CROSS REGULATION IN FLYBACK CONVERTRERS: SOLUTIONS*

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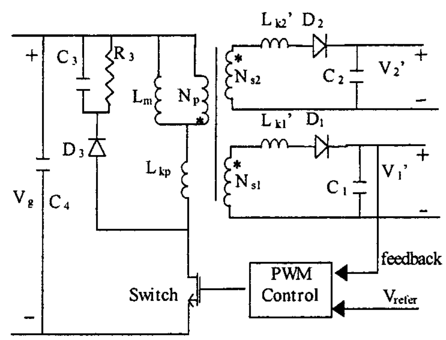
Abstract—Studies at UCI have shown that cross regulation in multiple output flyback converters can be significantly improved when the clamp voltage is lowered to slightly above the reflected output voltage. In this paper, three solutions for reducing the clamp voltage are investigated. In a conventional RC clamp, the clamp voltage can be reduced by decreasing its resistor value, but this leads to higher power losses. A new operating condition for the nondissipative LC snubber is found to achieve the goal by setting its resonant frequency below the switching frequency. In addition, a simple passive energy regenerative clamp is proposed that allows the clamp voltage to be lower than that of the RC clamp, thus improving the cross regulation without using an extra inductor as in the nondissipative LC snubber.

I. INTRODUCTION

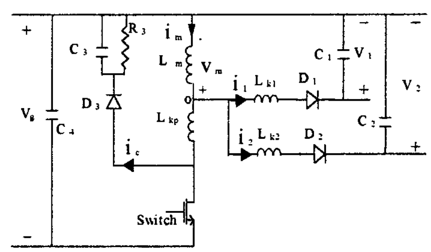
Both the analytical model and experimental results presented in article [1] show that cross regulation in multiple output flyback converters is greatly improved when the clamp voltage is maintained slightly above the reflected output voltage. However, using a traditional RC clamp to keep this voltage low results in higher power losses.

The objective of this paper is to explore lossless clamps that can lower the clamp voltage to improve cross regulation, while still recovering the stored energy in the leakage inductance of the primary winding. The problems of the RC clamp is discussed in Section II, a new operating condition for a nondissipative LC snubber is suggested in Section III, and the proposed passive energy regenerative clamp is given in Section IV. Finally, a comparison of the clamps is summarized in Section V.

II. RC CLAMP



(a)



(b)

Fig. 1. (a). A typical flyback converter of two outputs, (b) Equivalent transformer model on the primary side for analyzing the loss of RC clamp.

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Figure 1 (a) shows a typical two-output flyback converter with an RC clamp (C_3, R_3, and D_3). The transformer in Fig. 1 (a) is modeled as an ideal transformer with a magnetizing inductance, L_m, and equivalent leakage inductance, L_{kp}, L_{k1} and L_{k2}, corresponding to that of the primary winding, secondary winding1 and secondary winding2 respectively. Compared with the leakage inductance, the effect of the winding resistance is small enough to be neglected. To simplify the analysis, the ideal transformer is then removed by converting all of the leakage inductances to the primary side as shown in Fig. 1 (b). When the switch is turned off, all energy in the leakage inductance, L_{kp}, and part of the energy in the magnetizing inductance, L_m, are transferred into capacitor C_3 while the magnetizing current is commutating to the secondary.

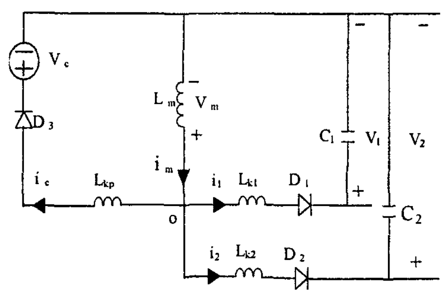

Fig. 2. Equivalent circuit charging clamp capacitor

After the switch is turned off and before the current in the leakage inductor, L_{kp}, drops to zero, the equivalent circuit is shown in Fig. 2. The clamp capacitor C_3 is replaced by voltage source V_c, assuming that the capacitor is large enough that its voltage is constant during the switching cycle and the voltage drops in all diodes are neglected.

