A New High Efficiency PWM Single-Switch Isolated Converter

Ki-Bum Park, Chong-Eun Kim, Gun-Woo Moon, and Myung-Joong Youn

Department of Electrical Engineering and Computer Science, Korea Advanced Institute of Science and Technology, 373-1 Gueseong-dong, Yuseong-Gu, Daejeon, Republic of Korea

Abstract

The flyback converter is one of the most attractive isolated converters in small power applications because of its simple structure. However, it suffers from high device stress, large transformer size, and high voltage stress across switch and diode. To solve these problems a new cost-effective PWM single-switch isolated converter is proposed. The proposed converter has no output filter inductor, reduced voltage stress on the secondary devices, and reduced transformer size. Moreover, the switch turn-off loss is reduced and no dissipative snubber across the secondary diode is required. Therefore, it features a simple structure, low cost, and high efficiency. The operational principle and characteristics of proposed converter are presented compared with flyback converter and verified experimentally

1. Introduction

Till now, the various types of isolated switching mode power converters have been proposed [1-6]. Among them, the flyback converter shown in Fig. 1 is a very favorite topology for its simple circuit configuration and an easy isolation compared with other topologies in low power applications [4-6]. That is, using only one switch, no output filter inductor and no additional transformer reset circuit make it very attractive. However, it suffers from the high voltage/current stress of devices and the large magnetizing current of transformer increases the transformer size. Moreover, the high primary peak current causes high switch turn-off loss and the leakage inductance of transformer causes high voltage spike and ringing across the switch and diode at the switching transitions, which requires snubbers.

To improve abovementioned drawbacks, a new cost-effective PWM single-switch isolated converter is proposed in this paper. As can be seen in Fig. 2, the proposed converter simply consists of switch Q, transformer T, capacitor C_S, diodes D_{S1} and D_{S2}, and auxiliary snubber network. In the proposed converter, the small transformer leakage inductor L_{lkg} drives the powering current, therefore no large filter inductor is required. The reset of transformer is automatically achieved by the secondary capacitor C_S and the offset magnetizing current of transformer is less than that of flyback converter resulting in the smaller transformer size. Moreover, the switch turn-off loss can be reduced by controlling the resonance between L_{lkg} and C_S. Furthermore, the voltage stresses of secondary diodes D_{S1} and D_{S2} are clamped to V_O, therefore it has basically less voltage stress and no snubber is needed.

2. Operational Principle

The key waveforms and topological states of proposed converter are presented in Figs. 3 and 4, respectively. The operation of one switching period is subdivided into two modes as follows. Before t_0, the transformer magnetizing inductor current I_{Lm} freewheels through the secondary capacitor C_S.

Mode 1 [$t_0 \sim t_1$] : After Q is turned on at t_0, the powering path from the input to the output is formed through L_{lkg}, C_S, D_{S1}, and Q. Therefore, L_{lkg} drives the powering current and resonated with C_S.

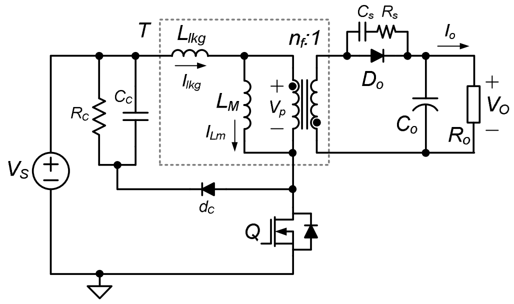


Figure 1. Flyback converter

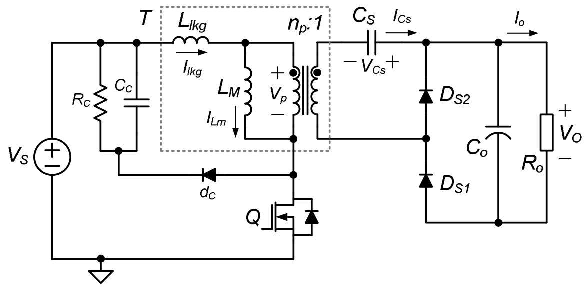


Figure 2. Proposed converter.

