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High step up voltage gain achieved in DC-DC converters using Linear Peak Current Mode control technique

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ABSTRACT

Conventional dc-dc boost converters are unable to provide high step-up voltage gains due to the effect of power switches, rectifier diodes, and the equivalent series resistance of inductors and capacitors. In this paper a Linear Peak Current Mode Controller (LPCM) for a transformerless dc-dc converters is proposed to achieve high step-up voltage gain without an extremely high duty ratio. In the proposed converter, two inductors with same level of inductance are charged in parallel during the switch-on period and are discharged in series during the switch-off period. The structures of the proposed converter and controller are very simple. Only one power stage is used. Moreover, the steady-state analyses of voltage gains and boundary operating conditions are discussed in detail.

Index Terms—**DC–DC boost converter, high step-up voltage gain, power stage.**

1. INTRODUCTION

The DC-DC converter with high step up gain is used for many applications, such as high-intensity- discharge (HID) lamp ballast for automobile headlamps, fuel- cell energy conversion systems, solar-cell energy conversion systems and the battery back-up system for uninterruptible power supplies. Theoretically, the DC-DC boost converter can achieve high step-up voltage gain with an extremely high duty ratio. However, in practice, the step-up voltage gain is limited due to the effect of power switches, rectifier diodes and theequivalent series resistance (ESR) of inductors andcapacitors.Also, the extremely high duty-ratio operation will result in serious reverse-recovery problem. Many topologies have been presented to provide high step-up voltage gainwithout anextremely high duty ratio. The DC-DC fly back converter is a very simple structure with high step-up voltage gain and electrical isolation, but the active switch of this converter will suffer high voltage stress due to the leakage inductance of the transformer. For recycling the energy of the leakage inductance and minimizing the voltage stress on the active switch, some energy-regeneration techniques have proposed to clamp the voltage stress on the active switch and to recycle the leakage-inductance energy. The coupled-inductor techniques provide solutions to achieve high voltage gain, low voltage stress on the active switchand high efficiency without the penalty of high duty ratio. The transformer less DC-DC converters, which include the cascade boost type , the quadratic boost type , the voltage-lift type, the capacitor-diodevoltage multiplier type and the boost type integrating with switched-capacitor technique . However, these types are all complex and higher cost. For getting higher step-up voltage gain, the other DC-DC converters are also presented.Compared with the converter as shown in fig.1, proposed converter has the following merits: (i) two power devices exist in the current-flow path during the switch-on period, and one power device exists in the current-flow path during the switch-off period. (ii) The voltage stresses on the active switches are less than the output voltage. (iii) Under the same operating conditions, including input voltage, output voltage, and output power, the current stress on the active switch during the switch-on period equals a half of the current stress on the active switch of the converter in fig.1. The proposed DC-DC converters fig.2 utilize the switched-inductor technique, which two inductors with same level of inductance are charged in parallel during the switch-on period and are discharged in series during the switch-off period, to achieve high step-up voltage gain without the extremely high duty ratio. The operating principles and steady-state analysis are discussed in the following sections. To analyze the steady-state characteristics of the proposed converters, some conditions are assumed as: (1)All components are ideal. The on-state resistance R_{DS} (ON) of the active switches, the forward voltage drop of the diodes and the ESRs of the inductors and capacitors are ignored.(2) All capacitors are sufficiently large and the voltages across the capacitors can be treated as constant.The modified boost type with switched-inductor technique is shown in fig.1.

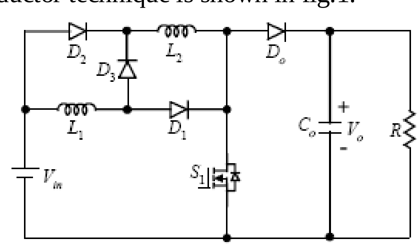

Figure 1Tranformerless DC-DC high step-up converter

The structure of this converter is very simple. Only one power stage is used in this converter. However, this converter has two issues: (i) threepower devices exist in the current-flow path during the switch-on period and two power devices exist in the current-flow path during the switch-off period. (ii) The voltage stress on the active switch equals the output voltage. When switch S_1 is ON the output voltage is zero, during this period Inductors L_1 and L_2 are charged. When switch S_1isOFF, the output voltage is appears.

