Novel Dual DC-DC Flyback Converter with Leakage-Energy Recycling

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Abstract

A novel dual DC-DC flyback converter with leakage-energy recycling is presented in this paper. Only one active switch is used for this converter. A pulse-width-modulation strategy is adopted to control this switch. Two transformers are employed for the proposed converter. During the switch ON-period, the primary windings of the two transformers store energies. At the switch OFF-period, the energies stored in the primary windings of the two transformers are released to the output via the secondary windings of the two transformers. Meanwhile, the leakage energies of the two transformers can be recycled. The operating principles and steady-state analyses of the proposed converter are described in detail. A prototype circuit of the proposed converter is implemented for verifying the performances.

Key words: Flyback converter, Leakage-energy recycling, Pulse width modulation

I. INTRODUCTION

DC power sources are widely utilized in many products, such as communication equipment, medical instruments, and both industrial and commercial devices. It includes nearly all electronics products. Switching mode power converters are applied to the DC power sources. High switching frequencies are used for these converters. Therefore, these converters can provide a high power density, high stability, fast regulation etc. Many topologies, including buck converters [1]-[3], boost converters [4]-[6], buck-boost converters [7], [8], Cuk converters [9], [10] and SEPIC converters [11], [12], have been researched for non-isolated applications. For isolated applications, the forward and flyback converters are used for low-power applications [13]-[18]. In addition, half-bridge, full-bridge and push-pull converters have been utilized for high-power applications [19]-[24]. For isolated and low-power applications, flyback converters are attractive due to their simple structure and low cost. However, the main drawback of flyback converters is the existence of leakage inductance of the transformer. It results in a high voltage spike on the active switch and high power losses. In order to improve this problem, some techniques have been studied. An RCD snubber has been used to limit the voltage spikes on the active switches [25]. However, the energy stored in the leakage inductance of the transformer is dissipated into the resistor of the RCD snubber. A two-switch flyback converter has been used to recycle the leakage-energy of transformers [26], [27]. Nevertheless, this topology need two active switches, which results in higher costs. An active clamp circuit has been applied to eliminate the voltage spike on active switches [28], [29]. In addition, this converter can provide zero voltage switching for the active switch. However, two active switches are employed for this converter. The interleaved flyback converter employs two transformers for high power applications [30]. The leakage energies of the two transformers can be recycled. However, two active switches are employed for this converter as well.

In this paper, a novel single switch dual DC-DC flyback converter with leakage-energy recycling is presented. The circuit configuration of the proposed converter is displayed in Fig. 1. The proposed converter is composed of one switch $S_1$, two transformers $TR_1$ and $TR_2$, three diodes $D_1$, $D_2$ and $D_3$, and three capacitors $C_1$, $C_2$ and $C_0$. The windings in the transformers $TR_1$ and $TR_2$ have same turns, $N_{11} = N_{21}$ and $N_{12} = N_{22}$, and turns ratios $n = N_{12} / N_{11} = N_{22} / N_{21}$. The leakage energies of the transformers $TR_1$ and $TR_2$ can be recycled into the capacitors $C_1$ and $C_2$ via the diode $D_1$ at the switch $S_1$ OFF-period. Therefore, the power losses can be reduced. For
simplifying the circuit analysis of the proposed converter, it is assumed that all of the components are ideal. Therefore, the conducting resistance of the active switch $S_1$, the forward voltage drop of the diodes $D_1$, $D_2$ and $D_3$, the equivalent series resistance (ESR) of the transformers $TR_1$ and $TR_2$ and capacitors $C_1$, $C_2$ and $C_o$ are ignored. In addition, these capacitors are sufficiently large. The voltages across these capacitors are considered constant at each switching period.

