1 Introduction

With ever-increasing demand for power from the ac line and more stringent requirements for power quality, power-factor correction (PFC) is becoming an integral part of power supplies. A variety of topologies are available for PFC [1, 2, 3]. While discontinuous-mode converters are well suited for low-power applications, peak-and average-mode controlled boost converters operating in continuous-conduction mode (CCM) are still the primary choice for many medium and high-power applications [1, 4].

Generally, the efficiency and the maximum switching frequency of the conventional PWM CCM boost converter are limited by the reverse-recovery loss of the output rectifier. To reduce or even eliminate this loss, the rate of change of the rectifier current during turnoff must be slowed down. The required rectifier-current wave shaping can be achieved by either active or passive snubbers [5, 6, 7, 8]. Generally, active snubbers consist of an auxiliary switch with a few passive components that are used to control di/dt rate of the rectifier current and to create conditions for zero-voltage switching (ZVS) of the rectifier current and the rectifier [5, 6]. A potential drawback of the active-snubber approaches is a need for an additional switch with current and voltage ratings comparable to
2 Passive Snubber

The circuit diagram of the boost converter with the passive, lossless snubber evaluated in this work is shown in Fig. 1. The snubber consists of inductor $L_D$, capacitors $C_1$ and $C_2$, and diodes $D_1$, $D_2$, and $D_3$. Since inductor $L_D$ is connected in series with transistor $Q$ and rectifier $D$, its value can be used to control the reverse-recovery characteristic of the rectifier by controlling the rectifier di/dt rate [9].

To facilitate the explanation of the operation of the circuit, Fig. 2 show the key waveforms. The waveforms in Fig. 2 are obtained by a Spice simulation assuming that $C_1 = C_2$. The basic principle of operation does not change if the capacitor values are different.

To simplify the analysis of operation, it is assumed that all the components in the circuit are ideal, and that inductance $L_F$ is large enough to be modeled as a constant-current source ($I_{in}$).

Before transistor $Q$ is turned on at $t = T_0$, only diode $D$ is conducting. At the same time, capacitor $C_2$ is fully discharged, i.e., $V_{C2} = 0$, while capacitor $C_1$ is charged up to $V_{C1} = V_0$. When the transistor is turned on at $t = T_0$, inductor current $i_L$ starts increasing. Since during this time rectifier $D$ is still conducting the current difference $I_{in} - i_L$, the voltage across the inductor is clamped to $V_0$, as can be seen from the equivalent circuit shown in Fig. 3(a). As a result, during this topological stage, diodes $D_1$, $D_2$, and $D_3$ are off.

The rate of change of the rectifier current is determined by $V_0/L_D$. By controlling this rate through a proper selection of the $L_D$ value, the recovered charge of the rectifier can be controlled, too. Generally, a slower rate of current change results in a smaller recovered charge [9].

Rectifier $D$ ceases to conduct at $t = T_1$ when current $i_L$ becomes equal to input current $I_{in}$. At the same time, diode $D_2$ starts conducting, and $C_1$, $C_2$, and $L_D$ form a series-resonant circuit, as shown in Fig. 3(b). Due to the presence of diode $D_2$, this circuit can conduct current only in one direction. As a result, during the $T_1 - T_2$ interval, the charge from $C_1$ is transferred to $C_2$ in a resonant manner. The voltage across $C_2$ at the end of the resonant interval is equal to the voltage of $C_1$ at the beginning of the resonant interval. Therefore, at $t = T_2$, $V_{C1} = 0$ and $V_{C2} = V_0$. The peak of the resonant current, which determines the current stress of the transistor is equal to $I_{pk} = V_0/Z_n$, where $Z_n = \sqrt{L_D(C_1 + C_2)/C_1 C_2}$ is the characteristic impedance of the resonant tank.

During the $T_2 - T_3$ interval, all the diodes are off, and constant input current $I_{in}$ is flowing through the switch, as shown in Fig. 3(c). When, at $t = T_3$, the transistor is turned off, the transistor current is diverted...
Fig. 2: Key waveforms of ideal boost converter with passive snubber for $C_1 = C_2$. 
into $C_1$ and $C_2$, as shown in Fig. 3(d). Capacitor $C_1$ starts charging toward $V_0$, while capacitor $C_2$ starts discharging toward zero through conducting diode $D_3$. At $t = T_1$, $V_{C_2} = 0$ and rectifier D starts conducting as shown in Fig. 3(e). At $t = T_0$, the transistor is turned on again, and the circuit enters the next switching cycle.