2. PROPOSEDCONVERTER

A transformer less DC-DC high step-up converter is proposed as shown in figure 2. which consists of two active switches (S_1 and S_2),two inductors (L_1 and L_2) that have the same level of inductance, one output diode Do, and one output capacitor Co. Switches S_1 and S_2 are controlled simultaneously by using one control signal. Figure 2(d) and 2(e) shows some typical waveforms obtained during continuous conduction mode (CCM) and discontinuous conduction mode (DCM). The operating principles and steady-state analysis of CCM and DCM are presented in detail as follows.

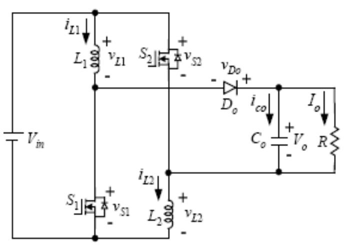

Figure 2 Proposed high step-up DC-DC converter I

2.1 CCM Operation:

The operating modes can be divided into two modes, defined as modes 1 and 2.

Mode 1 [t0, t1]:

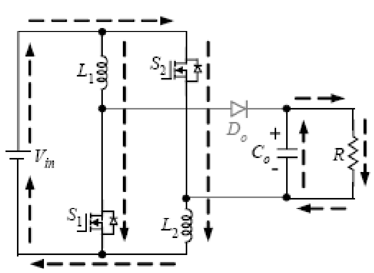

Figure 2 (a) Equivalent circuit of proposed converter I when switches ON

During this time interval, switches S_1 and S_2 are turned on. The equivalent circuit is shown in figure 2(a). Inductors L_1 and L_2 are charged in parallel from the DC Sourceand the energy stored in output capacitor C_0 is released to the load. Thus, the voltages across L_1 and L_2 are given as:

$$v_{L1} = v_{L2} = v_{in} \tag{1}$$

Mode 2 [t1, t2]:

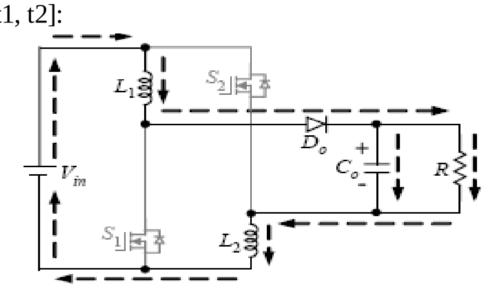

Figure 2(b) Equivalent circuit of proposed converter I when switches OFF

During this time interval, S_1 and S_2 are turned off. The equivalent circuit is shown in Fig. 2(b). The DC source, L_1 and L_2 are series-connected to transfer the energies to Co and the load. Thus, the voltages across L_1 and L_2 are derived as

$$v_{L1} = v_{L2} = \left(\frac{v_{in} - v_o}{2}\right) \tag{2}$$

By using the volt-second balance principle on $L_1$and L_2, the following equation can be obtained:

$$\int_0^{DT_s} v_{in} dt + \int_{DT_s}^{T_s} \left(\frac{v_{in} - v_o}{2}\right) dt = 0 \tag{3}$$

Simplifying (3), the voltage gain is given by

$$M_{ccm} = \left(\frac{v_o}{v_{in}}\right) = \left(\frac{1+D}{1-D}\right) \tag{4}$$

From figure 2(d), the voltage stresses on S_1, S_2, and D_O are derived as

$$v_{s1} = v_{s2} = \left(\frac{v_o + v_{in}}{2}\right)$$

$$v_{DO} = v_o + v_{in} \tag{5}$$

2.2 DCM Operation:

The operating modescan be divided into three modes, defined as modes 1, 2, and 3.

Mode 1 [t0, t1]: During this time interval, S_1 and S_2 are turned on. The equivalent circuit is shown in fig. 2(a).The two peak currents of L_1 and L_2 can be found as

$$I_{L1_P} = I_{L2_P} = \frac{v_{in}}{2} D T_s \tag{6}$$

Where L is the inductance of L_1 and L_2.

Mode 2 [t1, t2]: During this time interval, S_1 and S_2 are turned off. The equivalent circuit is shown in Figure 2(b). The DC source, L_1, and L_2 are series-connected to transfer the energies to Co and the load. Inductor currents i_{L1} and i_{L2} are decreased to zero at t = t2. Another expression of I_{L1p} and I_{L2p} is given as

$$I_{L1_P} = I_{L2_P} = \left(\frac{V_o - V_{in}}{2}\right) D_2 T_s \quad (7)$$

Mode 3 [t2, t3]:During this time interval, S_1 and S_2 are still turned off. The equivalent circuit is shown in Fig. 2(c).The energies stored in L_1 and L_2 are zero. Thus, only the energy stored in C_0 is discharged to the load.

