NEW TWO-INDUCTOR BOOST CONVERTER
WITH AUXILIARY TRANSFORMER

Yungtaek Jang and Milan M. Jovanović

Delta Products Corporation
Power Electronics Laboratory
P.O. Box 12173, 5101 Davis Drive
Research Triangle Park, NC 27709

Abstract — A new, two-inductor, two-switch boost converter topology and its variations suitable for applications with a large difference between the input and output voltage are described. The output voltage regulation of the proposed converters is achieved in a wide load and input-voltage range with constant-frequency control by employing an auxiliary transformer that couples the current paths of the two boost inductors.

I. INTRODUCTION

Generally, a single-inductor, single-switch boost converter topology and its variations exhibit a satisfactory performance in the majority of applications where the output voltage is greater than the input voltage. Nevertheless, in a number of high-power applications, the performance of the boost converter can be improved by implementing a boost converter with multiple switches and/or multiple boost inductors. Usually, multiple-switch and/or multiple-inductor boost topologies are employed in high input-current and/or high input-to-output voltage conversion applications. So far, a number of isolated and non-isolated multiple-switch and/or multiple-inductor topologies have been proposed, analyzed, and evaluated [1] – [8].

As an example, an interleaved boost topology is sometimes used in high-power applications to eliminate reverse-recovery losses of the boost rectifier by operating the two boost converters at the boundary of continuous-conduction mode (CCM) and discontinuous-conduction mode (DCM) so that the boost switches are turned on when the current through the corresponding boost rectifier is zero [1] – [4]. Generally, interleaving is employed to reduce the input current ripple and, therefore, to minimize the size of the input filter that would be relatively large if a single DCM boost converter were used. However, to achieve the operation at the CCM/DCM boundary under varying line and load-current conditions, the interleaved boost converter requires variable switching frequency control which is often complex to implement than constant-frequency control [1] – [4]. In addition, variable-frequency control in some applications is not desirable.

Another multi-switch boost converter is proposed in [5] for high-power applications that require an isolated PFC implementation. The circuit can also be employed in applications that require a high input-to-output-voltage conversion ratio. However, this current-fed push-pull converter suffers from a high voltage stress on the primary switches and high peak currents of the boost inductor and output capacitor.

Finally, the two-inductor, two-switch circuit shown in Fig. 1 that is described in [6] and [7] and analyzed in [8] exhibits some interesting properties. Specifically, the main feature of this circuit is that the voltage stress of each switch is one half the voltage stress of the switches in the single-inductor implementation in [5]. In addition, the input current is distributed evenly through the two boost inductors so that the current ripple in the output capacitor is smaller than in the single-inductor implementation. However, the major limitation of the two-inductor circuit in Fig. 1 is its inability to regulate the load in a wide range with constant-frequency control. To facilitate the explanation of this limitation, Fig. 2 shows key waveforms of the circuit in Fig. 1. As can be seen from Fig. 2, current \( i_{L1} \) in inductor \( L_1 \) increases during the entire on time of switch \( S_1 \) and decreases during the entire off time of switch \( S_1 \). Similarly, current \( i_{L2} \) in inductor \( L_2 \) increases during the on time of switch \( S_2 \) and decreases during its off time. As a result, even when converter’s duty cycle \( D \), which is defined as the ratio of the overlapping conduction time of the two switches and half of their switching period \( T_s \), is reduced to zero, the inductors continue to store energy since switches \( S_1 \) and \( S_2 \) are on for half of switching period \( T_s \). To reduce the stored energy and extend the load regulation range, it is necessary to shorten the conduction time of the switches. This can be accomplished by
increasing the switching frequency. Therefore, the circuit in Fig. 1 requires variable-frequency control to maintain the output voltage regulation in a wide load range.

In this paper, a new two-inductor, two-switch boost converter topology that can achieve output-voltage regulation from full load to no load in a wide input-voltage range using constant-frequency control is introduced. This topology employs an auxiliary transformer with a unity turns ratio to couple the current paths of the two boost inductors so that both inductors conduct identical currents. Due to this current-mirror effect of the auxiliary transformer, no energy is stored in the inductors when there is no overlapping of conduction times of the two switches, i.e., when $D=0$. This auxiliary-transformer approach can be applied to isolated or nonisolated two-inductor, two-switch topologies with any type of output rectifier.