The primary current $I_{lkg}(t)$ contains this powering current and $I_{Lm}(t)$ as well. $I_{lkg}(t)$ and $I_{Lm}(t)$ are expressed as follows.

$$I_{lkg}(t) = \left(V_S + V_{Cs}(t_0) - nV_O\right)\frac{1}{Z_R}\sin\left(\omega_r(t-t_0)\right) + I_{Lm}(t-t_0),$$
$$Z_R = n_p\sqrt{\frac{L_{lkg}}{C_S}}, \qquad \omega_r = \frac{n_p}{\sqrt{L_{lkg}C_S}} \tag{1}$$

$$I_{Lm}(t) = I_{Lm}(t_0)\cos\left(\omega_m(t-t_0)\right) + \left(n_pV_O - n_pV_{Cs}(t_0)\right)\frac{1}{Z_M}\sin\left(\omega_m(t-t_0)\right),$$
$$Z_M = n_p\sqrt{\frac{L_M}{C_S}}, \qquad \omega_m = \frac{n_p}{\sqrt{L_MC_S}} \tag{2}$$

Provided that the ripple of $V_{Cs}(t)$ is small enough $I_{Lm}(t)$ can be approximated as (3), where V_{Cs_avg} means the average value of $V_{Cs}(t)$.

$$I_{Lm}(t) \approx I_{Lm}(t_0) + \frac{\left(n_pV_O - n_pV_{Cs\_avg}\right)(t-t_0)}{L_M} \tag{3}$$

The real powering current which transfers the power from the input to the output is the difference between $I_{lkg}(t)$ and $I_{Lm}(t)$. This current flows to the output through C_S and D_{S1} and expressed as follow.

$$I_{Ds1}(t) = n_p\left(V_S + V_{Cs}(t_0) - n_pV_O\right)\frac{1}{Z_R}\sin\left(\omega_r(t-t_0)\right) \tag{4}$$

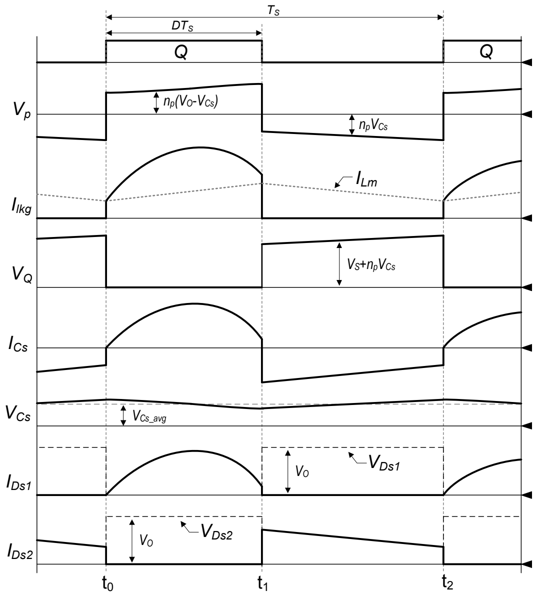


Figure 3. Key waveforms of proposed converter.

Mode 2 [$t_1 \sim t_2$] : After Q is turned off at t_1, $I_{lkg}(t)$ is decreased to zero rapidly since the primary current path is blocked. Therefore, $I_{Lm}(t)$ which used to flow in the primary flows though C_S and D_{S2} of the secondary. Hence, $nV_{CS}(t)$ is applied to L_m reversely and the reset of transformer is achieved.

$I_{Lm}(t)$ is expressed as (5). Provided that the ripple of $V_{CS}(t)$ is small enough $I_{Lm}(t)$ can be approximated as (6).

$$I_{Lm}(t)=I_{Lm}(t_1)\cos\left(\omega_m(t-t_1)\right)+\frac{n_pV_{Cs}(t_1)}{Z_M}\sin\left(\omega_m(t-t_1)\right) \tag{5}$$

$$I_{Lm}(t)\cong I_{Lm}(t_1)-\frac{n_pV_{Cs\_avg}(t-t_1)}{L_M}. \tag{6}$$

In this mode, the voltage spike of $V_Q(t)$ at the turn-off instant by L_{lkg}'s energy is not considered. At t_2, one period is completed and the same operation is repeated.

