposed converter is higher than the clamp-mode boost converter. The operation of the proposed converter was described and analyzed, leading to the mathematical expressions for the voltage gain, switch clamp voltage, and various current quantities. The calculation by the derived expressions was found to be highly consistent with the measurement results from the prototype circuit, confirming the validity of the presented analysis. The experimental prototype converter was tested for step-up gain of 10 and 5. Both cases yield the similar value of maximum efficiency.

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