II. ANALYSIS OF OPERATION

A nonisolated implementation of the boost circuit described in this paper is shown in Fig. 3. The input side of the circuit consists of two switches $S_1$ an $S_2$, two boost inductors $L_1$ and $L_2$, and auxiliary transformer ATR. To maximize the voltage gain of the converter, the output side of the circuit is configured as a voltage doubler rectifier that consists of boost rectifiers $D_1$ and $D_2$ and output filter capacitors $C_{F1}$ and $C_{F2}$ connected across load $R_L$.

To facilitate the explanation of the circuit operation, Fig. 4 shows a simplified circuit diagram of the circuit in Fig. 3. In the simplified circuit, auxiliary transformer ATR is modeled as an ideal transformer with turns ratio $n_{ATR}=1$ by assuming that its magnetizing inductance is high so that it can be neglected. In addition, it is assumed that filter capacitors $C_{F1}$ and $C_{F2}$ are large enough so that the voltage ripple across them is small compared to their dc voltages. Finally, in this analysis it is also assumed that all semiconductor components are ideal, i.e., that they represent zero impedances while in the on state and infinite impedances while in the off state.

To further facilitate the analysis of operation, Fig. 5 shows the topological stages of the circuit in Fig. 3 during a switching cycle, whereas Fig. 6 shows its key waveforms. The reference directions of currents and voltages plotted in Fig. 6 are shown in Fig. 4. As can be seen from the timing diagrams of the control signals for switches $S_1$ and $S_2$ shown in Fig. 6, switches $S_1$ and $S_2$ conduct simultaneously, i.e., they operate with overlapping control signals. The time of the simultaneous conduction, defined from the turn-on moment of one switch until the turn-off instant of the other switch, represents duty cycle period $D T_s/2$ of the converter, as indicated in Fig. 6.

From Fig. 4 it can be observed that because inductors $L_1$ and $L_2$ are connected in series with their corresponding transformer windings, inductor currents $i_{L1}$ and $i_{L2}$ must be equal at any given instant if the turns ratio of the transformer is unity, i.e., if $n_{ATR}=1$. Therefore, during the time interval when both switches are on, i.e., during the time interval $T_0 - T_1$ in Fig. 6, inductor currents $i_{L1}$ and $i_{L2}$ are increasing at the same rate. The rate of change of $i_{L1}$ and $i_{L2}$ can be calculated from Fig. 5(a), which represents the equivalent circuit diagram of the converter during the time interval $T_0 - T_1$. Since according to Fig. 5(a),

$$V_{IN} = v_1 + L_1 \frac{di_{L1}}{dt}$$  (1)

and

$$V_{OUT} = v_2 - L_2 \frac{di_{L2}}{dt}$$

Fig. 2. Key waveforms of conventional two-inductor boost converter shown in Fig. 1.
\[ v_1 + L_1 \frac{di_{l1}}{dt} = -v_2 + L_2 \frac{di_{l2}}{dt}, \quad (2) \]

it follows that the same rate of change of \(i_{l1}\) and \(i_{l2}\), i.e., \(di_{l1}/dt=di_{l2}/dt\), can only be satisfied if

\[ v = \frac{L_2 - L_1}{L_2 + L_1} V_{IN}, \quad (3) \]

where \(v=v_1=v_2\) because \(n_{ATR}=1\). Therefore, from Eqs. (1) through (3)

\[ \frac{di_{l1}}{dt} = \frac{di_{l2}}{dt} = \frac{V_{IN} - v}{L_1} = \frac{V_{IN} + v}{L_2}. \quad (4) \]

If both inductances have the same value \(L=L_1=L_2\), it follows that \(v=0\) and

\[ \frac{di_{l1}}{dt} = \frac{di_{l2}}{dt} = \frac{V_{IN}}{L}, \quad (5) \]

as indicated in Fig. 6.

The output is decoupled from the input when both switches are on and rectifiers \(D_1\) and \(D_2\) are reverse biased. As a result, during this stage the load current is supplied from the filter capacitors and capacitor voltages \(V_{CF1}\) and \(V_{CF2}\) slowly decrease, as seen in Fig. 6. Since output voltage \(V_O=V_{CF1}+V_{CF2}\), the output voltage also slowly decreases.