3. Characteristics

3.1 DC conversion ratio

The voltage-second balance across L_M can be expressed as (7) using V_{Cs_avg}, then the relationship between V_{Cs_avg} and V_O is expressed as (8).

$$D\left(n_pV_O-n_pV_{Cs\_avg}\right)=(1-D)n_pV_{Cs\_avg} \tag{7}$$

$$V_{Cs\_avg}=DV_O \tag{8}$$

In mode 1 the average current flows through C_S is the same as the load current I_O. Therefore, the voltage ripple of $V_{CS}(t)$, ΔV_{CS} can be expressed as (9).

$$\Delta V_{Cs}=\frac{I_OT_S}{C_S}=\frac{V_OT_S}{R_OC_S} \tag{9}$$

Since the averaging current of I_{Ds1} is equal to I_O, it can be expressed as (10). $V_{CS}(t_0)$ is expressed as (11) from (8) and (9).

$$I_O=\frac{V_O}{R_O}=\frac{C_S}{n_pT_S}\left(V_S+n_pV_{Cs}(t_0)-n_pV_O\right)\left(1-\cos\left(\omega_rDT_S\right)\right) \tag{10}$$

$$V_{Cs}(t_0)=V_{Cs\_avg}+\frac{\Delta V_{Cs}}{2}=DV_O+\frac{V_OT_S}{2R_OC_S} \tag{11}$$

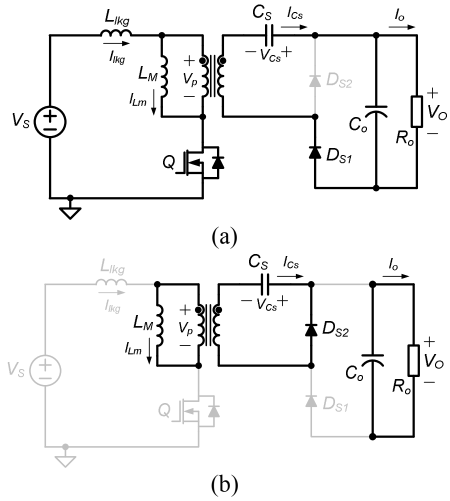


Figure 4. Topological states of operational mode. (a) Mode 1 (t_0~t_1). (b) Mode 2 (t_1~t_2)

By substituting (11) to (10), the input-output voltage conversion ratio is obtained as follow.

$$\frac{V_O}{V_S}=\frac{1}{n_p\left[A+1-D\right]},\quad A=\frac{T_S}{R_OC_S}\left(\frac{1}{2}-\frac{1}{1-\cos\left(\omega_rDT_S\right)}\right) \tag{12}$$

If the value of '*A*' is small enough, (12) can be approximated as (13) like the boost conversion ratio and V_{Cs_avg} also can be approximated as (14). In other words, provided that the voltage applied to L_{lkg} is quite small to be ignored, the voltage applied to L_M would be similar to that of flyback converter. Therefore, V_{CS} can be expressed as (14). As a similar way, since V_S, V_{CS}, and V_O are connected in series when switch conducting state, V_O can be approximated as (13).

$$\frac{V_O}{V_S}\cong\frac{1}{n_p(1-D)} \tag{13}$$

$$V_{Cs\_avg}\cong\frac{DV_S}{n_p(1-D)} \tag{14}$$

Since the reset action of L_M is similar to flyback converter, it also can be operated in the discontinuous conduction mode (DCM) at the light load, i.e., I_{DS2} can be decreased to zero during the switch-off state. In DCM, the DC-conversion ratio of proposed converter can be approximated as follow.

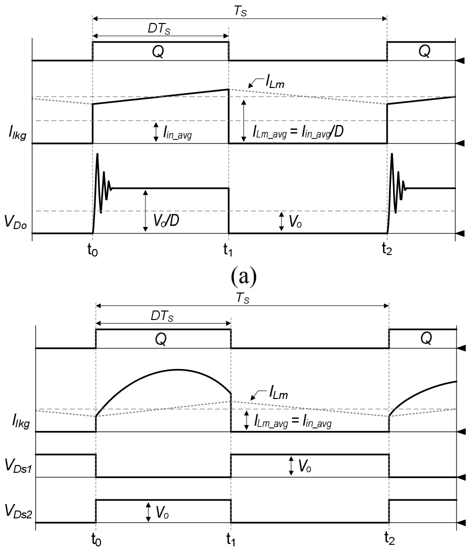
$$\frac{V_O}{V_S}\cong\frac{1+\sqrt{1+2n_p^{\,2}D^2T_SR_O/L_M}}{2n_p} \tag{15}$$

