A Passive Lossless Snubber Cell for Nonisolated PWM DC/DC Converters

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Abstract—A passive lossless snubber cell is proposed to improve the turn-on and turnoff transients of the MOSFET’s in nonisolated pulsewidth modulated (PWM) dc/dc converters. Switching losses and EMI noise are reduced by restricting $di/dt$ of the reverse-recovery current and $dv/dt$ of the drain–source voltage. The MOSFET operates at zero-voltage-switching (ZVS) turnoff and near zero-current-switching (ZCS) turn-on. The freewheeling diode is also commutated under ZVS. As an example, operation principles, theoretical analysis, relevant equations, and experimental results of a boost converter equipped with the proposed snubber cell are presented in detail. Efficiency of 96% has also been measured in the experimental results reported for a 1-kW 100-kHz prototype in the laboratory. Six basic nonisolated PWM dc/dc converters (buck, boost, buck-boost, Cuk, Sepic, and Zeta) equipped with the proposed general snubber cells are also shown in this paper.

Index Terms—Converters, pulsewidth modulation, switching circuits.

I. INTRODUCTION

PULSEWIDTH modulated (PWM) dc/dc converters have been widely used as switched-mode power supplies in industry. 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 noise. High switching losses reduce the power capabilities, while serious EMI noise interferes with the control of PWM dc/dc converters.

Switching losses and EMI noise of PWM dc/dc converters are mainly generated during turn-on and turnoff switching transients. According to [1], there are three nonideal 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 the turn-on process. This is the dominant source of turn-on loss and $dv/dt$ EMI noise.

2) Discharge of the parasitic drain–source capacitance of the MOSFET occurs during the turn-on process. This mechanism can be eliminated only by resonant converter techniques or active snubbers.

3) Fast increase of the drain–source voltage during the turnoff process occurs. This is the source of $dv/dt$ EMI noise and turnoff loss.

To improve the problems resulting from the nonideal phenomena described above, several kinds of soft-switching technologies have been presented in the literature [2]–[13]. 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 power and control circuits. Synchronization problems between control signals of the two switches during transient also complicate the control strategy. The circuit cost is increased and the reliability is decreased by using active snubbers. Resistors, capacitors, and diodes (RCD) snubbers in [6] have the simplest structures and, hence, the lowest costs. However, they also have the worst performance, since the switching losses are dissipated in resistors and, thus, reduce the efficiency of the circuit. Resonant converters in [7] and [8] commutate with either zero voltage switching (ZVS) or zero current switching (ZCS) to reduce switching losses. However, conduction losses are increased due to the high circulating current involved. It is also hard to design an EMI filter and control circuit because of a wide switching frequency range. Compared with the three technologies discussed above, a passive lossless snubber can effectively restrict switching losses and EMI noise using no active components and no power dissipative components [9]–[13]. Increasing the rates of the drain current and the drain–source voltage are restricted by inductors and capacitors, respectively. The control strategy is scarcely interfered with and the circulating energy generated is comparatively low. The circuit structure is as simple as RCD snubbers and the 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.

As an example, a boost converter equipped with the proposed snubber cell is investigated in depth. Snubber operation principles are analyzed and component parameters can be
mathematically determined. Experimental results of a 1-kW 100-kHz boost converter are used to verify the analysis. Formation of the general snubber cell is also discussed. Six basic nonisolated PWM dc/dc converters equipped with the proposed snubber cells are illustrated in this paper.

II. A BOOST CONVERTER WITH THE PROPOSED SNUBBER CELL

A. Principle of Operation

Shown in Fig. 1 is a boost converter with the proposed passive lossless snubber cell, which is encircled by dotted lines. During the turn-on process, injected charges in the low-doped middle region of diode $D_1$ cause transient reverse-recovery current flowing reversely through diode $D_2$. The surge current is the major part of the switching losses. Increasing the rate of the reverse-recovery current is restricted by the snubber inductor $L_s$ to suppress the switching loss. The MOSFET commutates near ZCS turn-on, since part of the turn-on loss resulting from the discharge of the parasitic drain–source capacitance cannot be removed by a passive snubber.

During the turnoff process, the drain–source voltage increases immediately to the output voltage. Fast $dv/dt$ increases...
the turnoff loss and, of more importance, it generates serious 
EMI noise. Increasing the rate of the drain–source voltage is 
restricted by the snubber capacitor $C_s$ to obtain ZVS turnoff 
and to reduce EMI noise. Notice that the freewheeling diode is 
also commutated under ZVS during both turn-on and turnoff. 