When at \(t=T_1\) switch \(S_2\) is turned off, inductor current \(i_{l2}\) is diverted from the switch to rectifier \(D_2\), as shown in Fig. 5(b), and the energy stored in inductor \(L_2\) starts to discharge into filter capacitor \(C_{F2}\). Since during this stage current \(i_{l2}\) decreases, current \(i_{l1}\) has to decrease at the same rate because the currents in the windings of the transformer are always equal. The rate of current decrease can be found from the equivalent circuit shown in Fig. 5(b). Therefore,

\[ V_{IN} = v_1 + L_1 \frac{di_{l1}}{dt} = v + L_1 \frac{di_{l1}}{dt} \quad (6) \]

and

\[ V_{AB} = -V_{CF2} = -L_1 \frac{di_{l1}}{dt} - v_1 - v_2 + L_2 \frac{di_{l2}}{dt} = -2v. \quad (7) \]

![Fig. 4. Simplified circuit model of proposed converter that shows reference directions of currents and voltages.](image)

![Fig. 5. Topological stages of proposed converter.](image)
From Eqs. (6) and (7), assuming \( L = L_1 = L_2 \), it follows that
\[
v = \frac{V_{CF2}}{2} = \frac{V_{CF}}{2} \tag{8}
\]
and
\[
\frac{di_{L1}}{dt} = \frac{di_{L2}}{dt} = \frac{1}{L} \left( V_{IN} - \frac{V_{CF2}}{2} \right) = \frac{1}{L} \left( V_{IN} - \frac{V_{CF}}{2} \right), \tag{9}
\]
where \( V_{CF} = V_{CF1} = V_{CF2} \).

Because during the topological stage shown in Fig. 5(b), current \( i_{L2} \) charges capacitor \( C_{F2} \), capacitor voltage \( V_{CF2} \) increases. At the same time, voltage \( V_{CF1} \) across capacitor \( C_{F1} \) continues to decrease because this capacitor continues to be discharged by the load current, as shown in Fig. 6.

When at \( t = T_2 \) switch \( S_2 \) is turned on again, the circuit enters the topological stage shown in Fig. 5(c), which is identical to the topological stage in Fig. 5(a). During this stage both switches are on and both inductor currents \( i_{L1} \) and \( i_{L2} \) increase at the same rate given by Eq. (5). At the same time, both output filter capacitors are being discharged by the load current since rectifiers \( D_1 \) and \( D_2 \) are reverse biased and the output part of the circuit is decoupled from the input part.

The converter enters the final topological stage shown in Fig. 5(d) at \( t = T_3 \) when switch \( S_1 \) is turned off and current \( i_{L1} \) is commutated from the switch into rectifier \( D_1 \). During this stage, energy stored in inductors \( L_1 \) and \( L_2 \) during the preceding topological stage discharges into capacitor \( C_{F1} \). The rate of decrease of currents \( i_{L1} \) and \( i_{L2} \) is given in Eq. (9). Due to the flow of current \( i_{L1} \) into capacitor \( C_{F1} \), voltage \( V_{CF1} \) increases, whereas voltage \( V_{CF2} \) continues to decrease because capacitor \( C_{F2} \) continues to be discharged by the load current. The circuit enter a new switching cycle at \( t = T_4 \) when switch \( S_1 \) is turned on again.

The voltage conversion ratio of the circuit can be calculated from the volt-second balance on the boost inductors. From Figs. 5 and 6, the volt-second balance equation for \( L_1 \) is
\[
V_{IN}D_1 \frac{T_s}{2} = \left( \frac{V_{CF}}{2} - V_{IN} \right) \left( \frac{T_s}{2} - D \frac{T_s}{2} \right), \tag{10}
\]
so that
\[
\frac{V_O}{V_{IN}} = \frac{4}{1 - D}, \tag{11}
\]
since \( V_O = 2V_{CF} \). As can be seen from Eq. (11), the output voltage of the converter in Fig. 3 is at least four times larger than the input voltage. This high conversion ratio makes this converter very suitable for applications with a large difference between the output and input voltage.