As presented in (13) the DC-conversion ratio of proposed converter in continuous conduction mode (CCM) is mainly dependent on the duty, not on the load. Therefore, the duty is rarely changed under the load variation. On the other hands, as shown in (15), the DC-conversion ratio in DCM is strongly affected by the load as well. Thus, the load variation in DCM rather changes the operating duty as other converters.

3.2 Transformer

In the proposed converter, the reset of transformer is automatically achieved by V_{CS} without an auxiliary circuitry like flyback converter. However, this reset operation causes additional conduction loss in the secondary since the reflected magnetizing current flows through D_{S2}.

In general, a transformer size is considerably influenced by the



(a)

(b)

Figure 5. Comparison between (a) flybcak converter and (b) proposed converter

offset of transformer magnetizing current. That is, a large offset current increases a transformer size [6,7]. Therefore, the transformer size of flyback converter is inevitably large and it is one of the main drawbacks of flyback converter that limits its rated power. Fig. 5 presents the transformer primary current I_{lkg} and magnetizing current I_{Lm} of proposed converter and flyback converter, where I_{in_avg}, means the average of input current. In the proposed converter, the average current of magnetizing current I_{Lm_avg} is the same as I_{in_avg}. On the other hand, I_{Lm_avg} is equal to I_{in_avg}/D in flyback converter. That is, the proposed converter has a less offset of magnetizing current, which can result in smaller transformer size. However, if both converters are designed to be operated in DCM, the transformer size would be similar.

3.3 Stress of devices

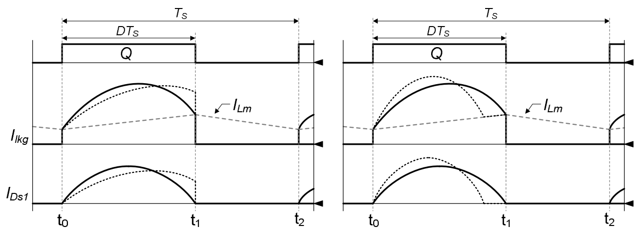
Using (14) the switch voltage stress of proposed converter can be approximated as $V_S/(1\text{-}D)$ and is the same as that of flyback converter. In both converter, a RCD-snubber is required to prevent the voltage overshoot and ringing caused by L_{lkg} as can be seen in Figs. 1 and 2. Fig. 5 shows the voltage waveform of secondary diode. In flyback converter, the voltage stress of secondary diode is V_O/D and a snubber is required to damp the ringing caused by L_{lkg}. On the other hand, in the proposed converter the voltage stress of secondary diodes D_{S1} and D_{S2} are clamped to V_O since two diodes are connected in series across the output V_O. Therefore, it has basically less voltage stress than flyback converter, moreover no snubber is required

The current stresses of proposed converter are basically higher than those of flyback converter because of the resonance between L_{lkg} and C_S. By the current-second balance of C_S, the average of I_{Ds2} is equal to the average of I_{Ds1}. I_{Ds2} is the reflected magnetizing current, thus its peak current is dependent on the magnetizing inductance. The peak current of I_{Ds2} is similar to that of flyback converter.

3.4 Switch turn-off loss and snubber loss

The switch turn-off loss is mainly determined by the switch current at the instant of turn-off. As can be seen in Fig. 5, the switch current of proposed converter at the instant of turn-off is less than that of flyback converter by the help of resonance between L_{lkg} and C_S. Therefore, the proposed converter has less turn-off loss.

In the case of using RCD-snubber in the primary as shown in Figs. 1 and 2, the loss dissipated by the snubber is dependent on the



(a) (b)

Figure 6. Comparison of current waveform.
(a) $T_R/2 > DT_S$. (b) $T_R/2 < DT_S$

energy stored in L_{lkg} [8]. That is, I_{lkg} at the instant of turn-off mainly determines the loss. Therefore, the proposed converter has also less dissipation by the snubber compared with flyback converter.

