Passive Lossless Snubbers for DC/DC Converters

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Abstract - Passive lossless snubbers are proposed to improve the turn-on and turn-off transients of the MOSFET's in PWM dc/dc converters. Switching losses and EMI noises are reduced by restricting di/dt of the reverse recovery current and dv/dt of drain-source voltage. The MOSFET operates at ZVS turn-off and near ZCS turn-on. The freewheeling diode is also commutated with ZVS. To be a demonstration, operation principles, theoretical analyses, relevant equations and experimental results of a boost converter equipped with the proposed snubber are presented in detail. Efficiency of 96% has also been measured in the experimental results reported for a 1kW, 100kHz prototype in the laboratory. Six basic nonisolated PWM dc/dc converters (buck, boost, buck-boost, Cúk, Sepic and Zeta) equipped with the proposed passive lossless snubbers are also shown in this paper.

I. INTRODUCTION

Pulse width modulated (PWM) dc/dc converters have been widely used as switched mode power supplies in industries. The PWM technique is praised for its high power capability and ease of control. Higher power density and faster transient response of PWM dc/dc converters can be achieved by increasing the switching frequency. However, as the switching frequency increases, so do the switching losses and EMI noises. High switching losses reduce the power capabilities while serious EMI noises interfere the control of PWM dc/dc converters.

Switching losses and EMI noises of PWM dc/dc converters are mainly generated during turn-on and turn-off switching transients. According to [1], there are three different non-ideal commutation phenomena when MOSFET's are used as power switches:

1. A surge current flows through the MOSFET caused by the reverse recovery current of the freewheeling diode during turn-on process. This is the dominant part of switching losses and the di/dt EMI noise source.
2. Discharge of the parasitic drain-source capacitance of the MOSFET during turn-on process. This mechanism can be reduced only by resonant converter techniques or active snubbers.

3. Fast increase of the drain-source voltage during turn-off process. This is the source of dv/dt EMI noise and turn-off loss.

To improve the problems resulted from the non-ideal phenomena described above, several kinds of soft switching technologies are presented in literature [2-9]. Active snubbers as introduced in [2-5] can reduce all three loss mechanisms by using auxiliary switches. Unfortunately, an auxiliary switch increases the complexity of both the power circuit and the control circuit. Synchronization problems between control signals of two switches during transient also complicate the control strategy. Circuit cost is increased and reliability is decreased by using active snubbers. RCD snubbers have the simplest structures and hence lowest costs [6]. However, it also has the worst performance since the switching losses are dissipated through resistors and thus reduce the efficiency of the circuit. Resonant converters commute with either zero-voltage-switching (ZVS) or zero-current-switching (ZCS) to reduce switching losses [7-8]. However, conduction losses and current stresses are increased due to high circulating current involved. It is also hard to design EMI filter and control circuit because of a wide switching frequency range. Comparing with three technologies discussed above, a passive lossless snubber can effectively restrict switching losses and EMI noises using no active components and no power dissipative components [9]. No additional control is needed and no circulating energy is generated. Circuit structure is as simple as RCD snubbers while circuit efficiency is as high as active snubbers and resonant converters. Low cost, high performance and high reliability are the distinct advantages of a passive lossless snubber.

To be a demonstration, a boost converter equipped with the proposed snubber is investigated in depth. Snubber operation principles are analyzed and the component parameters can be mathematically determined by the design rules presented. Experimental results of a 1kW, 100kHz boost converter are reported to verify the analysis. Six basic PWM dc/dc converters equipped with the proposed snubbers are also illustrated in this paper.

II. A BOOST CONVERTER EQUIPPED WITH THE PASSIVE LOSSLESS SNUBBER

A. Principle of Operation

Shown in Fig. 1 is a boost converter with the proposed passive lossless snubber, which is encircled by dotted lines. During turn-on process, injected charges in the low-doped middle region of diode D1 cause transient reverse recovery current flowing reversely through diode D1. The surge current is the major cause of the switching losses. Growth rate of the reverse recovery current is restricted by the snubber inductor L_s to suppress the switching loss. During turn-off process, drain-source voltage increases immediately to output voltage. Fast dv/dt of drain-source voltage increases the turn-off loss, and of more importance, it generates serious EMI noises. Growth rate of drain-source voltage is restricted by the snubber capacitor C_s to obtain ZVS turn-off and to reduce EMI noises. Notice that the freewheeling diode is also commutated with ZVS during both turn-on and turn-off.