Turn-on and turnoff 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, no 
power is dissipated or accumulated in the passive components 
of this snubber.

B. Equivalent Circuit Analysis

To analyze the steady-state operation of the circuit shown 
in Fig. 1, the 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) Input voltage $V_s$ is constant.
3) All semiconductor devices are ideal, except the free-
wheeling diode $D_s$. 
4) Main inductor $L_m$ is much greater than snubber inductor 
$L_s$.

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

Combining the turn-on and turnoff snubber cells described 
above, the proposed passive lossless snubber cell for noniso-
lated 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 $D_s$, respectively. Node $A'$ 
is connected to the component which is connected to the anode 
of the freewheeling diode.

Stage 1 [Fig. 2(a); $t_0 < t < t_1$]: $S_1$ turns on at $t_0$. During 
the turn-on process, $D_s$ is not turned off immediately because 
of the reverse-recovery phenomenon. Increasing the rate of the 
drain current is restricted by the snubber inductor to achieve 
ZCS turn-on of the MOSFET. The current of the snubber 
inductor $L_s$ is given by

$$I_{L_s}(t) = I_{Ls}(t_0) - \frac{V_o}{L_s}(t - t_0).$$

Stage 2 [Fig. 2(b); $t_1 < t < t_2$]: The reverse-recovery phe-

omenon finishes at $t_1$. As soon as $D_s$ is turned off, diode 
$D_3$ is naturally turned on, because $V_{C_b}$ is 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 $V_o - C_s - D_3 - C_b - L_s - S_1$. Increasing the 
rate of the voltage across $D_3$, which is equal to $V_{C_b} + V_{C_s}$, 
is restricted to achieve ZVS turnoff of the freewheeling diode 
$D_3$. Snubber inductor current, snubber capacitor voltage, and 
buffer capacitor voltage are

$$I_{Ls}(t) = \frac{V_o}{Z_1} \sin(\omega_1(t - t_1)) - I_{rr} \cos(\omega_1(t - t_1))$$

$$V(t) = I_{rr} Z_1 \sin(\omega_1(t - t_1)) - V_o \cos(\omega_1(t - t_1))$$

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

$$E(L_s(t_2)) + E(C_v(t_2)) = \frac{1}{2} L_s I_{Ls}^2 (t_2) + \frac{1}{2} C_v V_{C_v}^2 (t_2)$$

$$= \frac{1}{2} L_s I_{Ls}^2 + \frac{1}{2} C_v V_{C_v}^2.$$  \hspace{1cm} (10)

Stage 3 [Fig. 2(c); $t_2 < t < t_3$]: After $V_{C_v}$ is charged to the output voltage level at $t_2$, $D_2$ is turned on and $V_{C_v}$ keeps constant. The current in $L_s$ starts to charge $C_b$ through the second resonant path $L_s - D_2 - D_b - C_b$. $L_s$ and $C_b$ are performing one-way resonance because of diodes $D_2$ and $D_b$. The current through $L_s$ and the voltage across $C_b$ are given by

$$I_L(t) = \frac{C_s V_o}{C_b Z_2} \text{sin} (\omega_2 (t - t_2)) - I_{S2} \text{cos} (\omega_2 (t - t_2))$$

$$V_{C_b}(t) = I_{S2} Z_2 \text{sin} (\omega_2 (t - t_2)) + \frac{C_s}{C_b} V_o \text{cos} (\omega_2 (t - t_2))$$ \hspace{1cm} (11) \hspace{1cm} (12)

where

$$I_{S2} = \frac{V_o}{Z_1} \text{sin} (\omega_1 (t_2 - t_1)) + I_{Lr} \text{cos} (\omega_1 (t_2 - t_1))$$

$$Z_2 = \sqrt{\frac{L_s}{C_b}}$$

$$\omega_2 = \sqrt{\frac{1}{L_s C_b}}.$$ \hspace{1cm} (13) \hspace{1cm} (14) \hspace{1cm} (15)

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

$$I_{Ls}(t_2) = \frac{(I_{rr} Z_1)^2 + V_o^2 - \left(\frac{V_o C_s}{C_b}\right)^2}{Z_1}.$$ \hspace{1cm} (9)
Fig. 7. Waveforms of snubber inductor current \( I_{L_s} \), snubber capacitor voltage \( V_{C_s} \), and buffer capacitor voltage \( V_{C_b} \). \( a \) Waveforms of \( I_{L_s} \) and \( V_{C_s} \). \( b \) Waveforms of \( I_{L_s} \) and \( V_{C_b} \).