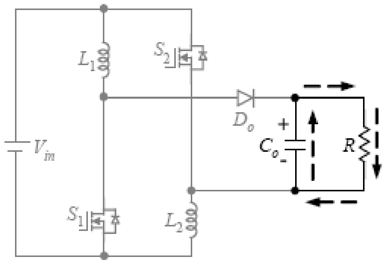

Figure 2(c) Equivalent circuit of proposed converter I when switches OFF in DCM mode

From (6) and (7), D_2 is derived as follows:

$$D_2 = \left(\frac{2D V_{in}}{V_o - V_{in}}\right) \quad (8)$$

From Figure 2(e), the average value of output-capacitor current during each switching period is given by

$$I_{CO} = \left(\frac{\frac{1}{2} D_2 T_2 I_{L1_P} - I_o T_S}{T_S}\right) = \frac{1}{2} D_2 I_{L1_P} - I_o \quad (9)$$

Substituting (6) and (8) into (9), Ico is derived as

$$I_{CO} = \left(\frac{D^2 V_{in} T_S}{L(V_o - V_{in})}\right) - \left(\frac{V_o}{R}\right) \quad (10)$$

Since I_{co}equals zero under steady state, equation (10) can bere-written as follows

$$\left(\frac{D^2 V_{in}^2 T_2}{L(V_o - V_{in})}\right) = \left(\frac{V_o}{R}\right) \quad (11)$$

Then, the normalized inductor time constant is defined as

$$\Gamma_L \equiv \left(\frac{L f_s}{R}\right) \quad (12)$$

wherefs is the switching frequency (fs = 1/Ts).

Substituting (12) into (11), the voltage gain is given by

$$M_{DCM} = \left(\frac{V_o}{V_{in}}\right) = \frac{1}{2} + \sqrt{\left(\frac{1}{4} + \frac{D^2}{\Gamma_L}\right)} \quad (13)$$

Typical waveforms for proposed converter I,

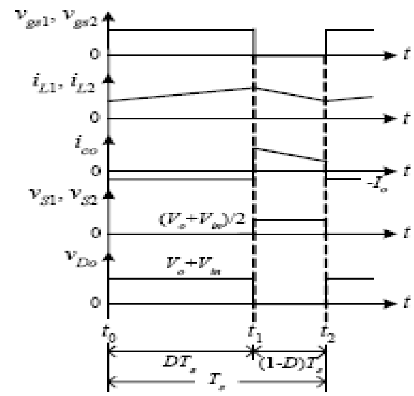

Figure 2 (d) CCM operation

2.3 Boundary Operating Condition between CCM and DCM

If proposed converter I is operated in boundary conduction mode (BCM), the voltage gain of CCM operation equals the voltage gain of DCM operation. From (4) and (13), the boundary normalized inductor time constant τ_{LB} can be derived as follows:

$$\Gamma_{LB} = \frac{D(1-D)^2}{2(1+D)} \quad (14)$$

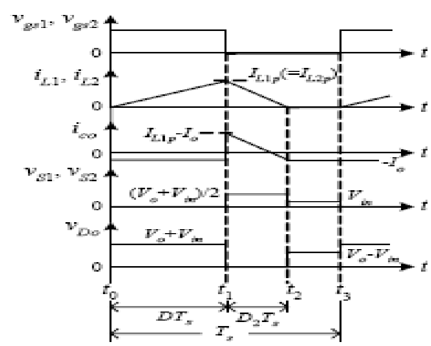

Figure 2(e) DCM operation.