4. Design Consideration

4.1 Selecting duty and turn ratio

The primary device stress of proposed converter is similar to that of flyback converter since it also utilizes only one switch. Generally, single-switch type converters such as flyback or forward converters suffer from high voltage stress to reset the transformer as its duty increases. On the other hands, a smaller duty increases the current stress of switch. Therefore, the duty and transformer turn ratio of proposed converter are chosen to accommodate as low voltage rating for the switch as possible while having a reasonable current stress of switch using an adequate trade-off.

4.2 Selecting resonant capacitor C_S

Fig. 6 shows the current waveform of proposed converter, where T_R is the resonant period between L_{lkg} and C_S, i.e., $T_R = 1/2\pi\omega_r$. The proper selection of T_R can improve the converter performance. In the case of that $T_R/2>DT_S$, the current stress of device can be reduced, however the switch turn-off loss is increased. On the other hands, in the case of that $T_R/2<DT_S$ the switch turn-off loss is reduced and D_{S1} achieves zero-current-switching (ZCS) turn-off which minimize the reverse recovery of diode. However, the current stress and conduction loss of devices are rather increased. Therefore, T_R is selected in the midpoint as (16) to achieve ZCS of diode while minimizing the switch turn-off loss as presented in the solid line of Fig. 6.

$$\frac{T_R}{2} = DT_S \tag{16}$$

To minimize the switch turn-off snubber loss, the smaller L_{lkg} is desirable. Thus, once L_{lkg} is determined from the fabricated transformer, L_{lkg} is set as it is and C_S can be selected as follow.

$$C_S = \frac{4\left(\pi n_p DT_S\right)^2}{L_{lkg}} \tag{17}$$

The condition (16) can be maintained as long as the converter is operated in CCM. However, at light load where the converter is operated in DCM, the duty is decreased and the condition (16) cannot be guaranteed any more.

4.3 Selecting magnetizing inductance L_M

In the proposed converter, the effect of L_M is similar to that of flyback converter. The large inductance of L_M resulting in a small current ripple of I_{Lm} can reduce the switch turn-off loss and the current stress of D_{S2}. Moreover, the large inductance enlarges a CCM range which guarantees the condition (16). However it increases the transformer size. Therefore, the reasonable trade-off between these factors is required.

