A Leakage-Inductance-Based ZVS Two-Inductor Boost Converter With Integrated Magnetics
Quan Li, Student Member, IEEE, and Peter Wolfs, Senior Member, IEEE,

**Abstract**—A two-inductor boost converter topology has conduction loss and transformer utilization advantages in converting low-voltage higher current inputs to high output voltages. In this letter, a new zero-voltage switching (ZVS) two-inductor boost converter with integrated magnetics is proposed. In the new topology, the two current source inductors, a resonant inductor and a two-winding transformer, are integrated into one single magnetic core with three windings. Two windings simultaneously perform the functions of the current source inductors and the transformer primary. The transformer leakage inductance forms the resonant inductance. This leads to a much more compact converter design with a significant reduction in the number of core and winding components. A theoretical analysis establishes the operating point of the ZVS converter. Both of the theoretical and experimental waveforms, including flux waveforms for the legs of the integrated core structure, are presented at the end of the letter.

**Index Terms**—Integrated magnetics, two-inductor boost converter, zero-voltage switching (ZVS).

I. INTRODUCTION

The two-inductor boost converter is well suited for low-voltage input, high-voltage output applications [1]. In order to combat the parasitic effects of the transformer leakage inductance in the hard-switched converter, a zero-voltage switching converter has been developed [2]. Fig. 1 shows the ZVS two-inductor boost converter with a voltage doubler rectifier. The quasiresonant converter is able to use the transformer leakage inductance and the MOSFET output capacitance as part of the resonant elements. This allows the switches to turn on at zero voltage and theoretically eliminates the switching losses. Higher efficiencies can be achieved under high switching frequencies and the converter volume and weight are minimized. However, the converter does require one additional resonant inductor and two additional resonant capacitors to form a resonant network [2]. This, in conjunction with the three existing magnetic cores required for the two inductors and the one transformer, contributes to bulk and weight.

This letter proposes a new ZVS two-inductor boost converter with integrated magnetics. In this topology, the two inductors and the one transformer are implemented with three copper windings on a single magnetic core and the resonant inductance is purely formed by the transformer leakage inductance. Compared with the discrete magnetics approach, the proposed converter has a significantly smaller component count and fewer interconnections between the windings. This results in a more compact design with a higher power density and better cost effectiveness.

II. ZVS TWO-INDUCTOR BOOST CONVERTER WITH INTEGRATED MAGNETICS

The demand for three separate magnetic components in the two-inductor boost converter tends to depart from the philosophy of “more silicon and less iron” in the design of the modern power electronic converters [3]. The last two decades have seen increasing interest in magnetic integration including both core and winding integration techniques [4], [5]. Among these, one particular application uses a core with three windings to implement a transformer and the two inductors in a current-doubler rectifier circuit [6]. In this application, the low-voltage transformer secondary and the individual inductor windings are integrated into a single winding located on each of the outer legs of a planer E core.

The current-doubler rectifier is related to the two-inductor boost converter by the principle of bilateral inversion [7]. If the diodes of the current doubler rectifier are replaced with active devices and the power flow reversed, a two-inductor boost converter results. As the primary side of the two-inductor boost converter shares a similar topology with the current doubler rectifier, the integrated magnetic scheme employed in [6] can be adopted by the two-inductor topology to merge the two input inductors and the one transformer into a single core with three windings.

The original current-doubler rectifier circuit using an integrated magnetic solution features a significant level of leakage inductance for the equivalent transformer as the primary and secondary windings are placed on different core legs. The current-doubler rectifier solution was proposed as a hard-switched converter [6]. The major effect of the leakage inductance in that application is a diode commutation overlap which reduces the available output voltage. The MOSFETs do not encounter over-voltage at turn-off as the current in one MOSFET is naturally taken by the body diode of the other. In the two-inductor...
A ZVS topology was developed to overcome this disadvantage of the conventional transformer designs, as shown in Fig. 2. The ZVS two-inductor boost converter proposed in this letter offers ZVS for the main switches [10]. The design limits the diode turn-off $di/dt$ and both turn-on and turn-off $di/dt$ for the main switches.

The resonant solution proposed in this letter offers ZVS for the main switches and additionally offers inherently favorable operating conditions for the output rectifiers which now process currents that are close to a sinusoidal waveform over the converter operating cycle and have even lower $di/dt$ values at their zero crossing. The ZVS two-inductor boost converter proposed is galvanically isolated by a high-frequency transformer with a simple magnetic structure. The design actively uses the major circuit parasitic components including transformer leakage and MOSFET capacitance to achieve a very low component count.