The curve of $\tau_{LB}$is plotted in Figure 2(f).If τ_L is larger than τ_{LB}, proposed converter I is operated in CCM

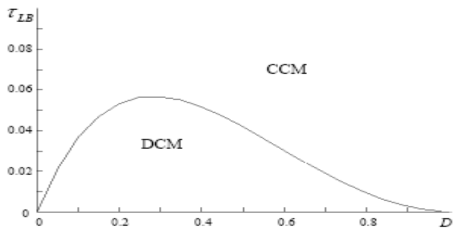

Figure 2(f) Boundary condition of proposed converter I

3. LINEAR PEAK CURRENT MODE CONTROL

Linear Peak Current Mode Control (LPCMC)-enables CCM operated rectifiers to be controlled using a much simpler controller Fig.3 (a).LPCMC offers the following advantages: Elimination of the controller multiplier and input voltages sensing circuits, uniconditional stability of the current loop, and ease of implementation using low standard PWM control IC's.The control technique is based on designing a current loop whose static gain is linearly dependent upon the off-duty cycle of the switch.

Current mode control:

For current-mode control there are three things to consider:

1. Current-mode operation. An ideal current-mode converter is only dependent on the dc or average inductor current. The inner current loop turns the inductor into a voltage controlled current source, effectively removing the inductor from the outer voltage control loop at dc and low frequency.
2. Modulator gain. The modulator gain is dependent on the effective slope of the ramp presented to the modulating comparator input. Each operating mode will have a unique characteristic equation for the modulator gain.
3. Slope compensation. The requirement for slope compensation is dependent on the relationship of the average current to the value of current at the time when the sample is taken. For fixed-frequency operation, if the sampled current were equal to the average current, there would be no requirement for slope compensation.

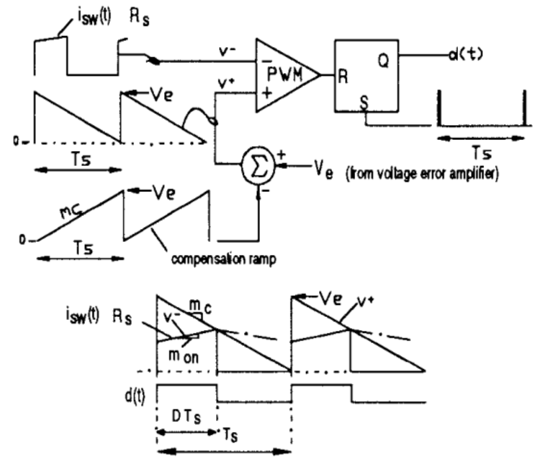

Figure 3(a) Linear Peak Current Mode Controller

The peak inductor current is expressed as

$$i_L(\text{peak}) = \frac{\mathrm{v_e}}{\mathrm{R_s}} \qquad (15)$$

Where $\mathrm{I_{ref}} = \frac{\mathrm{v_e}}{\mathrm{T_s}}$

Ve is the voltage error amplifier output signal, and RS is the current sensing resistor, the peak inductor current becomes

$$i_L(\text{peak}) = \mathrm{I_{ref}} - \frac{\mathrm{m_c T_s}}{\mathrm{R_s}} D \qquad (16)$$

mc is the slope of the compensating ramp, Ts is the switching period ,and D is the duty cycle. Consider re-writing equation (16) in terms of D'=1-D

$$i_L(\text{peak}) = \mathrm{I_{ref}} - \frac{\mathrm{m_c T_s}}{\mathrm{R_s}} D \quad (17)$$

By rearranging equation (17),we can express the static gain of current loop as

$$i_L(\text{peak}) = \left(\frac{\mathrm{v_e}}{\mathrm{R_s}} - \frac{\mathrm{m_c T_s}}{\mathrm{R_s}}\right) + \frac{\mathrm{m_c T_s}}{\mathrm{R_s}} D' \qquad (18)$$

Equation (18) shows a positive dependence of the static current loop gain of the off-duty cycle D'.Figure 3.2 plots iL(peak)/Iref versus D'.

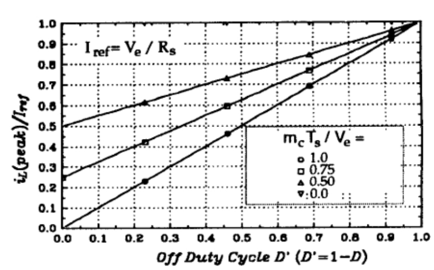

Figure 3(b) Reference current to peak current static gain

From equation (18), it is apparent that by choosing the compensation ramp appropriately, the first term will cancel, and the peak inductor current will be linearly related to D'.

$$i_L(\text{peak}) = \frac{m_c T_s}{R_s} D' \qquad (19)$$

Where $\frac{m_c T_s}{v_e} = 1$, or $m_c = \frac{v_e}{T_s}$ (20), (21)

With the compensating slope defined by equation (21) ,equation (19) can be rewritten as

$$i_L(\text{peak}) = \frac{v_e}{R_s} D' \qquad (23)$$

Equation (19) and (23) both reveal the linear relationship between the peak inductor current and the off-duty cycle D'.Hence the name Linear peak current control.