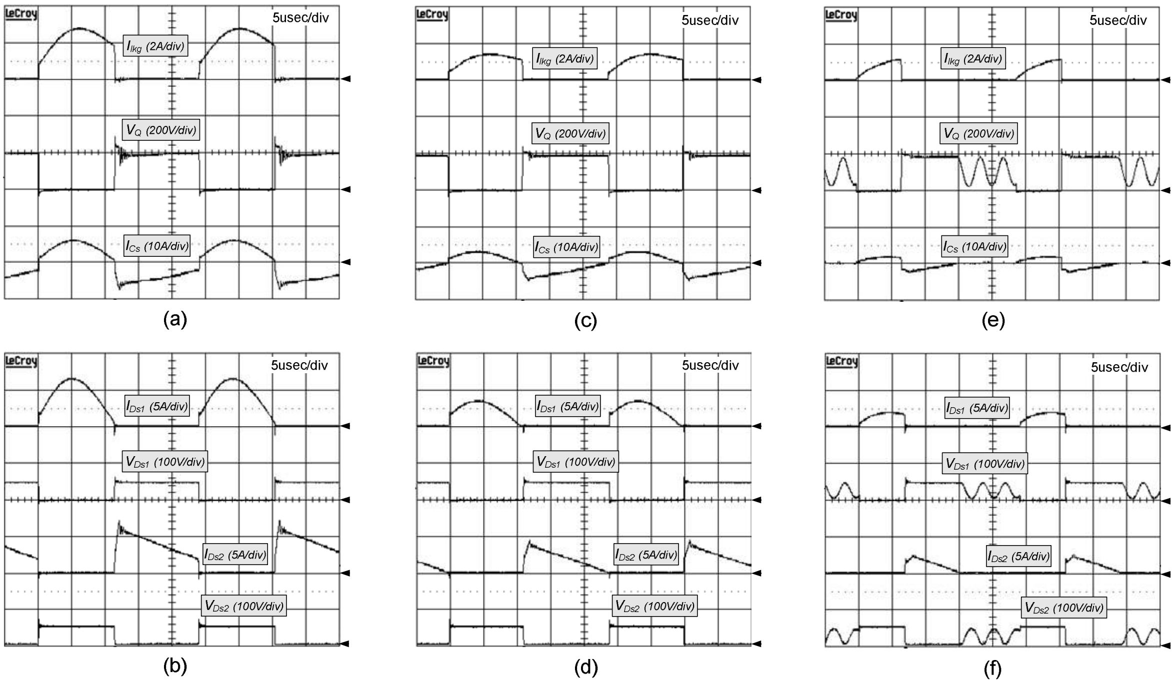


Fig. 7. Experimental waveforms. (a) I_{lkg}, V_Q, and I_{Cs} at full load. (b) I_{Ds1}, V_{Ds1}, I_{Ds2}, and V_{Ds2} at full load. (c) I_{lkg}, V_Q, and I_{Cs} at 50% load. (d) I_{Ds1}, V_{Ds1}, I_{Ds2}, and V_{Ds2} at 50% load. (e) I_{lkg}, V_Q, and I_{Cs} at 20% load. (f) I_{Ds1}, V_{Ds1}, I_{Ds2}, and V_{Ds2} at 20% load.

5. Experimental Results

The prototype of proposed converter has been built with following specifications; input voltage V_S = 100V, output voltage V_O = 48V, output power P_O = 100W, switching frequency f_S = 42kHz, switch Q: FQP17N40, secondary diodes D_{S1} and D_{S2}: 30CTQ060, transformer T: EER3435, turn ratio n_p=56/15, transformer leakage inductance L_{lkg}: 20uH, transformer magnetizing inductance L_m:950uH, and resonant capacitor C_S: 11uF, where a RCD-snubber (R_C=10kΩ)is employed across Q.

Figs. 7 (a) and (b) show the key experimental waveforms of proposed converter at full load condition. It can be seen that I_{lkg} drives the powering current with the resonance between L_{lkg} and C_S, and it decreases I_{lkg} at the instant of switch turn-off. Moreover, it turns-off the secondary diode D_{S1} smoothly, which minimize the reverse recovery. The voltage stresses of D_{S1} and D_{S2} are effectively limited by V_O without additional RC-snubbers.

Figs. 7(c) and (d) show the experimental waveforms at the 50% load condition, and Figs. 7(e) and (f) do at the 20% load condition. In the 50% load condition, it is still operated in the CCM, therefore the duty is rarely changed and ZCS turn-off of D_{S1} is still achieved. On the other hands, in the 20% load condition, it is operated under the DCM. Thus, the duty is considerably reduced, which increase I_{lkg} at the turn-off instant and lose ZCS turn-off condition of D_{S1}. Instead of that, D_{S2} is turned off smoothly.

Fig. 8 shows the measured efficiency of proposed converter and flyback converter. Since the proposed converter has the reduced turn-off loss, reduced RCD-snubber loss and no dissipative snubber in the secondary, it can achieve a higher efficiency than flyback converter along a wide load range.

6. Conclusion

A new single switch isolated converter is proposed, which utilizes the transformer leakage inductor to drive the powering current instead of a large inductor. The proposed converter has the reduced transformer size, reduced switch turn-off loss, and reduced voltage stress on the secondary diodes compared with flyback converter. Moreover, no dissipative snubber across the secondary diodes and no additional transformer reset circuit are required. Therefore, it features a simple structure, low cost, and high efficiency promising for small power applications.

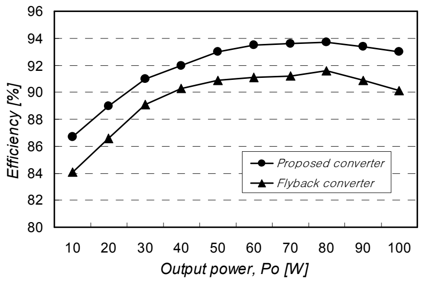


Figure 8. Measured efficiency.

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