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 converter is proposed to achieve high step-up voltage gain without an extremely high duty ratio. In the proposed converter, two inductors with the same level of inductance are charged in parallel during the switch-on period and 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 the equivalent series resistance (ESR) of inductors and capacitors. Also, the extremely high duty-ratio operation will result in serious reverse-recovery problems. Although many topologies have been presented to provide high step-up voltage gain without an extremely 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 switch, and high efficiency without the penalty of high duty ratio. The transformerless DC-DC converters, which include the cascade boost type, the quadratic boost type, the voltage-lift type, the capacitor-diode-voltage multiplier type, and the boost type integrating with switched-capacitor technique. However, these types are all complex and high 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, the 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 in 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 of fig.2 utilize the switched-inductor technique, which two inductors with the same level of inductance are charged in parallel during the switch-on period and 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 1 Transformerless 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) three power devices exist in the current flow path during the switch-on period and two power devices exist in the current flow 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 \( S_1 \) is OFF, the output voltage is appears.

### 2. PROPOSED CONVERTER

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 \( D_0 \), and one output capacitor \( C_0 \). 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](image)

**2.1 CCM Operation:**

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

**Mode 1 \([t_0, t_1]\):**

![Figure 2 (a) Equivalent circuit of proposed converter I when switches ON](image)

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 Source and the energy stored in the output capacitor \( C_0 \) is released to the load. Thus, the voltages across \( L_1 \) and \( L_2 \) are given as:

\[
V_{L1} = V_{L2} = V_{in} \quad (1)
\]

**Mode 2 \([t_1, t_2]\):**

![Figure 2(b) Equivalent circuit of proposed converter I when switches OFF](image)

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 \( C_0 \) and the load. Thus, the voltages across \( L_1 \) and \( L_2 \) are derived as

\[
V_{L1} = V_{L2} = \left( \frac{V_{in} - V_o}{2} \right) \quad (2)
\]

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

\[
DT \int_{V_{in}}^{V_o} dt + \frac{1}{DT} \int_{V_{in}}^{V_o} \left( \frac{V_{in} - V_o}{2} \right) dt = 0 \quad (3)
\]

Simplifying (3), the voltage gain is given by

\[
M_{ccm} = \left( \frac{V_o}{V_{in}} \right) = \left( 1 + \frac{D}{1 - D} \right) \quad (4)
\]

From figure 2(d), the voltage stresses on \( S_1 \), \( S_2 \), and \( D_0 \) are derived as

\[
V_{s1} = V_{s2} = \left( \frac{V_o + V_{in}}{2} \right) \quad (5)
\]

**2.2 DCM Operation:**

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

**Mode 1 \([t_0, t_1]\):** 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

\[
I_{L1} = I_{L2} = \frac{V_{in}}{2} DT \quad (6)
\]

Where \( L \) is the inductance of \( L_1 \) and \( L_2 \).
Mode 2 \([t_1, t_2]\): 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 \(C_0\) and the load. Inductor currents \(i_{L1}\) and \(i_{L2}\) are decreased to zero at \(t = t_2\). Another expression of \(I_{L1p}\) and \(I_{L2p}\) is given as

\[
I_{L1p} = I_{L2p} = \left(\frac{v_{in}}{2} - \frac{v_o}{2}\right)DT, \quad (7)
\]

Mode 3 \([t_2, t_3]\): 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:

\[
D_2 = \left(\frac{2DV_o}{v_{in} - v_o}\right) \quad (8)
\]

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

\[
I_{co} = \left(\frac{\sqrt{2}DTI_{L1} - I_T}{T_s}\right) = \frac{1}{2}D_I_{L1} - I_s
\]

Substituting (6) and (8) into (9), \(I_{co}\) is derived as

\[
I_{co} = \left(\frac{D^2v_oT_s}{I(V_{in} - v_o)}\right) - \left(\frac{v_o}{R}\right)
\]

Since \(I_{co}\) equals zero under steady state, equation (10) can be rewritten as follows

\[
\left(\frac{D^2v_{in}T_s}{I(V_{in} - v_o)}\right) = \left(\frac{v_o}{R}\right)
\]

Then, the normalized inductor time constant is defined as

\[
\Gamma_i = \left(\frac{f_s}{R}\right)
\]

where \(f_s\) is the switching frequency \((f_s = 1/T_s)\).

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

\[
M_{DCM} = \left(\frac{v_o}{v_{in}}\right) - \frac{1}{2} + \left[\frac{1 + D}{\Gamma_i}\right]
\]

Typical waveforms for proposed converter I.

![Figure 2 (d) CCM operation](image)

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 \(\tau_{LB}\) can be derived as follows:

\[
\Gamma_i = \left(\frac{D(1 - D)^2}{2(1 + D)}\right)
\]

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

![Figure 2 (e) DCM operation](image)
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. 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.

The peak inductor current is expressed as

\[ i_{L}(\text{peak}) = \frac{V_e}{R_s} \]  \tag{15}

Where \( I_{\text{ref}} = \frac{V_e}{R_s} \) is the voltage error amplifier output signal, and RS is the current sensing resistor, the peak inductor current becomes

\[ i_{L}(\text{peak}) = I_{\text{ref}} - m_c T_s R_s D \]  \tag{16}

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

\[ i_{L}(\text{peak}) = I_{\text{ref}} - \frac{m_c T_s}{R_s} D' \]  \tag{17}

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

\[ i_{L}(\text{peak}) = \left( \frac{V_e}{R_s} - \frac{m_c T_s}{R_s} \right) + \frac{m_c T_s}{R_s} D' \]  \tag{18}

Equation (18) shows a positive dependence of the static current loop gain of the off-duty cycle \( D' \). Figure 3.2 plots \( i_L(\text{peak})/I_{\text{ref}} \) versus \( D' \).

Figure 3(a) Linear Peak Current Mode Controller

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 $D'$:

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

(19)

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

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

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

(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 is shown in figure 4. Simulation results are shown in fig. 5 and fig. 6.

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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