When the circuit parasitics, primarily diode junction capacitance, are taken into consideration, the basic circuit operation remains unchanged. The most noticeable effect of the parasitics is observed on the rectifier voltage $V_D$, which shows low-energy, high-frequency ringings. These ringings can be easily suppressed by a low-power RC snubber placed across the rectifier.

As can be seen from Fig. 2, the addition of the snubber does not change the voltage stress of the transistor, which is $V_{DS(max)} = V_0$. However, the voltage stress on the diode is $V_{D(max)} = 2V_0$, and it is twice the stress when the circuit is operating without the snubber. On the other hand, the snubber does not change the diode current stress ($I_{D(max)} = I_0$), but increases the transistor current stress from $I_{DS(max)} = I_0$ without the snubber to $I_{DS(max)} = I_0 + V_0/Z_n$ with the snubber.

Since the voltage stress on the diode is independent of the values of the snubber's reactive components $C_1$, $C_2$, and $L_D$, it cannot be reduced by a snubber design op-
timization. The only way to reduce this stress is to place snubber inductor \( L_D \) directly in series with rectifier D rather than with transistor Q. However, this change would double the voltage stress on the transistor.

The current stress of the transistor can be minimized by maximizing characteristic impedance \( Z_n \). Since the value of \( L_D \) is determined from the desired rate of change of the rectifier current, the characteristic impedance can be maximized by minimizing the equivalent series capacitance of \( C_1 \) and \( C_3 \). However, the lower limit on the value of the snubber capacitances is determined by their energy-storage capacity. Namely, capacitor \( C_1 \), at \( V_0 \), ought to be able to store the energy at least equal to the energy in the snubber inductor at the moment of turn-off, i.e., \( 1/2C_1V_0^2 > 1/2L_DL_{in}^2 \). Otherwise, the excessive inductor energy would flow through \( D_1 \), \( D_2 \), and \( D_3 \) directly to the load. This is not desirable, since diodes \( D_1 - D_3 \) may introduce severe reverse-recovery problems.

3 ZVS Active Snubber

Recently, a new class of zero-voltage-transition (ZVT) PWM converters, which overcome some deficiencies of the existing soft-switched converters, has been introduced [6]. By using a resonant network with an auxiliary switch, the converters of this family achieve ZVS for both the main switch and the rectifier without increasing their voltage and current stresses. Due to the soft switching of the rectifier that virtually eliminates the problems related to the reverse recovery, this technique seems suitable for applications in high-power, PFC boost converters. Figure 4 shows the circuit diagram of the ZVT-PWM boost power stage, while Fig. 5 shows its key waveforms [6].

In the ZVT-PWM boost power stage in Fig. 4, an additional resonant network consisting of resonant inductor \( L_R \), auxiliary switch \( S_1 \), and diode \( D_1 \) is added. Due to applying the capacitor-shift rule, rectifier capacitance \( C_D \) can be placed across \( C_R \) so that it becomes part of \( C_R \). To simplify the analysis, the input filter inductance is assumed large enough to be considered as a constant current source.

As shown in Fig. 6, the converter goes through seven topological stages during a conversion cycle. Prior to \( t = T_0 \), main switch S and auxiliary switch \( S_1 \) are off, while rectifier D is conducting. At \( t = T_0 \), \( S_1 \) is turned on and current \( i_L \) starts flowing through \( L_R \). The equivalent circuit during this topological stage is shown in Fig. 6(a). Due to the constant voltage across the inductor, \( i_L \) increases linearly until, at \( t = T_1 \), it reaches \( I_{in} \), when rectifier D is turned off with soft switching.