III. BASIC PRINCIPLES OF OPERATION

As the two-inductor boost converter with integrated magnetics can be modeled by the converter with discrete magnetics as shown in Fig. 3(b), the operation can be analyzed based on the converter in Fig. 1 if the resonant inductance $L_{RS}$ is converted to its equivalent value $L_r$ on the transformer primary side. The duty ratios of the MOSFETs $Q_1$ and $Q_2$ must be greater than 50% to ensure the continuity of the flux in each leg of the magnetic core. According to Fig. 2, the fluxes in the two outer core legs are restricted by the following equations:

$$N_p \frac{d\phi_1}{dt} = E - V_{C2} \tag{1}$$

$$N_p \frac{d\phi_2}{dt} = E - V_{C1} \tag{2}$$
The analysis is only demonstrated for the cycle from Q₁ turn-off to Q₂ turn-off. After Q₂ turns off, the converter will move through a similar group of states. The resonant capacitor voltage and the flux in the outer leg will be the same while the resonant inductor current or the transformer primary current will be reversed. Before Q₁ turns off, both Q₁ and Q₂ are closed. At time \( t = 0 \), Q₁ turns off and the converter will move up to four possible states before Q₂ turns off, as shown in Fig. 4, [2]. \( I₀ \) is the current in the inductors \( I₁ \) or \( I₂ \) and \( V_d \) is the output capacitor \( C₃ \) or \( C₄ \) voltage reflected to the winding that performs as both the inductance and the transformer primary.

1) State (a) \((0 ≤ t ≤ t₁)\): This state only exists if the resonant inductor carries a nonzero current when the MOSFET Q₁ turns off. The initial conditions are \( i_{Lₚ}(0) = -\Delta₁·I₀, v_{C₁}(0) = 0 \), and \( ϕ₂(0) = Φ₂₀ \), where \( Δ₁ \) is the ratio of the initial inductor current to the current source \( I₀ \) and \( Φ₂₀ \) is the initial flux. Both the current source \( I₀ \) and the resonant inductor current charge the capacitor \( C₁ \). While \( v_{C₁} \) increases, \( ϕ₂ \) increases but with a reducing rate as long as \( v_{C₁} < E \). When \( v_{C₁} > E \), \( ϕ₂ \) decreases with an increasing rate. The capacitor voltage \( v_{C₁} \), the inductor current \( i_{Lₚ} \) and the flux \( ϕ₂ \) are as shown in (3)–(5), at the bottom of the page, where \( Z₀ = \sqrt{L_p/C₁} \) is the characteristic impedance and \( ω₀ = 1/\sqrt{L_pC₁} \) is the angular resonant frequency of the resonant tank.

\[
v_{C₁}(t) = \frac{(1 + Δ₁)I₀Z₀\sin(ω₀t)}{ω₀}\cos(ω₀t) + \frac{V_d\cos(ω₀t)}{ω₀} - V_d
\]

\[
i_{Lₚ}(t) = \frac{V_d}{I₀} \sin(ω₀t) - (1 + Δ₁)I₀ \cos(ω₀t) + I₀
\]

\[
ϕ₂(t) = \frac{(E + V_d)ω₀t + (1 + Δ₁)I₀Z₀(\cos(ω₀t) - 1) - V_d \sin ω₀t + Φ₂₀}{ω₀V_p}
\]

2) State (b) \((t₁ ≤ t ≤ t₂)\): This state starts when the resonant inductor current reaches zero and will be bypassed if the resonant capacitor voltage is greater than \( V_d \) at this point. In this state, only the current source charges the capacitor. The capacitor voltage increases linearly while the inductor current stays at zero. The flux \( ϕ₂ \) encounters the same situation as in State (a).