Fig. 1 A boost converter with the passive lossless snubber.

Turn-on and turn-off switching losses are reduced by the snubber inductor and the snubber capacitor, respectively. The energy transferred to the buffer capacitor C_b can be viewed as the summation of the energy absorbed in snubber inductor L_s and snubber capacitor C_s . Energy recovery is achieved by discharging the buffer capacitor C_b to the output. Ideally, the absorbed power is neither dissipated nor accumulated in the passive components of this snubber.

B. Equivalent Circuit Analysis

To analyze the steady state operations of the circuit shown in Fig. 1, following assumptions are made during one switching cycle:

- (1). The output capacitor C_o is large enough to assume that the output voltage V_o is constant and ripple-free.
- (2). The input voltage V_s is constant.
- (3). All semiconductor devices are ideal, except the freewheeling diode D1.
- (4). The main inductor L_m is much greater than the snubber inductor L_s .

Based on these assumptions, circuit operations in one switching cycle can be divided into eight stages as shown in Fig. 2 (a)-(h), respectively.

Fig. 2. Equivalent circuits during one switching cycle.

Stage 1 (Fig 2.a ; $t_0 < t < t_1$) :

The switch S_1 turns on at t_0 . During turn-on process, freewheeling diode D1 is not immediately turned off because of the reverse recovery phenomenon. Growth rate of the drain current is restricted by the snubber inductor to softly turn on the MOSFET. Current of the snubber inductor L_s is given by

$$I_{L_s}(t) = I_F(t_0) - \frac{V_o}{L_s}(t - t_0) \quad (1)$$

where $I_F(t)$ is the forward current flowing from the input.

Stage 2 (Fig 2.b ; $t_1 < t < t_2$) :

The reverse recovery phenomenon finishes at t_1 . As soon as D1 is turned off, diode D3 is naturally turned on because V_{C_s} and V_{C_b} are equal to zero. Snubber

inductor L_s , snubber capacitor C_s and buffer capacitor C_b are charged by the output through the first resonant path $V_o - C_s - D_3 - C_b - L_s - S_1$. Growth rate of the voltage across D_1 , which is equal to $V_{Cs} + V_{Cb}$, is restricted to achieve ZVS turn-off of freewheeling diode D_1 . Snubber inductor current, snubber capacitor voltage and buffer capacitor voltage are:

$$I_{L_s}(t) = -\frac{V_o}{Z_1} \sin(\omega_1(t-t_1)) - I_r \cos(\omega_1(t-t_1)) \quad (2)$$

$$V(t) = I_r Z_1 \sin(\omega_1(t-t_1)) - V_o \cos(\omega_1(t-t_1)) + V_o \quad (3)$$

$$V_{Cs}(t) = \frac{C_b}{C_s + C_b} V(t), \quad V_{Cb}(t) = \frac{C_s}{C_s + C_b} V(t) \quad (4)$$

where

$$I_r = \frac{V_o}{L_s} (t_1 - t_0) - I_r(t_0) \quad (5)$$

$$Z_1 = \sqrt{\frac{L_s(C_s + C_b)}{C_s C_b}} \quad (6)$$

$$\omega_1 = \sqrt{\frac{C_s + C_b}{L_s C_s C_b}} \quad (7)$$

Peak value of the drain current of switch S_1 is obtained by the summation of forward current I_F and peak snubber inductor current $I_{Ls,p}$. The peak value appears when $V_{Cb} + V_{Cs}$ is equal to V_o , and it is given by

$$I_{Ls,p} = \frac{\sqrt{V_o^2 + (I_r Z_1)^2}}{Z_1} \quad (8)$$

The first resonance stops at t_2 when $V_{Cs}(t_2)$ equals V_o because diode D_2 is turned on. By using reciprocity theorem, snubber inductor current at t_2 is given by

$$I_{Ls}(t_2) = \frac{\sqrt{(I_r Z_1)^2 + V_o^2 - (V_o \frac{C_s}{C_b})^2}}{Z_1} \quad (9)$$

