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Single switch quasi-resonant ZVS converter with tapped inductor

Streszczenie. *Artykuł przedstawia quasi-rezonansowy przekształtnik podwyższający napięcie z dławikiem dzielonym przełączany przy zerowym napięciu. Wykorzystanie dławika dzielonego umożliwia uzyskanie wysokiego współczynnika wzmocnienia napięciowego. Obwód rezonansowy zapewnia przełączanie tranzystora przekształtnika przy zerowym napięciu. Miękkie przełączanie tranzystora daje możliwość zwiększenia częstotliwości pracy przekształtnika. W artykule szczegółowo omówiono pracę przekształtnika, została wyprowadzona jego charakterystyka sterowania. Omówiono zasadę optymalnego doboru wartości elementów rezonansowych. Artykuł kończy się przedstawieniem wyników badań laboratoryjnych potwierdzających prawidłowe działanie przekształtnika. (**Quasi – rezonansowy przekształtnik podwyższający napięcie z dławikiem dzielonym przełączany przy zerowym napięciu**).*

Abstract. *This paper presents a quasi-resonant zero-voltage switching boost converter with tapped inductor. Thanks to the use of a tapped inductor, a high voltage gain is achieved. A resonant circuit ensures that the converter transistor is switched at zero voltage. Soft transistor switching enables increase of converter operation frequency. The control is done by pulse width modulation and constant turn off time of the transistor. The soft-switching technique enables the converter to operate at higher frequencies. We provide a detailed discussion of the operation of the converter and present its control characteristics. Next, we discuss the importance of optimally choosing the values of resonant elements. Finally, the results of laboratory tests are presented.*

Słowa kluczowe: ZVS, tapped inductor, quasi-resonant, boost converter.
Keywords: przekształtnik podwyższający napięcie, dławik dzielony, quasi-rezonans, przełączanie przy zerowym napięciu.

Introduction

In recent years, boost converters have been used in many areas of industry. They are used in renewable energy systems, in the operating of photovoltaic panels, in uninterruptible power supply, in discharge lamps, in electric and hybrid vehicles, and in telecommunications. An increased interest in boost converters can also be observed in literature [1]. An important problematic issue with these converters is how to obtain a high voltage gain. In the case of a basic boost converter, this is not possible for several reasons. Firstly, in order to achieve a high gain, the converter transistor must be operated at a high duty cycle, which increases power losses in the transistor and parasitic resistances in the converter. Secondly, a high output voltage requires the use of a transistor with a high resistance $R_{DS(on)}$, which increases conduction losses in the transistor. Thirdly, in a plain boost converter, transistor switching losses and the reverse recovery time of the output diode put a limit on the transistor switching frequency. One of the ways to achieve a high voltage gain is by using a tapped inductor. However, this solution is problematic due to the need to suppress overvoltage caused by the current changes in leakage inductance to which the transistor is exposed. Literature describes many passive methods to reduce voltage stress of transistor [2 - 8]. One popular method is to use an additional diode connected in series with the output inductor and a capacitor for storing energy form leakage inductance [2, 3, 6]. An additional diode introduced into the main current loop of a converter adversely affects efficiency, due to reverse recovery current. Another method is to utilize a capacitor connected in series with the output inductance of the tapped inductor and add to the converter a circuit located outside the main current loop, responsible for reduced voltage stress [3, 4, 6, 7]. In converters modified in this way, the operating voltage of the transistor is reduced, enabling the use of transistors with lower resistance $R_{DS(on)}$. However, these converters require use of additional reactance and semiconductor components. A separate family of converters are quasi-resonant converters, in which sinusoidal current and voltage change enables soft switching of transistor and diode. So far scientific literature has not described ZVS quasi-resonant converters with tapped inductor running in autotransformer connection. The closest analogues are converters with a coupled inductor. In these converter, a capacitor is placed in series with one of the inductors, forming a resonant circuit along with leakage inductance [5, 8]. This solution eliminated the adverse effects associated with the reverse recovery of diodes and improves the operating conditions of the transistor. However, these topologies contain many additional reactance and semiconductor elements and are rather complex in comparison with a basic boost converter.