The second resonance stops at \( t_3 \) when \( I_{L_s}(t_3) \) becomes 0. 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

\[
\frac{1}{2} C_b V_{C_b}^2(t_3) = E_{CB}(t_3) = E_{LS}(t_2) + E_{CV}(t_2) = \frac{1}{2} L_s I_{IM}^2(t_2) + \frac{1}{2} C_s V_o^2.
\]

(16)

Also, the peak buffer capacitor voltage \( V_{C_b,p} \) is given by

\[
V_{C_b,p} = V_{C_b}(t_3) = \sqrt{\frac{L_s I_{IM}^2 + C_s V_o^2}{C_b}}.
\]

(17)

It also determines the voltage stress of the freewheeling diode, which is equal to \( V_o \) plus \( V_{C_b,p} \).

Stage 4 [Fig. 2(d); \( t_3 < t < t_4 \)]: At \( t_3 \), \( I_{L_s} \) is decreased to zero, while \( D_2 \) and \( D_3 \) are turned off simultaneously. 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 was absorbed in \( L_s \) and \( C_s \).

Stage 5 [Fig. 2(e); \( t_4 < t < t_5 \)]: After the switch \( S_1 \) turns off at \( t_4 \), main inductor current \( I_{IM}(t_4) \) flows through \( D_2 \) to discharge \( C_s \) to the output. \( D_3 \) and \( D_4 \) are not turned on because they are reverse biased by \( V_{C_b} \). The drain–source voltage of \( S_1 \) is equal to \( V_o - V_{C_b} \). Slower \( dv/dt \) of the drain–source voltage is obtained while \( V_{C_b} \) is discharged from \( V_o \) to 0. Assuming that \( I_{IM} \) is constant during this stage, \( V_{C_b} \) is given by

\[
V_{C_b}(t) = V_o - \frac{I_{IM}(t_4)}{C_b} (t - t_4).
\]

(18)

Stage 6 [Fig. 2(f); \( t_5 < t < t_6 \)]: Diodes \( D_3 \) and \( D_4 \) are turned on by the main inductor current \( I_{IM}(t_5) \) when \( V_{C_b} \) is discharged to zero at \( t_5 \). Voltage across \( L_s \) is equal to \( V_{C_b} \) and, thus, \( I_{L_s} \) increases to discharge \( C_b \) to the output. Circuit operation is similar to the second resonance in Stage 2. \( I_{L_s} \) and \( V_{C_b} \) are given by

\[
I_{L_s}(t) = \frac{V_{C_b}(t_5)}{Z_2} \sin(\omega(t - t_5))
\]

(19)

\[
V_{C_b}(t) = V_{C_b}(t_5) \cos(\omega(t - t_5)).
\]

(20)

Stage 7 [Fig. 2(g); \( t_6 < t < t_7 \)]: \( I_{L_s} \) is increased to \( I_{IM}(t_6) \) at \( t_6 \); \( D_2 \), and \( D_3 \) are turned off. After \( t_6 \), \( I_{IM} \) discharges \( C_b \) to the output through \( D_4 \). Assuming that \( I_{IM} \) is constant in this stage, \( V_{C_b} \) is given by

\[
V_{C_b}(t) = V_{C_b}(t_6) - \frac{I_{IM}(t_6)}{C_b} (t - t_6).
\]

(21)

ZVS turn-on of diode \( D_4 \) is achieved by slow \( dv/dt \) of \( V_{C_b} \).

Stage 8 [Fig. 2(h); \( t_7 < t < t_8 \)]: \( V_{C_b} \) is discharged to zero at \( t_7 \). \( D_4 \) is turned off and \( D_1 \) is turned on. The energy recovery process of the snubber is finished when all energy in the buffer capacitor \( C_b \) is transferred to the output. After that,
main inductor current $I_{Lm}$ flows through diode $D_3$ instead of diode $D_4$ 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.

Based on the analysis presented above, key waveforms of the boost converter with the proposed snubber cell are shown in Fig. 3.

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. The following rules should be noted when designing $L_sC$ values.