From (9), the energy stored in L_s and C_s can be given by

$$\begin{aligned} E_{Ls}(t_2) + E_{Cs}(t_2) &= \frac{1}{2} L_s I_{Ls}^2(t_2) + \frac{1}{2} C_b V_{Cb}^2(t_2) \\ &= \frac{1}{2} L_s I_r^2 + \frac{1}{2} C_s V_o^2 \end{aligned} \quad (10)$$

Stage 3 (Fig 2.c ; $t_2 < t < t_3$) :

After V_{Cs} is charged to output voltage V_o at t_2 , D_2 is turned on and V_{Cs} keeps constant. The current in L_s starts to charge C_b through the second resonant path $L_s - D_2 - D_3 - C_b$. L_s and C_b are performing one-way

resonance because of diode D_2 and D_3 . The current through L_s and the voltage across C_b are given by

$$I_{Ls}(t) = \frac{C_s}{C_b} \frac{V_o}{Z_2} \sin(\omega_2(t-t_2)) - I_{s2} \cos(\omega_2(t-t_2)) \quad (11)$$

$$V_{Cb}(t) = I_{s2} Z_2 \sin(\omega_2(t-t_2)) + \frac{C_s}{C_b} V_o \cos(\omega_2(t-t_2)) \quad (12)$$

where

$$I_{s2} = \frac{V_o}{Z_1} \sin(\omega_1(t_2-t_1)) + I_r \cos(\omega_1(t_2-t_1)) \quad (13)$$

$$Z_2 = \sqrt{\frac{L_s}{C_b}} \quad (14)$$

$$\omega_2 = \sqrt{\frac{1}{L_s C_b}} \quad (15)$$

The second resonance stops at t_3 when $I_{Ls}(t_3)$ equals zero. Since the energy in L_s is completely transferred to C_b in this stage, the energy stored in C_b at t_3 can be found following (10) to be

$$\begin{aligned} \frac{1}{2} C_b V_{Cb}^2(t_3) &= E_{Cb(t_3)} = E_{Ls(t_2)} + E_{Cb(t_2)} \\ &= \frac{1}{2} L_s I_r^2 + \frac{1}{2} C_s V_o^2 \end{aligned} \quad (16)$$

Also, the peak voltage of the buffer capacitor $V_{Cb,p}$ is given by

$$V_{Cb,p} = V_{Cb}(t_3) = \sqrt{\frac{L_s I_r^2 + C_s V_o^2}{C_b}} \quad (17)$$

It also determines the voltage stress of freewheeling diode, which is equal to V_o plus $V_{Cb,p}$.

Stage 4 (Fig 2.d ; $t_3 < t < t_4$) :

At t_3 , I_{Ls} is decreased to zero while D_2 and D_3 are turned off. The current through L_s keeps zero and the voltage across C_b keeps constant after t_3 . From (16), the total energy transferred to C_b can be viewed as the summation of the energy which were absorbed in L_s and C_s . The circuit operation in this stage is identical to the normal turn-off state of a conventional boost converter.

Stage 5 (Fig 2.e ; $t_4 < t < t_5$) :

After the switch S_1 turns off at t_4 , forward current $I_F(t_4)$ flows through D_2 to discharge C_s to the output. Diodes D_3 and D_4 are not turned on because they are reverse biased by V_{Cs} . Drain-source voltage of S_1 is equal to $V_o - V_{Cs}$. Slow dv/dt of drain-source voltage is obtained while V_{Cs} is discharged from V_o to zero.

Assuming that $I_F(t)$ is constant during this stage, V_{Cs} is given by

$$V_{Cs}(t) = V_o - \frac{I_F(t_4)}{C_s}(t - t_4) \quad (18)$$

Stage 6 (Fig 2.f ; $t_5 < t < t_6$) :

Diodes D3 and D4 are turned on by the forward current $I_F(t_5)$ when V_{Cs} is discharged to zero at t_5 . Voltage across L_s equals V_{Cb} and makes I_{Ls} increased to discharge C_b to output. Circuit operation is similar to the second resonance in stage 2. I_{Ls} and V_{Cb} are given by

$$I_{Ls}(t) = \frac{V_{Cb}(t_2)}{Z_2} \sin(\omega_2(t - t_5)) \quad (19)$$

$$V_{Cb}(t) = V_{Cb}(t_2) \cos(\omega_2(t - t_5)) \quad (20)$$

Stage 7 (Fig 2.g ; $t_6 < t < t_7$) :