Proposed converter

This article presents a zero voltage switching quasi-resonant boost converter with tapped inductor (Fig. 1). Similar converter with zero current switching (ZCS) has been presented in [9]. The converter is equipped with a resonant circuit consisting of inductor L_r and capacitor C_r. It ensures that the voltage change in the transistor and the current change in the output diode are quasi-sinusoidal. This almost completely eliminates switching losses of semiconductor components. Tapped inductor enables to achieved a high voltage gain. The converter does not require an additional circuit to reduce overvoltage caused by current changes in leakage inductance. The proposed converter design requires fewer components in relations to the previously described solutions, and enables operation at higher frequencies, which makes this converter economical.

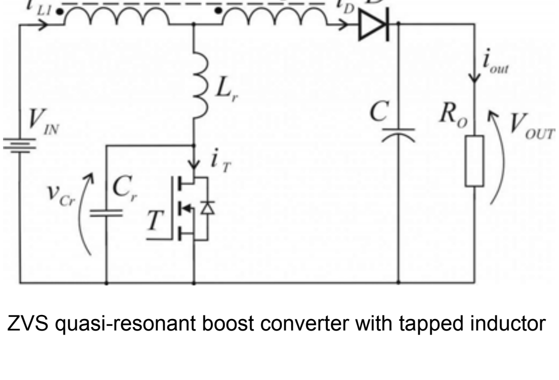

Fig.1. ZVS quasi-resonant boost converter with tapped inductor

The operation of the converter has been analyzed on the assumption that the amount of energy accumulated in the tapped inductor and the voltage of the source V_{IN} and capacitor C during operation are constant. The switching on/off times of diode D and transistor T as well as the parasitic properties converter components have been disregarded. The operation of the converter can be divided into six modes, presented in Figures 2 and 3.

Mode t_1 - t_2

At time t_1 transistor T is turned off at zero voltage. Diode D is polarized in the reverse direction. Energy from output capacitor C is transferred to the load. Capacitor C_r is charged with constant current until diode D starts to conducts. Interval t_1 - t_2 can be described as (1) – (3).

(1) $$V_{Cr} = \frac{V_{OUT} - V_{IN}}{N+1} + V_{IN} = \frac{V_{OUT} + V_{IN}N}{N+1}$$

(2) $$t_1 = \frac{V_{Cr}C_r}{I_{L1}}$$

(3) $$t_2 - t_1 = \frac{C_r\left(V_{OUT} + V_{IN}N\right)}{I_{L1}\left(N+1\right)}$$

where: I_{L1} - current i_{L1} at time t_1, N – turns ratio of tapped inductor, V_{Cr} - voltage on capacitor C_r at time t_2.

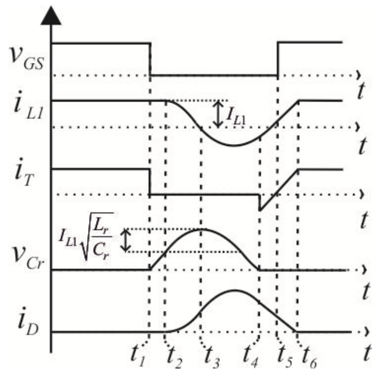

Fig.2. Key waveforms of the proposed dc–dc converter

Mode t_2 - t_3

At the time t_2 resonant operation between capacitor C_r and inductor L_r is started. Diode D polarized in the forward direction, the energy accumulated in the tapped inductor is transferred to the load consisting of resistor R_O and capacitor C connected in parallel. Input current is decreasing to zero. The resonant circuit of the converter can be described by the formulas in (4) – (6).

(4) $$\omega_r = \frac{1}{\sqrt{L_rC_r}}$$

(5) $$f_r = \frac{1}{2\pi\sqrt{L_rC_r}}$$

(6) $$Z = \sqrt{\frac{L_r}{C_r}}$$

Mode t_3 - t_4

At the time t_3 input current is equal to zero, voltage on the capacitor C_r reaches its maximum value. Inductor L_r and capacitor C_r continue their resonant operation. The current of capacitor C_r, inductor L_r and input during the time t_2 - t_4 can be described by the formulas in (7) – (8).

(7) $$i_{Cr}(t) = i_{Lr}(t) = I_{L1}\cos(\omega_r t)$$

(8) $$i_{L1}(t) = i_{Cr}(t) + i_D(t)$$

The relationship between the current of inductor L_1 and diode D is given by (9).