4. SIMULATION RESULTS

A Linear peak current mode control (LPCM) transformer less dc–dc converters with high step-up voltage gain has been simulated using MATLAB/Simulink. The simulation diagram of proposed controller to is shown in figure4.Simulation results are shown in fig.5 and fig.6.

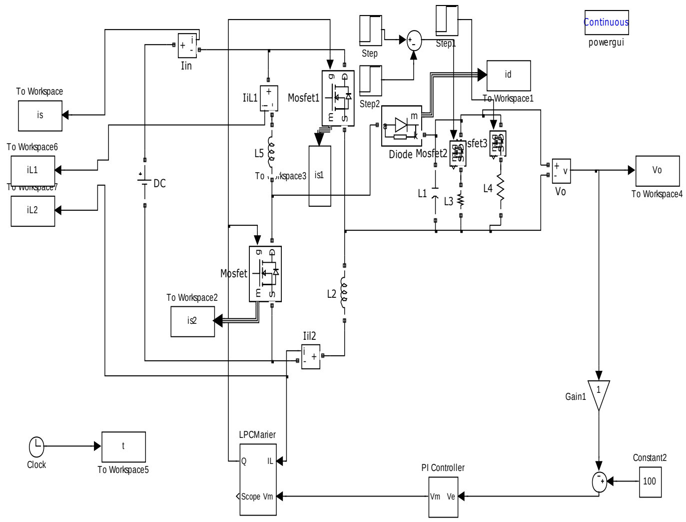

Fig4. Simulation diagram of LPCM Controller For Transformer Less Dc–Dc Converter With High Step-Up Voltage Gain

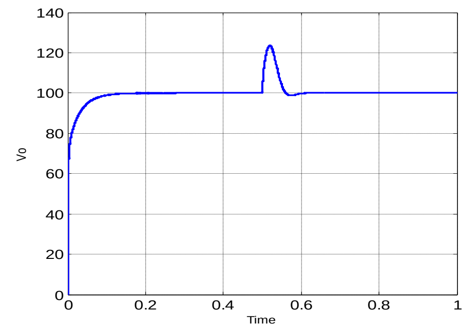

Fig5. Output voltage when load is changes from 40W to 5W at t=0.5Sec.

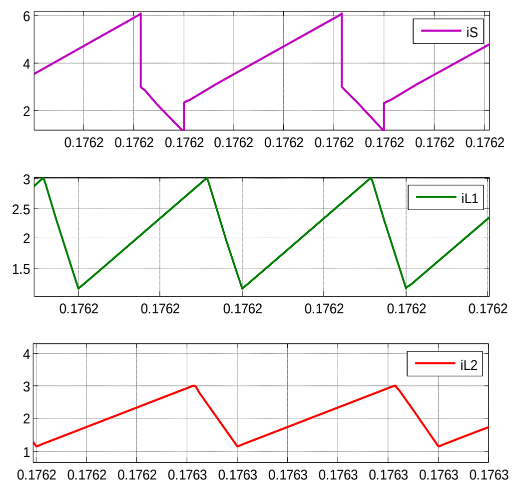

Fig6. Switch current, Inductor1 Current, Inductor2 Current waveform of Transformer Less Dc–Dc Converter With High Step-Up Voltage Gain.

5. CONCLUSION

A linear peak current mode control technique has been presented which enables simple, low cost. The technique possesses an inherently stable current loop, and the outer voltage loop is designed in a fashion similar to the other current mode control techniques. This paper has studied LPCM controller for transformerless dc–dc converter with high step-up voltage gain. Since the voltage stresses on the active switches are low, active switches with low voltage ratings and low ON-state resistance levels R_{DS}(ON) can be selected. The steady-state analyses of the voltage gain and the boundary operating condition are discussed in detail. Finally the controller concept was generalized to include average current mode controller. In this case the sensed signal was a filtered version of the inductor current. In LPCM controller multiplier present in conventional current mode controller is eliminated as a result of making profitable use of the inherent dutycycle dependent modulator gain.

6. REFERENCES

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