II. OPERATING PRINCIPLE

An equivalent circuit of the proposed converter is displayed in Fig. 2. The active switch $S_1$ is controlled by utilizing the pulse-width modulation strategy. The transformer $TR_1$ is modeled as the magnetizing inductance $L_{m11}$, the leakage inductance $L_{k11}$, and an ideal transformer. Similarly, the transformer $TR_2$ is modeled as the magnetizing inductance $L_{m21}$, the leakage inductance $L_{k21}$ and an ideal transformer. Because the windings of the transformers $TR_1$ and $TR_2$ have same turns, the magnetizing-inductance and leakage-inductance of the transformers $TR_1$ and $TR_2$ are assumed as follows:

$$L_{m11} = L_{m21} = L_m$$  \hspace{1cm} (1)
$$L_{k11} = L_{k21} = L_k$$  \hspace{1cm} (2)

The coupling coefficient $k$ of the transformers $TR_1$ and $TR_2$ is given as:

$$k = \frac{L_m}{L_m + L_k}$$  \hspace{1cm} (3)

Some typical waveforms during one switching period in continuous conduction mode (CCM) operation are displayed in Fig. 3. In addition, the equations are assumed as:

$$i_{Lm11} = i_{Lm21} = i_{Lm}$$  \hspace{1cm} (4)
$$v_{Lm11} = v_{Lm21} = v_{Lm}$$  \hspace{1cm} (5)
$$i_{Lk11} = i_{Lk21} = i_{Lk}$$  \hspace{1cm} (6)
$$v_{Lk11} = v_{Lk21} = v_{Lk}$$  \hspace{1cm} (7)
$$V_{e1} = V_{e2} = V_e$$  \hspace{1cm} (8)

The operating principles in CCM operation are described as follows.

Mode I [$t_0$, $t_1$]: The active switch $S_1$ is turned on. The current direction is displayed in Fig. 4(a). The energies stored
in the magnetizing inductances $L_{m1}$ and $L_{m2}$ of the transformers are released to the output capacitor $C_o$ and the load $R$ via ideal transformers and the diodes $D_2$ and $D_3$. The DC-source $V_{in}$ the capacitor $C_1$ and the magnetizing inductance $L_{m2}$ transfer their energies to the leakage inductance $L_{k2}$ in series via the active switch $S_1$. Similarly, the DC-source $V_{in}$ the magnetizing inductance $L_{m1}$ and the capacitor $C_2$ transfer their energies to the leakage inductance $L_{k1}$ in series via the active switch $S_1$. Thus, the magnetizing-inductance currents $i_{lm1}$ and $i_{lm2}$ are decreased and the leakage-inductance currents $i_{lk1}$ and $i_{lk2}$ are increased, as displayed in Fig. 3. This mode ends when the magnetizing-inductance currents $i_{lm1}$ and $i_{lm2}$ are equal to the leakage-inductance currents $i_{lk1}$ and $i_{lk2}$ at the moment $t = t_1$. The voltages across the magnetizing inductances $L_{m1}$ and $L_{m2}$ and the leakage inductances $L_{k1}$ and $L_{k2}$ are found as follows:

$$v_{lm}^1 = -\frac{V_o}{n}$$  \hspace{1cm} (9)

$$v_{lk}^1 = V_{in} + V_o + \frac{V_c}{n}$$  \hspace{1cm} (10)

From the above equations, the currents through the magnetizing inductances $L_{m1}$ and $L_{m2}$ and the leakage inductances $L_{k1}$ and $L_{k2}$ are derived as:

$$i_{lm}^1(t) = -\frac{V_o}{nL_m}(t-t_0) + i_{lm}(t_0)$$  \hspace{1cm} (11)

$$i_{lk}^1(t) = \frac{1}{L_k}(V_{in} + V_o + \frac{V_c}{n})(t-t_0)$$  \hspace{1cm} (12)