(9) $$i_{L1}(t) + Ni_D(t) = I_{L1}$$

Current of inductor L_1 and diode D in the interval t_2 - t_4 is given by (10) - (11).

(10) $$i_{L1}(t) = \frac{I_{L1}}{N+1}\left(N\cos(\omega_r t) + 1\right)$$

(11) $$i_D(t) = \frac{I_{L1}}{N+1}(1 - \cos(\omega_r t))$$

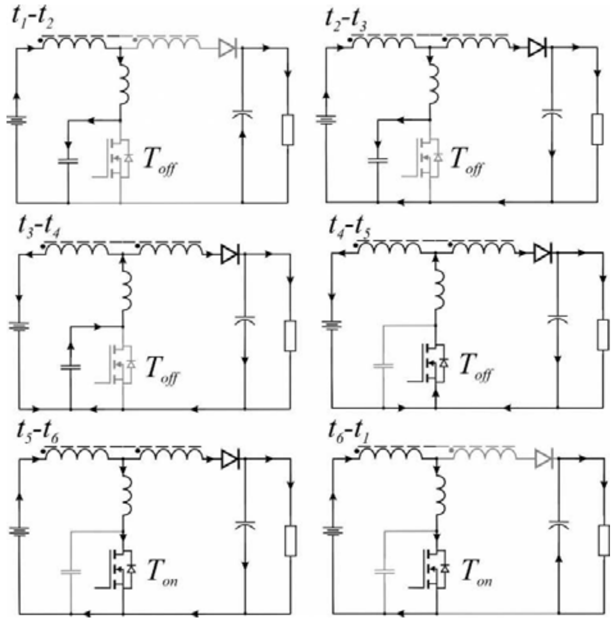

Fig.3. Operating modes of the proposed dc–dc converter

The t_2 - t_4 interval terminates with the discharge of resonant capacitor C_r. The duration of the interval depends on the voltage on the capacitor C_r, the current of inductor L_1 at moment t_2, the resonance impedance and turns ratio of tapped inductor. Introducing the parameter Ψ (formula (12)), time $t_2 - t_4$ can be given by (13).

(12) $$\psi = \frac{V_{Cr}}{I_{L1}\sqrt{L_{Lr}/C_r}} = \frac{V_{OUT} + V_{IN}N}{I_{L1}\sqrt{L_r/C_r}\left(N+1\right)} \le 1$$

(13) $$t_4 - t_2 = \frac{\pi + arc\sin\left(\psi\right)}{\omega_r}$$

Mode t_4 - t_6

At the time t_4, body diode of transistor T starts conducting. The current of resonant inductor L_r increase linearly. The input current increases, and the diode D current decreases to zero. At the time t_5, transistor T is turned on at zero current and zero voltage. The t_4 - t_6 interval terminates when the current of diode D drops to zero. The duration of this interval is dependent on the difference of inductor L_1 currents at times t_4 and t_6, and the voltage and inductance of resonant inductor L_r. The interval

t_4 - t_6 is described by the formulas in (14) – (16).

$$i_{Lr}(t_6)-i_{Lr}(t_4)=I_{L1}\left(1-\cos(\omega_r t_2)\right) \tag{14}$$

$$t_6-t_5=\frac{L_r I_{L1}\left(N+1\right)}{V_{OUT}+V_{IN}N} \tag{15}$$

$$t_6-t_4=\frac{L_r I_{L1}\left(1-\cos(\omega_r t_2)\right)\left(N+1\right)}{V_{OUT}+V_{IN}N} \tag{16}$$

Mode t_6 - t_1

In the interval t_6 - t_1, diode D is polarized in the reverse direction. The energy stored in the electric field of capacitor C is transferred to the load. The transistor T is turned on, and energy from the voltage source V_{IN} is stored in the magnetic field of the tapped inductor. The interval t_6 - t_1 is given by (17).

$$t_1-t_6=T_{on}-\left(t_6-t_5\right) \tag{17}$$

Voltage gain of the converter

Voltage gain of a converter can be determined by comparing its input and output power according to the equation in (18).

$$V_{IN}I_{IN}=\eta V_{OUT}I_{OUT} \Rightarrow \eta\frac{V_{OUT}}{V_{IN}}=\frac{I_{IN}}{I_{OUT}} \tag{18}$$

where: I_{out} – average load current, I_{in} - average input current η - converter efficiency.