It also should be noted that because of the converter’s unique property to simultaneously charge and discharge both boost inductors due to the coupling of inductor currents through the auxiliary transformer, the converter can maintain the regulation of the output voltage with a constant frequency control in a wide range of the load current. Namely, with the duty cycle close to unity, the maximum power is transferred from the input to the output since the maximum energy is stored in the inductors. As the duty cycle decreases toward zero, less and less energy is stored in both inductors, which enables the output voltage regulation leg down to very light loads. Finally, it should be noted that if \( L_1 = L_2 = L \) and \( n_{ATR} = 1 \), both inductors in Fig. 3 store and transfer the same amount of energy, i.e., each of the converter processes one half of the total power. Since the total power is processed in two parallel legs, the conduction loss of the circuit is reduced compared to a circuit with a single power path.

Since, according to Eq. (11), proper operation of the circuit in Fig. 3 requires the output voltage to be at least four times greater than the input voltage, the output capacitors need to be precharged during a start-up period. This precharging can be implemented solely by a proper design of the control circuit, i.e., without any additional components. To accomplish this, switches \( S_1 \) and \( S_2 \) should be operated with complimentary switching (\( D = 0 \)) for a fixed time interval.

![Fig. 6. Key waveforms of proposed converter.](image-url)
during the start-up period. Since the magnetizing inductance of ATR is much larger than the boost inductor, ATR operates as an auto transformer with a 1:2 turns ratio when switches $S_1$ and $S_2$ operate with complimentary switching so that each output capacitor is charged to two times the input voltage. After the output voltage reaches approximately two times the input voltage, the duty cycle of the converter should be linearly increased to the steady-state value to obtain the desired output voltage.

Finally, it should be noted that a number of other implementations of the two-inductor, two-switch converter with the ATR are possible. As an illustration, Fig. 7(a) shows the implementation with an isolation transformer and a full-wave rectifier. The operation of this isolated circuit and its properties are similar to those of the non-isolated converter in Fig. 3. The circuit in Fig. 3 can also be implemented with a single magnetic component, as shown in Fig. 7(b). In this integrated magnetics implementation, the auxiliary transformer and the boost inductors can be integrated in a variety of ways. For example, the integration can be achieved by adjusting the coupling between the transformer windings so that the leakage inductance of the transformer windings is used as the boost inductances.

### III. EXPERIMENTAL RESULTS

#### A. Specifications

To verify the operation and evaluate the performance, a 1-kW, two-inductor boost rectifier was designed to the following specifications:

- Input Voltage $V_{in}$: 40 VDC - 70 VDC
- Input Current Ripple: < 5%
- Output Voltage $V_o$: 380 VDC (0 - 100% load)
- Power $P_o$: 1 kW
- Ripple Voltage: < 6.5 Vpeak-peak (80 kHz)
- Switch Frequency $f_s$: 40 kHz
- Efficiency: > 90% at full load
- Cooling: Forced Convection

#### B. Component Selections

**Semiconductors**

Since the drain voltages of switches $S_1$ and $S_2$ are clamped to output capacitors $C_{F1}$ and $C_{F2}$, respectively, the peak voltage stress on switches $S_1$ and $S_2$ is approximately 190 V. The peak current stress on switches $S_1$ and $S_2$ that occurs at full load and low line is approximately 27.5 A. Therefore, two IRFP264 MOSFETs ($V_{DSS} = 250$ V, $I_{DSS} = 38$ A, $R_{DS}$ = 0.075 Ω) from IR connected in parallel were used for each of the switches. Since output diodes $D_1$ and $D_2$ must block the output voltage and must conduct the peak load current which is 2.9 A, a RHRP1560 diode ($V_{RM} = 600$ V, $I_{AVM} = 15$ A) from Fairchild was used as output diodes $D_1$ and $D_2$. To reduce the conduction losses of the switches and output diodes, devices which have higher current ratings than the designed maximum current were selected.

**Boost Inductor**

To obtain the desired inductance of boost inductors $L_1$ and $L_2$ of approximately $L_1=L_2=30$ µH at full load, each boost inductor was built using a toroidal core (Magnetics, Kool-µ 77071-A7) and 58 turns of magnet wire (AWG #14). A ferrite toroidal core (Philips, Ferrite TX29/19/7.6) was used in parallel with a Kool-µ core to increase the continuous conduction mode operation boundary. Specifically, a Kool-µ core and a ferrite toroidal core were wound together using magnet wire.