1) In Stage 6, diodes $D_2$ and $D_3$ should be naturally turned off before the voltage of $C_b$ is discharged to zero, or the residential current will turn on $D_2$, $D_3$, and $D_4$ in the entire switching period. In other words, the following inequality has to be obeyed:

$$\frac{1}{2} L_s^2 V_o^2 < \frac{1}{2} L_s^2 P_{in} + \frac{1}{2} C_b V_o^2. \quad (22)$$

It requires higher $I_{tr}$ or larger $C_s$.

2) Current stress of the MOSFET and voltage stress of the 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 $C_s$ to limit $V_{CB}$ to 100 V when the output voltage is 400 V.

Practically, $C_b$ should be about 30 times $C_s$, considering reverse-recovery energy.

4) Snubber inductor $L_s$ should be selected as large as possible to decrease reverse-recovery loss. However, according to the following equation in [14], larger $L_s$ results in lower $I_{tr}$:

$$I_{tr} \propto \sqrt{\frac{dP_t}{dt}} \propto \sqrt{\frac{I_F}{L_s}}. \quad (23)$$

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

Tradeoffs have to be made when designing $L_s$, $C_s$, and $C_b$. Voltage and current stresses of diodes $D_2$, $D_3$, and $D_4$ are determined by the output voltage and the input current. However, a lower rating is also acceptable due to short snubber operating time. Current stress of the MOSFET is increased by $I_{Lm}$; it is much lower than in the hard-switching conditions. Voltage stress of $D_4$ is increased by $V_{CB}$, since $V_{CB}$ is designed to be lower than 100 V, the increased stress has little effect on cost. Voltage stress of MOSFET and current stress of $D_4$ are the same as without the snubber cell 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 turnoff snubber cell.
Fig. 11. Commutation waveforms of freewheeling diode without snubber. (a) Turn-on transients. (b) Turnoff transients.

The proposed general snubber cell consists of one inductor, two capacitors, and three diodes. The snubber inductor is placed in series with the freewheeling diode. It is designed to restrict \( \frac{di}{dt} \) of the reverse-recovery current to achieve ZCS turn-on. The snubber capacitor is placed in parallel with the drain–source voltage. It is designed to restrict \( \frac{dv}{dt} \) of the drain–source voltage to achieve ZVS turnoff. ZVS turn-on and turnoff of the freewheeling diode are also obtained.

Switching losses and EMI noise during turn-on and turnoff are eliminated by the snubber cell. All energy absorbed in the snubber inductor and snubber capacitor are transferred to the buffer capacitor. Energy recovery is achieved by discharging the buffer capacitor to the output. Since the snubber cell deals with only small switching transient energy instead of the main power stage energy, the resulting circulating energy is insignificant compared with resonant converters. 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, Cuk, Sepic, and Zeta with the proposed snubber cells embedded are shown in Fig. 6. Although circuit size and cost are increased by adding these additional components, it is still tolerable, considering the numerous advantages provided by the snubber cell.

IV. EXPERIMENTAL RESULTS

A prototype of a 1-kW 100-kHz boost converter with the passive lossless snubber has been built to verify the principle of operation and the theoretical analysis. Output voltage is regulated at 380 V dc with the control circuit implemented by an L4981A. The components specifications are listed in Table I. A hard-switching boost converter with same specifications is also built for comparison.

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 cell embedded are shown in Figs. 8 and 9, respectively. Waveforms of the MOSFET and the freewheeling diode without snubber are shown in Figs. 10 and 11, respectively. The efficiencies of these two circuits are shown in Fig. 12 under different loadings. The maximum efficiency has been measured to be 96% at 700 W.

Waveforms in Fig. 7(a) and (b) are exactly the same as predicted in Fig. 3. Snubber operation analysis is proved to be valid. Comparing Figs. 8(a) and 10(a), it can be seen that \( \frac{di}{dt} \) of the drain current is restricted and commutation of the MOSFET is close to ZCS turn-on. The reason for no ZCS is the discharge of the parasitic drain–source capacitance of the MOSFET during the turn-on process. This switching loss mechanism can only be removed by resonant converter techniques or active snubbers. Comparing Figs. 8(b) and 10(b), it can also be seen that \( \frac{dv}{dt} \) of the drain–source voltage is restricted and the MOSFET commutates at ZVS turnoff. Fig. 9(a) and (b) shows that the freewheeling diode is also commutated at ZVS turn-on and turnoff. Switching transients of the freewheeling diode are significantly improved compared with the hard-switching counterpart shown in Fig. 11(a) and (b).