On the basis of the waveforms given in Figure 2 and using the formulas (1) - (17), average input current can be determined by (19) - (20). Output current is equal to the average diode D current (formulas (21) - (22)). Using the formulas in (19) - (22), one can determine the voltage gain according to the formula in (23). Using the equations (19) - (22) the voltage gain is given by (23).

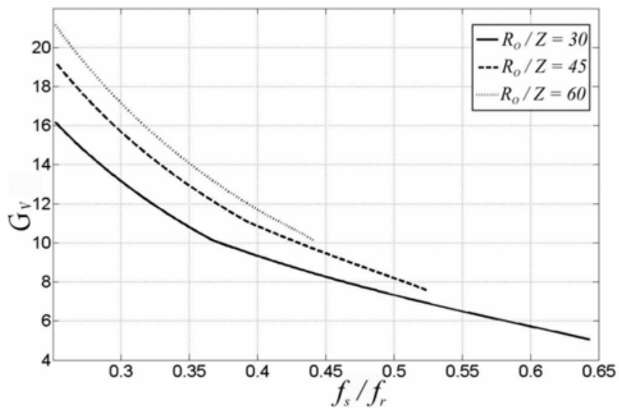

Fig.4. Voltage gain of proposed converter for N = 6 and η = 1

For transistor T to be switched on at zero voltage, the inequality (12) must be satisfied. The formula can also be expressed as in (24).

$$I_{IN}=\frac{1}{T_S}\left(I_{L1}T_S-\int_0^{t_4-t_2}\left(I_{L1}(1-\cos\left(\omega_r\left(t_4-t_2\right)\right)\right)dt-I_{L1}\left(1-\cos\left(\omega_r\left(t_4-t_2\right)\right)\right)\frac{t_6-t_4}{2}\right) \tag{19}$$

$$I_{IN}=I_{L1}\left(1-\frac{N\left(\pi+\arcsin\left(\psi\right)\right)+\psi+\frac{\left(1+\sqrt{1-\psi^2}\right)^2}{2\psi}}{2\pi(N+1)}\right)\frac{f_s}{f_r} \tag{20}$$

$$I_{OUT}=\frac{1}{T_S}\left(\int_0^{t_4-t_2}\left(I_{L1}(1-\cos\left(\omega_r\left(t_4-t_2\right)\right)\right)dt-I_{L1}\left(1-\cos\left(\omega_r\left(t_4-t_2\right)\right)\right)\frac{t_6-t_4}{2}\right) \tag{21}$$

$$I_{OUT}=\frac{I_{L1}\left(\pi+\arcsin\left(\psi\right)+\psi+\frac{\left(1+\sqrt{1-\psi^2}\right)^2}{2\psi}\right)}{2\pi(N+1)}\frac{f_s}{f_r} \tag{22}$$

$$G_V=\eta\frac{I_{IN}}{I_{OUT}}=\eta\frac{1-\frac{N\left(\pi+\arcsin\left(\psi\right)+\psi+\frac{\left(1+\sqrt{1-\psi^2}\right)^2}{2\psi}\right)}{2\pi(N+1)}}{\frac{\pi+\arcsin\left(\psi\right)+\psi+\frac{\left(1+\sqrt{1-\psi^2}\right)^2}{2\psi}}{2\pi(N+1)}\frac{f_s}{f_r}} \tag{23}$$

$$\psi=\frac{R_O\left(G_V+N\right)\left(\pi+arc\sin\left(\psi\right)+\psi+\frac{\left(1+\sqrt{1-\psi}\right)^2}{2\psi}\right)}{2\pi ZNG_V\left(N+1\right)}\frac{f_s}{f_r}\le 1 \tag{24}$$

Figure 4 shows the voltage gain characteristics of the analyzed converter for turn ratio of tapped inductor N = 6. The characteristics have been determined on the basis of the formula in (23), with allowance for the inequality in (24), for three different values of the relationship between the load resistance and the resonant impedance. Since the equations in (23) - (24) are implicit, the characteristics have been determined numerically. The switching frequency of transistor T should ensure its operation according to the control characteristics shown in Figure 4. Increasing the transistor switching frequency or decreasing the output resistance of the converter may prevent transistor T from switching on at zero voltage (due to failure to satisfy the inequality in (24)).

Experimental results

A prototype of the converter has been built. Its parameters are shown in table 1.