The capacitor voltage \( v_{C₁} \), the inductor current \( i_{Lₚ} \) and the flux \( ϕ₂ \) are

\[
v_{C₁}(t) = \frac{I₀}{C₁}(t - t₁) + v_{C₁}(t₁)
\]

\[
i_{Lₚ}(t) = 0
\]

\[
ϕ₂(t) = \frac{(E - v_{C₁}(t₁))(t - t₁) - \frac{V_d}{ω₀}(t - t₁)^2}{N_p} + ϕ₂(t₁).
\]

3) State (c) \((t₂ ≤ t ≤ t₃)\): In this state, the circuit starts to establish a resonant inductor current in the positive direction when the inductor resonates with the capacitor. The flux \( ϕ₂ \) keeps decreasing with an increasing rate until \( v_{C₁} \) reaches its peak and continues to decrease as long as \( v_{C₁} > E \). After \( v_{C₁} \) falls below E, \( ϕ₂ \) again increases at an increasing rate. The capacitor voltage \( v_{C₁} \), the inductor current \( i_{Lₚ} \) and the flux \( ϕ₂ \) are as shown in (9)–(11), at the bottom of the page. If \( k \) is defined by \( I₀Z₀ = kV_d \), a simplified ZVS condition is \( k ≥ 1 \) according to (9). Equation (9) also determines the peak MOSFET voltage, which must be controlled within a reasonable level to allow an affordable MOSFET with a low forward resistance to be applied.

4) State (d) \((t₃ ≤ t ≤ t₄)\): At the beginning of this state, the capacitor voltage reaches zero and the MOSFET Q₁ turns on. The resonant inductor is discharged linearly by \( V_d \). The flux \( ϕ₂ \) increases linearly. The capacitor voltage \( v_{C₁} \), the inductor current \( i_{Lₚ} \) and the flux \( ϕ₂ \) are

\[
v_{C₁}(t) = 0
\]

\[
i_{Lₚ}(t) = i_{Lₚ}(t₃) - \frac{V_d}{I₀}(t - t₃)
\]

\[
ϕ₂(t) = \frac{E}{N_p}(t - t₃) + ϕ₂(t₃).
\]
As the resonant inductance in Fig. 2 is the winding leakage inductance, the center leg flux obeys

\[ N_s \frac{d\phi_c}{dt} = v_s. \]

When the secondary current is positive, \( v_s > 0 \) and \( \phi_c \) increases linearly, and vice versa. Due to the loose coupling of the primary and secondary windings, the leakage flux is significant and the flux paths are not constrained within the core structure. Therefore

\[ \phi_L = \phi_1 + \phi_2 + \phi_{Lk}. \]

where \( \phi_{Lk} \) is the leakage flux in the air.

IV. THEORETICAL AND EXPERIMENTAL WAVEFORMS

The proposed topology is validated experimentally by a 20-V input 40-W converter. A conversion efficiency of 93% has been recorded by using the mathematics functions of a Tektronix TDS5034 oscilloscope equipped with input and output voltage and current probes. The key components and parameters used in the converter are listed below.

- The two inductors and the transformer are implemented on a Philips ETD29 core with 0.5-mm air gaps in both outer legs. \( N_p = 10 \) turns and \( N_s = 13 \) turns. The transformer leakage inductance referred to the secondary is 12.5 \( \mu \)H.
- The switching frequency \( f_s = 500 \) kHz and the duty ratio \( D = 0.6 \).
- \( k = 1, A_1 = 1, A_9 \), and \( V_d/E = 1.15 \).
- \( L_r = 7 \mu \)H and \( C_1 = C_2 = 6.6 \) nF.

The theoretical waveforms under the above operating conditions are presented in Fig. 5. The corresponding experimental waveforms are shown in Fig. 6. From top to bottom, Figs. 5 and 6 show the MOSFET gate voltage, the resonant capacitor voltage and the transformer secondary current. These agree very well and the ZVS two-inductor boost converter with the new magnetic structure clearly operates in the same manner as a resonant converter constructed with discrete components.

The top two waveforms in Fig. 7 are, respectively, the ac flux waveforms of \( \phi_2 \) and \( \phi_c \) as recovered by integrating the voltage of a single search turn wound on the transformer leg. The bottom two waveforms are, respectively, the resonant capacitor voltage \( v_{C1} \) and resonant inductor current \( i_s \) and they are repeated here as references for the flux waveforms. It can be observed that \( \phi_2 \) decreases when \( v_{C1} \) is greater than \( E \) and \( \phi_c \) increases linearly when \( i_s > 0 \) and decreases linearly when \( i_s < 0 \).
V. CONCLUSION

A new soft-switched two-inductor boost converter with integrated magnetics is presented. The converter utilizes one magnetic core and three winding coils to implement the two inductors and one transformer. The transformer leakage inductance functions as the resonant inductance and no extra resonant inductors are required. Therefore, a minimum magnetic component count is achieved in the ZVS converter. The theoretical analysis of the converter operation is briefly provided. The experimental waveforms from a 40-W proof-of-concept prototype converter validate the theoretical analysis.

REFERENCES