Mode II $[t_1, t_2]$: The active switch $S_1$ is still turned on. The current direction is displayed in Fig. 4(b). The DC-source $V_{in}$ and the capacitor $C_1$ are in series to transfer their energies for the magnetizing inductance $L_{m1}$ and the leakage inductance $L_{k2}$ via the active switch $S_1$. Similarly, the DC-source $V_{in}$ and the capacitor $C_2$ are in series to transfer their energies for the magnetizing inductance $L_{m2}$ and the leakage inductance $L_{k1}$ via the active switch $S_1$. The energy stored in the output capacitor $C_o$ is discharged to the load $R$. Therefore, the magnetizing-inductance currents $i_{lm1}$ and $i_{lm2}$, and the leakage-inductance currents $i_{lk1}$ and $i_{lk2}$ are increased, as displayed in Fig. 3. This mode ends when the leakage-inductance currents $i_{lk1}$ and $i_{lk2}$ are equal to zero at this moment $t = t_2$. The following equations are given as:

$$v_{lm}^2 = V_{in} + V_c$$  \hspace{1cm} (13)

$$i_{lm}^2 = i_{lk}^2$$  \hspace{1cm} (14)

Owing to:

$$v_{lk}^2 = \frac{L_k}{L_m}\frac{di_{lm}^2}{dt} = \frac{1-k}{k}L_m\frac{di_{lm}^2}{dt}$$  \hspace{1cm} (15)

The voltage $v_{lk}$ is rewritten as:

$$v_{lk}^2 = \frac{1-k}{k}v_{lm}^2$$  \hspace{1cm} (16)

Substituting (16) into (13) yields:

$$v_{lm}^2 = k(V_o + V_c)$$  \hspace{1cm} (17)

Using (12) and (20), the current $i_{lm}^2$ is derived as:

$$i_{lm}^2(t) = \frac{k(V_o + V_c)}{L_m}(t-t_0) + i_{lm}(t_1)$$  \hspace{1cm} (18)

Mode III $[t_2, t_3]$: The active switch $S_1$ is turned off. The current direction is displayed in Fig. 4(c). The magnetizing inductance $L_{m2}$ and the leakage inductance $L_{k2}$ are in series to release their energies to the capacitor $C_1$ via the diode $D_1$. Similarly, the magnetizing inductance $L_{m1}$ and the leakage inductance $L_{k1}$ are series to release their energies to the capacitor $C_1$ via the diode $D_1$. Thus, the leakage energies have been recycled. The energies stored in the magnetizing inductances $L_{m1}$ and $L_{m2}$ are released to the load $R$ via ideal transformers and the diodes $D_2$ and $D_3$. The energy of the output capacitor $C_o$ is also discharged to the load $R$. Therefore, the magnetizing-inductance currents $i_{lm1}$ and $i_{lm2}$, and the leakage-inductance currents $i_{lk1}$ and $i_{lk2}$ are decreased, as displayed in Fig. 3. This mode ends when the leakage-inductance currents $i_{lk1}$ and $i_{lk2}$ are equal to zero at this moment $t = t_3$. The voltages across the magnetizing inductances $L_{m1}$ and $L_{m2}$, and the leakage inductances $L_{k1}$ and $L_{k2}$ are obtained as follows:

$$v_{lm}^3 = -\frac{V_o}{n}$$  \hspace{1cm} (19)

$$v_{lk}^3 = \frac{V_o}{n} - V_c$$  \hspace{1cm} (20)

Thus:

$$i_{lm}^3(t) = -\frac{V_o}{nL_m}(t-t_0) + i_{lm}(t_1)$$  \hspace{1cm} (21)

$$i_{lk}^3(t) = \frac{1}{L_k}(V_o - V_c)(t-t_0) + i_{lk}(t_2)$$  \hspace{1cm} (22)

Mode IV $[t_3, t_4]$: The active switch $S_1$ is still turned off. The current direction is displayed in Fig. 4(d). The energies stored in the magnetizing inductances $L_{m1}$ and $L_{m2}$ are released to the output capacitor $C_o$ and the load $R$ via ideal transformers and the diodes $D_2$ and $D_3$. Thus, the magnetizing-inductance currents $i_{lm1}$ and $i_{lm2}$ are decreased, as displayed in Fig. 3. This mode ends when the active switch $S_1$ is turned on at the beginning of the next switching period. The voltages across the magnetizing inductances $L_{m1}$ and $L_{m2}$ are found as follows:

$$v_{lm}^4 = \frac{V_o}{n}$$  \hspace{1cm} (23)