Table 1. Parameters of the laboratory prototype

Parameter	Value / part number
input voltage V_{IN}	(30 - 40) V
output voltage V_{OUT}	(180 - 440) V
output power P_{OUT}	(25 - 100) W
switching frequency f_s	(84 - 125) kHz
resonant frequency f_r	235 kHz
input inductor L_1	50 μH
output inductor L_2	1800 μH
turns ratio N	6 (17 :102) on powder cores MS-184075
coupled coefficient k	0.98
resonant capacitor C_r	22 nF (polypropylene)
resonant inductor L_r	21 μH on powder cores MS-090026
diode D	SiC diode C4D05120A (V_F = 1.2 V)
transistor T	IRFP4868 (V_{DSS} = 300V, R_{DSon} = 25.5 mΩ)

The values of individual components of the converter were selected for the output power of (25 – 100) W, input voltage of (30 - 40) V, and a voltage gain of 6 to 10. The control characteristics in Figure 4 were determined under the assumption that the energy accumulated in the tapped inductor remains constant during the converter operation. In a real system, this energy is variable; the inductance of the tapped inductor and the switching time of transistor T affect the current ripples of inductor L_1, which in its turn affects the control characteristics of the converter (formula (12)). Assuming that the transistor is switched at the boundary conduction mode (BCM), in voltage gain formula (12), (23), (24) current I_{L1} should be multiplied by two when determining the control characteristics. The voltage gain characteristics determined in this way, together with the experimental results, are shown in Figures 5 - 6. A comparison between the characteristics thus derived and the characteristics presented in Figure 4 reveals that an increase in the current ripples of the tapped inductor enables the transistor to operate in a wider range of frequency changes. At the same time, it allows for soft switching of the transistor at a lower voltage gain and output resistance.

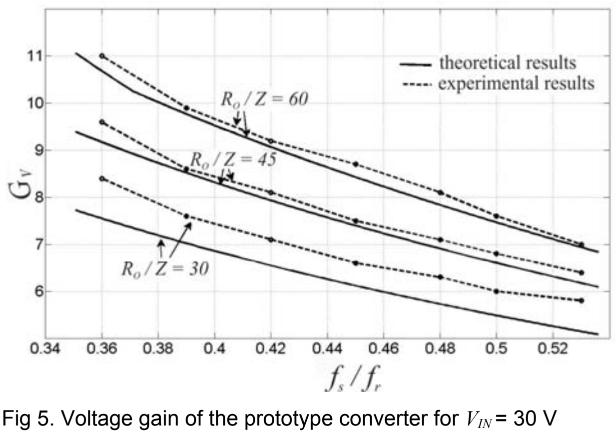

Fig 5. Voltage gain of the prototype converter for V_{IN} = 30 V

The values of resonant elements were selected using the assumed resonance frequency (f_r = 250 kHz) and resonance impedance (Z = 30 Ω) according to the formulas in (25) - (26).

$$C_r = \frac{1}{2\pi f_r Z} = \frac{1}{2\pi \cdot 250000 \cdot 30} \approx 22nF \tag{25}$$

$$L_r = C_r Z = 22 \cdot 10^{-9} \cdot 30^2 \approx 20\mu H \tag{26}$$

Maximum output power is limited by maximum inductor current I_{L1}, maximum inductor current is limited by transistor T voltage V_{DS} in accordance with inequality (27).

$$V_{DSS} > \frac{V_{OUT} + V_{IN}}{N+1} + I_{L1}\sqrt{\frac{L_r}{C_r}} \tag{27}$$

The voltage gain characteristics of the converter obtained experimentally are close to the theoretical values (Fig. 5 - 6). The slight deviations are due to the continuity of inductor current (when determining the characteristics, it was assumed that the transistor is switched at the BCM). This is particularly evident at a high voltage gain and high output power.