The sum of currents at node O equals zero:

$$i_m = i_c + i_1 + i_2 \qquad (1)$$

Differentiating of (1) yields

$$\frac{di_m}{dt} = \frac{di_c}{dt} + \frac{di_1}{dt} + \frac{di_2}{dt} \qquad (2)$$

Substitution of the current changing rates of i_m, i_c, i_1 and i_2 into (2) gives

$$-\frac{V_m}{L_m} = \frac{V_m - V_c}{L_{KP}} + \frac{V_m - V_1}{L_{K1}} + \frac{V_m - V_2}{L_{K2}} \qquad (3)$$

Let $K_1=L_m/L_{k1}$, $K_2=L_m/L_{k2}$ and $K_p=L_m/L_{kp}$, solving (3) for V_m yields

$$V_m = \frac{K_1V_1 + K_2V_2 + K_pV_c}{1 + K_1 + K_2 + K_p} \qquad (4)$$

The current flowing into C_3 is

$$i_c(t) = \frac{V_m - V_c}{L_{kp}} \cdot t + I_o \qquad (5)$$

where I_0 is magnetizing current at the moment when the switch is turned off.

The diode D_3 conduction time, T, is obtained by letting $i_c(t)=0$

$$T = \frac{L_mI_o}{K_p} \cdot \frac{1 + K_1 + K_2 + K_p}{V_c + V_cK_1 + V_cK_2 - K_1V_1 - K_2V_2} \qquad (6)$$

The loss in the RC clamp is

$$P_c = f \cdot \int_0^T V_c \cdot i_c(t) \cdot dt =$$

$$\frac{1}{2} \cdot L_m \cdot I_0^2 \cdot \frac{(1 + K_1 + K_2 + K_p) \cdot f}{\left[1 + K_1 \cdot (1 - \frac{V_1}{V_c}) + K_2 \cdot (1 - \frac{V_2}{V_c})\right] \cdot K_p} \qquad (7)$$

Equation (7) shows that lowering the clamp voltage leads to more losses, especially when the clamp voltage approaches the reflected output voltage V_{ref}. The energy charged into the clamp capacitor is plotted in Fig. 3 in terms of the clamp voltage. From this figure, it is clear that better cross regulation is achieved by decreasing the designed RC clamp voltage is at the cost of the efficiency. It is necessary to make a tradeoff between the cross regulation and efficiency if an RC clamp is used.

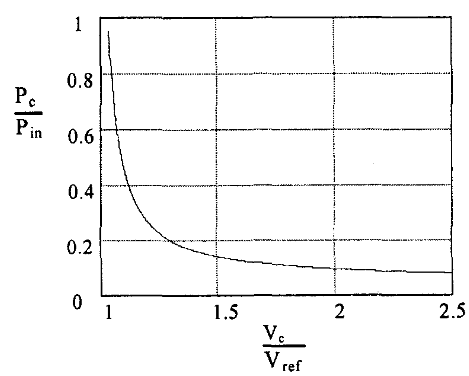

Fig. 3. Loss in RC clamp increases when the clamp voltage approaches reflected output voltage

Equation (7) illustrates several important characteristics of the RC clamp circuit. For a certain resistance of R_3 in the RC clamp, the clamp voltage is determined by the energy into the capacitor. Since it is proportional to the converted overall power, $\frac{1}{2}*L_m*I_0^2$, the cross regulation will vary with load current. However, the lost energy is independent of the input voltage. Therefore, the cross regulation of flyback converter with an RC clamp remains the same for the universal input voltage range if the feedback control has an infinite gain. It is the simplest and cheapest approach to limit the voltage stress cross the switch, but efficiency suffers.

III. NONDISSIPATIVE LC SNUBBER

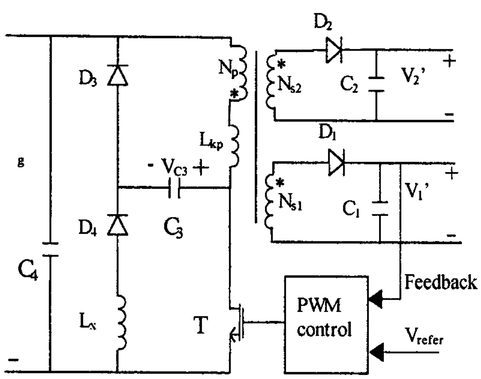

Fig. 4. Nondissipative LC clamp

Figure 4 shows a flyback converter with the nondissipative LC turn-off snubber that is proposed in [2-3]. It comprises of C_3, D_4, L_x and D_3. The transformer primary winding has a leakage inductance of L_{kp}. In terms of the mount of energy stored in the leakage inductance of the primary winding and the input voltage level, the authors reported four operation modes in which the oscillation frequency between C_3 and L_x is assumed to be much higher than switching frequency. This assumption results in reversely changing the polarity of clamp capacitor in the intervals of switch turn-on and turn-off and creating high voltage on the clamp capacitor. In general when the switch is turned off, the voltage cross the clamp capacitor C_3 will be charged up to:

$$V_{C3} = V_{ref} + I_0 \cdot \sqrt{L_{kp} / C_3} \quad (8)$$

where I_0 is the current in the primary winding at the moment when the switch is turned off, V_{ref} is the reflected output voltage, and the magnetizing inductance is assumed to be much larger than the leakage inductance, L_{kp}.