In addition:

$$i_{lm}^4(t) = -\frac{V_o}{nL_m}(t-t_0) + i_{lm}(t_3)$$  \hspace{1cm} (24)
III. STEADY-STATE ANALYSIS

A. Voltage Gain

Since the time durations of mode I are very short when compared to a switching period, mode I is neglected in the following analysis. By using the voltage-second balance principle for the magnetizing inductance $L_{m1}$, the following equation is given as:

$$v_{lm}^H D_{t_1} + v_{lm}^H (t_3 - t_2) + v_{lm}^H (t_4 - t_3) = 0$$  \( (25) \)

Substituting (17), (19) and (23) into (25), the following equation can be found.

$$V_{in} c^o = V_{in} c - V_n (1 - D)$$  \( (26) \)

If the leakage inductances of the transformers are neglected, the coupling-coefficient $k$ is equal to 1. In addition, the voltage $V_c$ can be written as:

$$V_c = \frac{V_n}{n}$$  \( (27) \)

Substituting (27) and $k = 1$ into (26), the voltage gain of the proposed converter is found as follows:

$$M = \frac{V_o}{V_{in}} = \frac{I_o}{I_{in}} = \frac{nD}{1 - 2D}$$  \( (28) \)

B. Boundary Operating Condition

For the sake of simplicity, the leakage inductance of the transformer is ignored. When the proposed converter is operated in the boundary conduction mode, the peak value of the magnetizing-inductance current can be found from (18).
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Fig. 6. Boundary condition of the proposed converter at \( n = 0.75 \).

\[
I_{Lmp} = \frac{V_m + V_S}{L_m} DT_s = \frac{nV_m + V_o}{nL_m} DT_s \tag{29}
\]

The average value of the output-capacitor current can be written as:

\[
I_{co} = \frac{I_{Lmp}}{n}(1-D) - I_o \tag{30}
\]

Using the ampere-second balance principle for the output capacitor, it can be seen that the average value of the output-capacitor current \( I_{co} \) is equal to 0. Therefore:

\[
I_{Lmp}(1-D) = I_o = \frac{V_o}{R} \tag{31}
\]

The magnetizing-inductance time constant is defined as:

\[
\tau_{Lm} = \frac{L_m}{RT_s} \tag{32}
\]

Using (28)-(29) and (31)-(32), the boundary magnetizing-inductance time constant \( \tau_{LmB} \) is derived as:

\[
\tau_{LmB} = \left( \frac{1-D}{n} \right)^2 \tag{33}
\]

A curve of the boundary magnetizing-inductance time constant \( \tau_{LmB} \) is plotted in Fig. 6. At \( \tau_{Lm} > \tau_{LmB} \), the proposed converter is operated in the CCM.

C. Voltage Stresses on the Power Devices

From the operating principle analysis, the voltage stresses on the active switch \( S_1 \) and the diodes \( D_1, D_2, D_3 \) are given as:

\[
V_{S1} = V_{D1} = V_m + \frac{2V_o}{n} \tag{34}
\]

\[
V_{D2} = V_{D3} = nV_m + 2V_o \tag{35}
\]

D. Power Losses Analysis

The average of the input current is written as:

\[
I_{in(\text{av})} = \frac{nD}{1-2D} I_o \tag{36}
\]

During the switch \( S_1 \) ON-period, the average and root-mean-square of the switch-current are derived as:

\[
I_{S1(\text{av})} = \frac{n}{1-2D} I_o \tag{37}
\]

\[
I_{S1(\text{rms})} = \frac{n\sqrt{D}}{1-2D} I_o \tag{38}
\]