**Auxiliary Transformer**

ATR was built using a pair of ferrite cores (Philips, ER28L-3F3) with an air gap (2 mils). Two AWG #14 magnet wires with equal number of turns (11 turns:11 turns) were used to obtain a desired magnetizing inductance which is approximately 100 µH.
Capacitors

Two aluminum capacitors (680 µF, 250 VDC) were used for output capacitors $C_{F1}$ and $C_{F2}$. Both capacitors have equal voltage stresses which is equal to half the output voltage, since the converter naturally balances the voltages of output capacitors $C_{F1}$ and $C_{F2}$.

Clamp Circuit

It should be noted that an accidental simultaneous opening of both switches would lead to a catastrophic circuit failure since the energy stored in the boost inductors would not have a path to discharge. Therefore, to prevent the circuit failure, it is necessary to provide a discharging path for the energy of the boost inductors when both switches are open. Figure 8 shows a protection circuit implemented with a RCD snubber connected across the switches. The clamp circuit consists of two diodes $D_{C1}$ and $D_{C2}$, capacitor $C_C$, and resistor $R_C$. An aluminum capacitor (25 µF, 250 VDC) was used for clamp capacitor $C_C$ and a carbon resistor (36 kΩ/2 W) was used for clamp resistor $R_C$.

C. Experimental Results

The performance of the proposed two-inductor boost converter was verified on a 1-kW prototype circuit that was designed to operate from a 40-70-V battery input and deliver up to 2.9 A at a 380-V output. The simplified schematic diagram is shown in Fig. 8. This converter was operated at 40 kHz switching frequency. However, the ripple of the input current and the ripple of the output voltage are 80 kHz, since the switching periods of the two switches are interleaved as shown in Fig. 6. The efficiency measurements for the prototype converter at 40 V and 70 V input are summarized in Fig. 9. As can be seen from the figure, the measured full-load efficiency was around 92% at the minimum line voltage.

Figure 10 shows the measured waveforms of the proposed two-inductor boost converter at 40 V and 70 V input voltages under the full load condition. As can be seen in the figure, the boost inductor current increases during the period when both switches $S_1$ and $S_2$ are on. During the period when one of the switches is off, the boost inductor current decreases. There is a good agreement between the experimental and theoretical waveforms. It should be noted that the blocking voltage of the switch is approximately 190 V which is half of the output voltage. Moreover, inductor current $i_L$ is also half of the

Fig. 8. Schematic diagram of 1-kW, two-inductor boost converter with auxiliary transformer.

Fig. 9. Measured efficiency of prototype circuit for $V_O=380$ V.
input current, since two inductors $L_1$ and $L_2$ share the current equally.

IV. SUMMARY

A new two-inductor, two-switch boost converter topology and its variations that can regulate the output voltage in a wide range of load current and input voltage with a constant-frequency control are introduced. The constant-frequency control is achieved with the employment of a unity-turns-ratio auxiliary transformer that couples the current paths of the two inductors and forces them to be virtually identical. Since in this topology no energy can be stored in the inductors when the conduction times of the switches do not overlap ($D=0$), the output voltage can be regulated from full load down to practically no load.

The described two-inductor topologies with the auxiliary transformer are suitable for applications that require a high input-to-output voltage conversion ratio. Specifically, a nonisolated implementation with a voltage-doubler rectifier exhibits a voltage gain that is four times of the corresponding gain of the conventional nonisolated boost converter.

The performance of the proposed two-inductor boost converter was verified on a 1-kW prototype circuit that was designed to operate from a 40-70-V battery input and deliver up to 2.9 A at a 380-V output.

REFERENCES


Fig. 10. Measured gate-to-source voltage waveforms $V_{GS1}$ and $V_{GS2}$, boost-inductor current waveform $i_L$, and drain-to-source voltage waveform $V_{S1}$, at: (a) $V_{IN}=40$ V, $P_O=1087$ W, and $V_O=380$ V; (b) $V_{IN}=70$ V, $P_O=1089$ W, and $V_O=380$. Time base 5 $\mu$s/div.