After the switch is turned on, the capacitor, C_3, is discharged through the loop of T, L_x and D_4. Then C_3 and L_x will oscillate until the voltage across the capacitor C_3 is charged to a negative input voltage V_g. At this moment D_3 will conduct, which causes the rest of the energy stored in L_x to be charged into the input filter capacitor C_4. Therefore, this circulating energy is:

$$W = \frac{1}{2} \cdot C_3 \cdot (V_{C3}^2 - V_g^2) \cdot \quad (9)$$

In this operating condition, the clamp voltage is independent of the input voltage V_g but is a function of the load as shown by (8). Since the LC oscillates at high resonant frequency, the capacitance of C3 can not be designed to be very large. As a result, the clamp voltage is much higher than the reflected output voltage in heavy load conditions (usually it is designed to be 2 to 4 times of V_{ref}). Therefore, it was very hard for the LC operation mode discussed in [2] to render a clamp voltage close to the reflected output voltage so as to achieve better cross regulation.

A new operating condition is discovered when the resonant frequency is lower than switch frequency. In this operating condition, only a portion of the energy stored in the clamp capacitor C_3 is transferred to inductor L_x through oscillation during the switch turn-on period. After the switch is turned off, the diode, D_3, conducts and the energy in L_x is transfer to input filter capacitor C_4. The polarity of the clamp capacitor remains unchanged, unlike in the case described in [2]. The capacitance of C_3 is assumed to be large enough that the voltage ripple on it can be neglected during the whole switching cycle. Therefore, when the switch is turned off, the current flowing into the clamp capacitor is:

$$i_{C3}(t) = I_0 - \frac{1}{L_{kp}} \cdot (V_{C3} - V_{ref}) \cdot t \quad (10)$$

The D_3 conduction time T can be obtained by letting $i_{C3}(t)=0$

$$T = \frac{I_0 \cdot L_{kp}}{V_{C3} - V_{ref}} \quad (11)$$

The energy increased in the clamp capacitor during the turn-off period is

$$\Delta W_{in} = \frac{1}{2} \cdot L_{kp} \cdot I_0^2 \cdot \frac{V_{C3}}{V_{C3} - V_{ref}} \quad (12)$$

During turn-on period, the energy discharged into inductor L_x is

$$\Delta W_o = \frac{1}{2} \cdot \frac{V_{C3}^2}{L_x} \cdot T_{on}^2 \quad (13)$$

where $T_{on}=(L_m+L_{kp})I_o/V_g$.

In steady-state condition, the energy flowing into and out is balanced. Combination of (12) and (13) yields the clamp voltage,

$$V_{C3} = \frac{1}{2} \cdot V_{ref} + \frac{1}{2} \cdot \sqrt{V_{ref}^2 + \frac{4 \cdot L_{kp} \cdot L_x}{(L_m + L_{kp})^2} \cdot V_g^2} \quad (14)$$

Equation (14) shows that the clamp voltage can be designed to be close to the reflected output voltage, V_{ref}, since the leakage inductance, L_{kp}, is much less than the magnetizing inductance, L_m, in most applications. By selecting a proper size for L_x, the clamp voltage will become less dependent on the input voltage, V_g. For universal input applications, the clamp voltage can easily maintain a value slightly above the reflected output voltage and improve cross regulation. In addition, by operating the LC clamp in this condition, the clamp voltage is independent of the load unlike the case of the RC clamp. Current stress through the switch in DCM condition is

$$I_{sw} = I_0 + \frac{L_m + L_{kp}}{L_x} \cdot \frac{V_{C3}}{V_g} \cdot I_0 \quad (15)$$

This is higher than that of the converter with an RC clamp. For universal input, the switch current stress at the low end is higher than is at the high end. The inductance of Lx is required to be comparable with the magnetizing inductance of transformer so as to limit the current stress through switch.