Thus, the power loss of the switch \( S_1 \) is given as follows:

\[
P_{S1} = I_{S1(\text{rms})}^2 r_{S1} = \frac{n^2 D}{(1-2D)^2} I_o^2 r_{S1} \tag{39}
\]

where \( r_{S1} \) is the ON-state resistance of the switch \( S_1 \). The average of the capacitor-current are given as:

\[
I_{c1(\text{av})} = I_{c2(\text{av})} = \left\{ \begin{array}{ll}
\frac{I_{S1(\text{av})}}{2}, & S_1 \text{ ON} \\
\frac{D}{2(1-D)} I_{S1(\text{av})}, & S_1 \text{ OFF}
\end{array} \right. \tag{40}
\]

\[
I_{co(\text{av})} = \left\{ \begin{array}{ll}
-I_o, & S_1 \text{ ON} \\
\frac{D}{1-D} I_o, & S_1 \text{ OFF}
\end{array} \right. \tag{41}
\]

The root-mean-square of the capacitor-current is derived as:

\[
I_{c1(\text{rms})} = I_{c2(\text{rms})} = \left( \frac{D}{2(1-D)} \right)^2 I_{S1(\text{rms})}^2 r_c \tag{42}
\]

\[
I_{co(\text{rms})} = \frac{D}{1-D} I_o r_{co} \tag{43}
\]

Therefore, the power losses of the capacitors are written as:

\[
P_{c1} = P_{c2} = I_{c1(\text{rms})}^2 r_c = \frac{n^2 D}{4(1-D)(1-2D)^2} I_o^2 r_c \tag{44}
\]

\[
P_{co} = I_{co(\text{rms})}^2 r_{co} = \frac{D}{1-D} I_o^2 r_{co} \tag{45}
\]

where \( r_{c1} \) and \( r_{co} \) are the ESRs of the capacitors \( C_1 \) and \( C_o \). During the switch \( S_1 \) OFF-period, the average of the diode-current is derived as:

\[
I_{D1(\text{av})} = \frac{D}{1-D} I_{S1(\text{av})} = \frac{nD}{(1-D)(1-2D)} I_o \tag{46}
\]

\[
I_{D2(\text{av})} = I_{D3(\text{av})} = \frac{1}{2(1-D)} I_o \tag{47}
\]

\[
I_{D2(\text{rms})} = I_{D3(\text{rms})} = \frac{1}{2\sqrt{1-D}} I_o \tag{48}
\]

Thus, the power losses of the diodes \( D_1-D_3 \) in the ON-state forward voltage-drop are found as follows:

\[
P_{D1} = P_{F11} I_{D1(\text{av})} = \frac{nD V_{F11}}{1-D} I_o \tag{49}
\]

\[
P_{D2} = P_{F12} I_{D2(\text{av})} = \frac{V_{F22}}{2(1-D)} I_o \tag{50}
\]
where $V_{FD1}$ and $V_{FD2}$ are the ON-state forward voltage-drop of the diodes $D_1$ and $D_2$. The average of the leakage-inductor currents of the transformers $TR_1$ and $TR_2$ are given as:

$$I_{Lk1} = I_{Lk2} = \frac{I_{S1(on)}}{2D} = I_{S1(on)} S_1 \text{ ON}$$  \hspace{1cm} (51)

The root-mean-square of the leakage-inductor currents of $TR_1$ and $TR_2$ are derived as:

$$I_{Lk rms1} = I_{Lk rms2} = \frac{mI_o}{2(1-2D)}$$  \hspace{1cm} (52)

Thus, the power losses of the primary-winding of $TR_1$ and $TR_2$ are given as:

$$P_{11} = P_{21} = I_{Lk1}^2 r_{11} \frac{n^2 D}{(1-2D)(1-D)}$$  \hspace{1cm} (53)

where $r_{11}$ is the ESR of the primary winding of $TR_1$. The power losses of the secondary-winding of $TR_1$ and $TR_2$ are obtained as:

$$P_{12} = P_{22} = I_{Lk2}^2 r_{12} \frac{I_o r_{12}}{4(1-D)}$$  \hspace{1cm} (54)

where $r_{12}$ is the ESR of the secondary winding of $TR_1$.