Inductor current I_{Ls} is increased to $I_F(t_6)$ at t_6 , D2 and D3 are turned off. After t_6 , $I_F(t)$ discharges C_b to output through D4. ZVS turn-on of the diode D1 is achieved by slow dv/dt of V_{Cb} . Assuming that $I_F(t)$ is constant in this stage, V_{Cb} is given by

$$V_{Cb}(t) = V_{Cb}(t_6) - \frac{I_F(t_6)}{C_b}(t - t_6) \quad (21)$$

Stage 8 (Fig 2.h ; $t_7 < t < t_0$) :

Capacitor voltage V_{Cb} is discharged to zero at t_7 . D4 is turned off and D1 is turned on simultaneously. Snubber energy recovery process is accomplished when all energy in the buffer capacitor C_b is transferred to the output. After that, input current $I_F(t)$ flows through D1 instead of D4 to prevent C_s from being charged reversely. Circuit operation will be the same as in stage 1 when the switch S_1 turns on again at t_0 in the next switching cycle.

Fig 3. Key waveforms of the boost converter with the passive lossless snubber.

C. Design Considerations

The snubber inductor L_s , snubber capacitor C_s and buffer capacitor C_b are the three main elements to be designed. Following rules should be noticed when designing L , C values.

1. In stage 6, diodes D2 and D3 should be naturally turned off before the voltage of C_b is discharged to zero, or the residential current will turn on D2, D3 and D4 for the entire switching period. In other words, following inequality has to be obeyed :

$$\frac{1}{2} L_s I_F^2 < \frac{1}{2} L_s I_{rr}^2 + \frac{1}{2} C_s V_o^2 \quad (22)$$

It requires higher I_{rr} or larger C_s .

2. Current stress of MOSFET and voltage stress of freewheeling diode are given in (8) and (17), respectively. Larger C_s results in higher MOSFET current stress and higher diode voltage stress.
3. According to (17), C_b has to be at least 16 times as C_s to limit V_{Cb} to 100V with a 400V output. Practically, C_b should be about 30 times as C_s considering reverse recovery energy.
4. Snubber inductor L_s should be selected as large as possible to decrease reverse recovery loss. According to the following equation in [10], larger L_s results in lower I_{rr} .

$$I_{rr} \propto \sqrt{I_F \frac{dI_F}{dt}} \propto \sqrt{\frac{I_F}{L_s}} \quad (23)$$

5. Resonant frequency in (15) should be much larger than switching frequency to ensure correct operation of the snubber cell.

Trade-offs have to be made when designing L_s , C_s and C_b . Voltage and current stresses of diode D2, D3 and D4 are equal to the output voltage and the input current. However, lower component ratings are also acceptable due to short snubber operating time. Voltage stress of D1 and current stress of MOSFET are increased by $V_{Cb,p}$ and $I_{Ls,p}$, respectively. Voltage stress of MOSFET and current stress of D1 are the same as those without the snubber embedded.

III. THE GENERAL SNUBBER CELL FOR DC/DC CONVERTERS

The proposed snubber cell can be seen as the combination of a turn-on snubber cell and a turn-off snubber cell. The turn-on snubber, shown in Fig. 4(a), limits di/dt of the reverse recovery current by a snubber inductor in series

with freewheeling diode. Two diodes and one capacitor are added to recover the absorbed energy to output. Most of the turn-off snubber proposed in literature use a snubber capacitor parallel to the switch to limit dv/dt of drain-source voltage. However, in a switching circuit, the voltage across the freewheeling diode always varies with the drain-source voltage of the switch. In other words, dv/dt of drain-source voltage can also be limited in the turn-off snubber, shown in Fig. 4(b), by paralleling a snubber capacitor to the diode. An additional diode is added to isolate the switch from the snubber capacitor [2]. The isolation can prevent the snubber capacitor from being discharged in the switch at light load and high line.

Fig 4. General passive lossless turn-on and turn-off snubber cells

Combining the turn-on and turn-off snubber cells described above, the proposed passive lossless snubber cell for nonisolated PWM dc/dc converters is defined and shown in Fig. 5. Node *A* and *K* are connected to the anode and the cathode of the converter freewheeling diode *D*₁, respectively. Node *A'* is connected to the component which was connected to the anode of the freewheeling diode in the original circuit.