Voltage stress across switch is

$$V_{sw} = V_g + \frac{1}{2} \cdot V_{ref} + \frac{1}{2} \cdot \sqrt{V_{ref}^2 + \frac{4 \cdot L_{kp} \cdot L_x}{(L_m + L_{kp})^2} \cdot V_g^2} \quad (16)$$

Compared with the operating condition reported in [2], the voltage stress on switch is reduced and independent of load as shown in (16).

Compared with the RC clamp, the LC snubber can be designed for better cross regulation without resulting in power losses for universal input applications. But an extra inductor is required, the capacitor must be larger, and the current stress through the switch is higher.

IV. ENERGY REGENERATIVE CLAMP

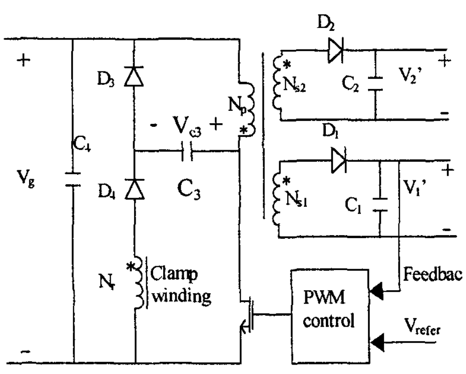

Fig. 5. Proposed energy regenerative clamp

Another simple way to losslessly maintain a low clamp voltage is to use the energy regenerative clamp shown in Fig. 5 and proposed in [4]. This clamp uses an extra winding, N_R that shares the *same core of the transformer*, reducing the number of components compared with the lossless LC snubber. A similar clamp winding for a forward converter was proposed in [5]. When the switch is turned off, the transformer's magnetizing and leakage inductance will initially conduct through C_3 and D_3, clamping the voltage across the switch to $V_g + V_c$. At this time, diode D_4 is reversely biased, no current flows through the clamp winding, and the energy in the leakage inductance is temporally stored in capacitor C_3. It is assumed that the capacitance of C_3 is large enough that the voltage across it approximately maintains constant from cycle to cycle. When the switch is turned on, the magnetizing inductance of the transformer will be charged by two energy sources. Initially it will be charged by C_3 through the clamp winding N_r to transformer core until the voltage across C_3 drops to the reflected input voltage, $V_g \cdot N_r/N_p$. Then it will be charged by the input voltage source V_g. The current waveforms are shown in Fig. 6, where I_p and I_c are the currents through the primary winding and the clamp capacitor, C_3, respectively.

When the switch is on, diode D_3 and all the diodes in output are reversely biased, and the voltage on the capacitors C_3 is approximately equal to that of the clamp winding N_r if the voltage drops on diode D_4 and switch are neglected. The voltages of primary and clamp winding satisfy during switch-on period:

$$V_{c3} = \frac{N_r}{N_p} \cdot V_g \quad (17)$$

Equation (17) indicates that the clamp voltage is determined by the designed turns of the clamp winding. It must be greater than the output reflected voltage so that energy stored in the magnetic core can be converted to the secondary side during switch-off period, otherwise it will be totally transferred to capacitor C_3, i.e.

$$V_{c3} > \frac{V_o \cdot N_p}{N_s} \quad (18)$$

Substitution of (17) into (18) gives design criteria for the clamp winding:

$$N_r > \frac{V_o \cdot N_p^2}{V_g \cdot N_s} \quad (19)$$

Voltage stress across switch is

$$V_{sw} = V_g + V_c = V_g \cdot (1 + \frac{N_r}{N_p}) \quad (20)$$

Equation (20) shows that the voltage stress across the switch is controllably determined by the turns of the clamp winding and does not change with the variations of leakage inductance nor the load currents unlike that in the RC clamp. But the clamp voltage is proportional to the input rms voltage, therefore, the cross regulation at high end (i.e. 220v AC) will be worse than that at low end (i.e. 110v AC) for an universal input.

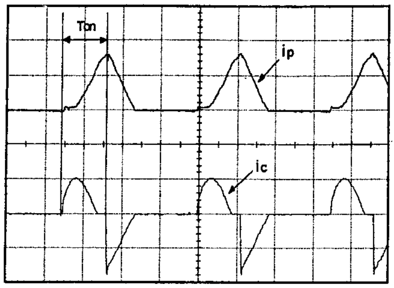

Fig. 6. The measured currents through the primary winding and the clamping capacitor C_3 in flyback converter with energy regenerative clamp.