IV. EXPERIMENTAL RESULTS

A prototype circuit has been built in the laboratory to verify the feasibility of the proposed converter. The circuit specifications and components are selected as input voltage $V_{in} = 100 \text{ V}$, output voltage $V_o = 48 \text{ V}$, output power $P_o = 250 \text{ W}$, switching frequency $f_s = 75 \text{ kHz}$, capacitors $C_1 = C_2 = 100 \mu \text{ F}$ and $C_o = 470 \mu \text{ F}$, turn-ratio of the transformers $n = 0.75$, switch $S_1$ IXTQ52N30P, and diodes $D_1$, $D_2$ and $D_3$: STTH6003CW. Thus, the voltage gain $M$ is equal to 0.48. Substituting the voltage gain $M = 0.48$ and the turns ratio $n = 0.75$ into (28), the duty ratio $D$ is found to be 0.28. Substituting the turns ratio $n = 0.75$ and the duty ratio $D = 0.28$ into (33), the boundary magnetizing-inductance time constant $\tau_{LmB}$ is derived as 0.92. It is assumed that the proposed converter is operated in the CCM from 40% of the full load. Thus, the load $R$ is 23 $\Omega$. At $\tau_{Lm} > \tau_{LmB}$, the proposed converter is operated in the CCM. Therefore:

$$L_m > \tau_{LmB} R T = 0.92 \times 23 \times 13.3 \mu \text{ H} = 281 \mu \text{ H}$$  \hspace{1cm} (55)

The magnetizing-inductance $L_m$ is selected as 285 $\mu \text{ H}$.

Under input voltage $V_{in} = 100 \text{ V}$, output voltage $V_o = 48 \text{ V}$ and output power $P_o = 250 \text{ W}$, some experimental waveforms are shown in Fig. 7. Fig. 7(a) shows waveforms of $i_{Lk1}$, $i_{Lk2}$ and $i_{D1}$. It can be seen that the waveforms $i_{Lk1}$ and $i_{Lk2}$ are almost same. In addition, the summation of the currents $i_{Lk1}$
and \(i_{L21}\) is equal to the current \(i_{D1}\) during the switch \(S_1\) OFF-period. Thus, it can be ensured that the leakage energies of the transformers can be recycled to the capacitors \(C_1\) and \(C_2\) via the diode \(D_1\). Fig. 7(b) shows the waveforms \(i_{S1}\), \(i_{S12}\) and \(i_{S22}\). It can be seen that the summation of the currents \(i_{L11}\) and \(i_{L21}\) is equal to the current \(i_{S1}\) during the switch \(S_1\) ON-period. Fig. 7(c) shows waveforms of \(V_{S1}, V_{D1}\) and \(V_{D2}\), which agree with the operating principle and steady-state analysis. Fig. 7(d) shows waveforms of \(V_{S1}, V_{C2}\) and \(V_o\) in the start-up process. It can be seen that the voltage \(V_o\) is controlled at the setting value. Fig. 8 shows experimental waveforms under \(V_m = 100\) V, \(V_o = 48\) V and \(P_o = 70\) W. The load \(R\) and the magnetizing-inductance time constant \(\tau_{Lm}\) are 33 \(\Omega\) and 0.648, respectively. Thus, \(\tau_{Lm}\) is less than \(\tau_{LmB}\). It can be seen that the proposed converter is operated in the discontinuous conduction mode. Fig. 9 shows the dynamic response of the proposed converter for a load change between 70W and 250W. Fig. 10 shows the measured efficiency of this prototype circuit. The maximum measured efficiency is 94.8% and the measured efficiency is 93.1% under the full-load condition.

V. CONCLUSIONS

The conventional DC-DC flyback converter has the merits of a simple structure and low cost. However, this converter also possesses leakage inductance of the transformer. This results in a lower efficiency. In this paper, only a single switch is used in the proposed converter. This proposed converter employs two transformers with the same inductance. During the switch OFF-period, the energies of the magnetizing-inductance of the transformers are released to the output. Meanwhile, the leakage energies of the transformers can be recycled. From the obtained experimental results, it can be seen that the leakage energies of the transformers have been recycled. In addition, the measured efficiency is 93.1% under the full-load condition and the maximum efficiency is 94.8%.

REFERENCES


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