Fig 5. The general passive lossless snubber cell proposed for nonisolated PWM dc/dc converters.

The proposed general snubber cell consists of one inductor *L*_s, two capacitors *C*_s and *C*_b and three diodes *D*₂, *D*₃ and *D*₄. The snubber inductor *L*_s is placed in series with the freewheeling diode *D*₁. It is designed to restrict di/dt of the reverse recovery current to achieve ZCS turn-on. The snubber capacitor *C*_s is placed in parallel with *D*₃ and *D*₄ and isolated by diode *D*₂. *D*₃ and *D*₄ are freewheeling during turn-off. It is designed to restrict dv/dt of drain-source voltage to achieve ZVS turn-off. ZVS turn-on and turn-off of freewheeling diode are also obtained. Switching losses and EMI noises during turn-on and turn-off are eliminated by the snubber cell. All energy absorbed in the snubber inductor and snubber capacitor are transferred to the buffer capacitor *C*_b. Energy recovery is achieved by discharging *C*_b to the output. Snubber operation principles

discussed for a boost converter can be extended to other topologies. Six basic nonisolated PWM dc/dc converters : Buck, Boost, Buck-boost, Cúk, Sepic and Zeta with the proposed snubbers embedded are shown in Fig. 6.

Fig 6. Six basic nonisolated PWM dc/dc converters with the proposed snubbers.

IV. EXPERIMENTAL RESULTS

A prototype of 1 kW, 100 kHz, 400V DC output boost converter with the passive lossless snubber has been built to verify the principle of operation and the theoretical analysis. The components specifications are listed in Table 1.

TABLE 1
PART LIST OF THE IMPLEMENTED PROTOTYPE POWER CIRCUIT.

Part	Type	Part	Value
S1	IRFP460	Ls	5μH
D1	HFA15TB60	Cs	3.3nF
D2	HFA15TB60	Cb	100nF
D3	HFA15TB60	Lm	180μH
D4	HFA15TB60	Co	940μF

Efficiency of 96% at 1kW has been measured. The snubber inductor current, snubber capacitor voltage and buffer capacitor voltage waveforms are shown in Fig. 7. The commutation waveforms of the MOSFET and the freewheeling diode with the proposed snubber embedded are shown in Fig. 8 and Fig. 9, respectively. Waveforms of MOSFET without snubber are shown in Fig. 10.

Fig 7. Waveforms of snubber inductor current I_{Ls} , snubber capacitor voltage V_{Cs} and buffer capacitor voltage V_{Cb} .

Fig 8. Waveforms of MOSFET commutation with snubber.

Fig 9. Waveforms of freewheeling diode commutation with snubber.

Fig 10. Waveforms of MOSFET commutation without snubber.

Waveforms in Fig. 7(a) and Fig. 7(b) are exactly the same as predicted in Fig. 3. Snubber operation analysis is approved to be valid. Comparing Fig. 8(a) and Fig. 10(a), it can be seen that di/dt of drain current is restricted and commutation of the MOSFET is close to ZCS turn-on. The reason that the MOSFET operates only close to ZCS is the discharge of the parasitic drain-source capacitance of MOSFET during turn-on process. This switching loss can only be removed by resonant converter techniques or active snubbers. Comparing Fig. 8(b) and Fig. 10(b), it can also be seen that dv/dt of drain-source voltage is restricted and the MOSFET commutates at ZVS turn-off. It can be seen from

Fig. 9(a) and Fig. 9(b) that the freewheeling diode is also commutated at ZVS turn-on and turn-off. EMI noises are reduced due to slower di/dt of drain current.

V. CONCLUSION

Passive lossless snubbers for PWM dc/dc converters are proposed in this paper. The general snubber cell is the combination of a turn-on snubber and a turn-off snubber. Energy recovery is achieved by passive components only. Component values of snubber inductor, snubber capacitor and buffer capacitor can be determined by the designing rules presented in this paper. Current stress of MOSFET and voltage stress of freewheeling diode are also presented clearly. A 1kW, 100kHz prototype of boost converter equipped with the snubber has been implemented in the laboratory to verify the analyses above. Experimental waveforms show that the growth rates of the reverse recovery current and drain-source voltage are successfully restricted by the snubber. The MOSFET operates at ZVS turn-off and close to ZCS turn-on, the freewheeling diode operates at ZVS turn-on and turn-off.

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