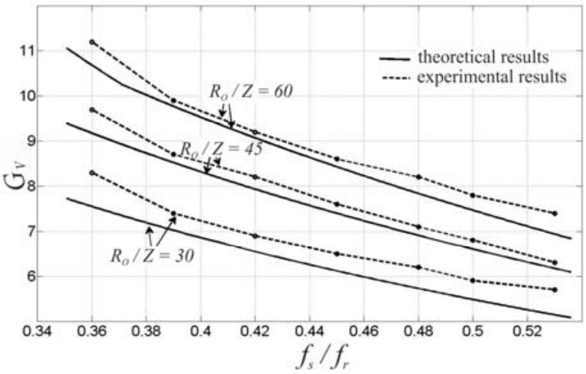

Fig.6. Voltage gain of the prototype converter for V_{IN} = 40 V

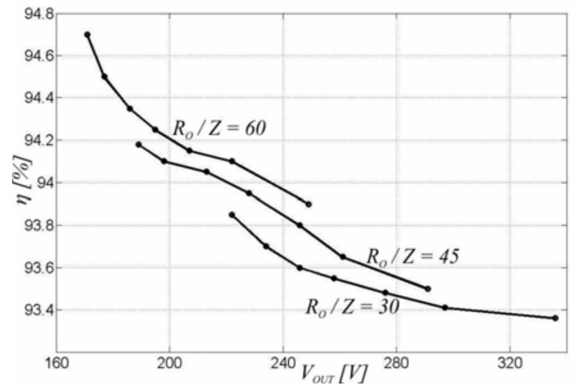

Fig.7. Efficiency as a function of output voltage for V_{IN} = 30 V

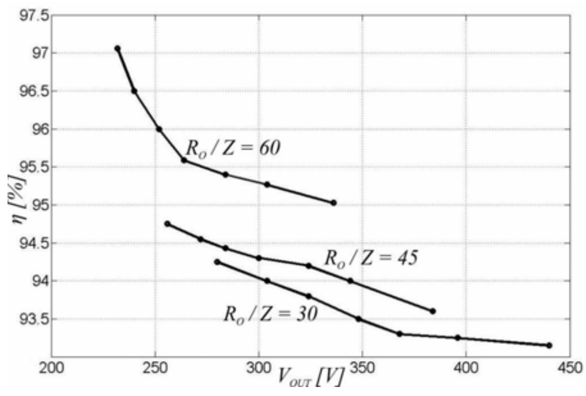

Fig.8. Efficiency as a function of output voltage for V_{IN} = 40 V

Figures 6 and 7 show efficiency as a function of output voltage for three different output resistances and two input voltages. Increased voltage gain is negatively correlated with converter efficiency. The highest efficiency value (97%) was obtained at V_{IN} = 40 V, V_{OUT} = 235 V and R_O = 2 kΩ. Figure 9 shows selected waveforms of converter voltages and currents. In accordance with assumptions, the converter transistor is switched on and off at zero voltage. The output diode current is quasi sinusoidal in nature. Ripple of the inductor L_1 and diode D current when transistor is switched on is due to parasitic resonance resulting from the capacitance of the output diode and the inductance of the tapped inductor.

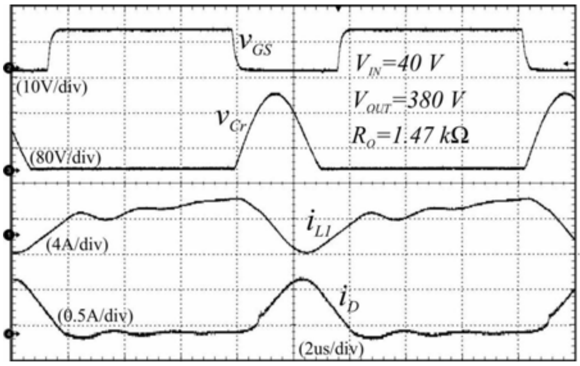

Fig.9. Selected waveforms of converter

Conclusion

In the article has successfully developed a novel quasi-resonant boost converter with tapped inductor. According of the experimental results, the maximum efficiency is 97%, while achieving voltage gain $G_V \approx 5.9$. High voltage gain has been accomplished by using a tapped inductor. The transistor of the converter was switched with a maximum frequency of 125 kHz, simultaneously soft switching of power transistor and output diode has been achieved. Therefore the switching frequency of this converter can be higher, this enables reduction size of the converter components. Soft switching of the transistor also enabled reduction of EMI problem. The converter does not require a snubber circuit, quasi-sinusoidal current changing in tapped inductor prevent overvoltage by leaked inductance. The ability to operate this converter at high frequencies, and consequently its miniaturization, and a small number of components used and also high voltage gain, makes this converter an attractive alternative choice to low power and high step-up converters.

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