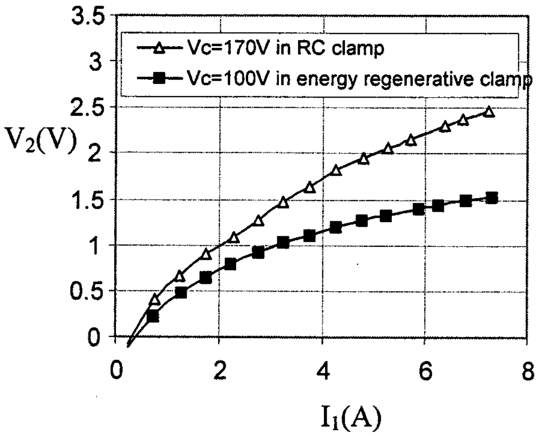

Fig. 7. Comparison of the errors in the unregulated output of the flyback converter with the energy regenerative clamp and that with RC clamp.

Comparative experiments were conducted in the flyback converter with the parameters listed in Table 1. When the flyback converter used the RC clamp, the clamp voltage was designed as 170V at full load (+5V output is 8A). Under this condition, the output error in the +7V terminal is more than 2.6V as shown in Fig. 7. Using the same flyback converter, a new clamp winding with 21 turns was added to the transformer. This winding gives a clamp voltage of 100V when the input AC voltage is about 120V. After the proposed energy regenerative clamp was employed, the error on +7 output was only 1.5V with the same full load current on +5V output as shown by the curve with diamond in Fig. 7. It was reduced by 42 percent compared with that of the RC clamp. However, because of the dependency of the clamp voltage on the input voltage, when the input AC is 220V, the clamp voltage becomes 170V, and cross regulation is close to that of the RC clamp. Another major advantage of the energy regenerative clamp is that it improves overall efficiency about 7.7 percent due to regeneration of the leakage inductance energy.

Table 1. Experimental flyback converter parameters

Turns of primary	36
Turns of output 1	3
Turns of output 2	4
L_m	293μH
L_{kp}	9.9μH
L_{k1}(converted to primary side)	9.8μH
L_{k2}(converted to primary side)	9.8μH
R_1 (Pre-load in output 1)	24Ω
R_2 (Pre-load in output 2)	66.7Ω
V_{ref}(output reflected voltage)	60V

V. CONCLUSION

Among the three clamps discussed above, RC clamp is the simplest and cheapest approach to limit the voltage stress across switch. Its clamp voltage is dependent on the output load of converters but not on the input voltage, which leads to the same cross regulation for universal input applications. Since it is a dissipative clamp, decreasing its designed clamp voltage to achieve better cross regulation is at the cost of the efficiency. Therefore, in the design procedure, it is indispensable to make a tradeoff between the cross regulation and efficiency when RC clamp is used. The new operating condition for the LC snubber provides a solution to lower the clamp voltage to near the reflected output voltage for universal input without resulting in dissipation. Its clamp voltage is independent on load unlike that in RC clamp, but the current stress through switch is much higher than that in the RC clamp. It also requires an extra inductor with a value comparable with the magnetizing inductance of transformer so as to limit the current stress through switch. The proposed energy regenerative clamp overcomes the defect in the LC snubber. It does not need an extra magnetic core by using a clamp winding to share the same core with the transformer, and allows the clamp voltage to be much lower than that of RC clamp without yielding losses by recovering leakage energy into transformer core. The switch current stress is less than that in the LC snubber with the suggested operating condition. It also has the merit that its clamp voltage doesn't alter in terms of the load in contrast with the RC clamp and it only uses the same number of components as the RC clamp. Therefore, it is a cost-effective approach to improve both efficiency and cross regulation in multiple output flyback converters. However, due to the dependency of its clamp voltage on the input voltage, the cross regulation cannot be maintained the same for universal input applications. It will suffer at the high end when the clamp is designed for low end input. The comparisons of the three discussed clamps are summarized in Table 2.

Table 2. Comparison of RC clamp, nondissipative LC snubber, and energy regenerative clamp.

	RC Clamp	Nondissipative LC Snubber	Energy Regenerative Clamp
Simplicity and Cost	Simplest	expensive	simple
Efficiency	Loss	lossless	lossless
Clamp Voltage Dependence on Input Voltage	No	not sensitive	linear
Clamp Voltage Dependence on Load	Yes	no, but depends on leakage induct.	no
Universal Input Voltage Range	best for universal	suitable for universal input	best for non-universal input
Voltage Stress across Switch	high	low	same as LC snubber
Current Stress through Switch	lowest	highest	medium
Extra Magnetic Core	no	yes	no

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