During the \( T_1 - T_2 \) interval, all semiconductor switches except auxiliary switch \( S_1 \) are off so that \( L_R \) and \( C_R \) form a resonant circuit. \( I_L \) continues to increase in a resonant manner, while \( C_R \) starts discharging. When, at \( t = T_2 \), the voltage across \( C_R \) becomes zero, the anti-parallel diode of main switch S starts conducting, as shown in Fig. 6(c). To obtain ZVS, the turn-on signal should be applied to switch S while its body diode is conducting. During the \( T_2 - T_3 \) interval, \( i_L \) stays constant because the inductor is shorted.

At \( t = T_4 \), switch \( S_1 \) is turned off and \( i_L \) is diverted from \( S_1 \) to \( D_1 \) as shown in Fig. 6(d). Due to the conduction of \( D_1 \), the voltage across switch \( S_1 \) is clamped to \( V_0 \). During this stage, the resonant-inductor cur-
rent decreases linearly, and the inductor energy is transferred to the output. When \( i_L \) becomes zero at \( t = T_4 \), diode \( D_1 \) turns off.

In the following interval, \( T_4 - T_5 \), main switch is on, while all other semiconductors are off. The input current flows through switch \( S \) until the switch is turned off at \( t = T_5 \), when the current is diverted into resonant capacitor \( C_R \) as shown in Fig. 6(f). Because of constant input current \( I_{in} \), the voltage across \( C_R \) increases linearly until it reaches \( V_0 \) at \( t = T_5 \), when rectifier \( D \) starts conducting as shown in Fig. 6(g). This stage is terminated at \( t = T_6 \), when switch \( S_1 \) is turned on and a new conversion cycle is initiated.

As can be seen from the above analysis, the main switch and the rectifier are "softly switched". The main switch is turned on at zero voltage, while the rectifier is turned off at zero voltage. As a result, the turn-on switching loss of the switch is eliminated. In addition, due to the controlled \( \text{di/dt} \) rate and soft switching of the rectifier, its reverse-recovery characteristic is greatly improved. Also, the current and voltage stresses on the main switch and the rectifier are the same as in the conventional,
Fig. 6: Equivalent circuits of active-snubber boost converter topological stages.

In this converter, the auxiliary switch is forced ("hard") commutated. The switch is turned on when the current, which is greater than the input current, is flowing through the switch. Consequently, this switch suffers from the turn-on switching loss and conduction loss. However, since the auxiliary switch conducts only during a brief time around the turn-on transitions of the main switch, it handles a small amount of power compared to the main switch. Therefore, the conduction loss of the switch is relatively small. In addition, if the auxiliary switch is selected with a relatively small output capacitance, its turn-on switching loss is also small. Usually, this relatively small power loss on the auxiliary switch is more than offset by the reduction of the power loss due to the improved reverse-recovery characteristic of the rectifier.

It should be noted that the boost circuit described in [5] has the same features as the above-analyzed converter. However, the implementations of the active snubber circuits slightly differ in these two circuits. The circuit described in this section requires fewer components. In addition, it is only one member of a larger converter family of ZVT-PWM converters. The approach described in [5] is only introduced to alleviate problems associated with the reverse-recovery characteristics of the rectifier in the boost converter.

4 Evaluation Results

The experimental evaluation of the described passive and active snubber was performed on a 600-W, PFC boost converter with average current-mode control. The converter was designed for the 90-260-Vac input-voltage range, and with the output voltage of 380 Vdc. Figure 7 shows the circuit diagram of the experimental converter with the active snubber.
The efficiency of the converter without the snubber (i.e., when $S_1$, $L_R$, $C_R$, $L_S$, and $D_1$ are removed from the circuit) as a function of the input voltage is shown in Fig. 8 for different rectifiers [9, 10]. Rectifiers BYT79-500 and HFA15TB60 are available on the market [9, 10], while the rectifier labeled HEXFRED10 is in the development stage at IR. As can be seen, the ultra-fast HEXFRED10 rectifier results in the best efficiency of the converter, which ranges from 91% at low line up to 98% at high line. The HFA15TB60 rectifier achieves the same efficiency at low line, but slightly lower (97.5%) at high line. The measured efficiency for the BYT79-500, which among the tested diodes has the longest recovery time, was in the 89% - 97.5% range. The measured efficiency of the converter with the active snubber is shown in Fig. 9. Due to a controlled $di/dt$ rate and ZVS of the rectifier, the efficiency of the converter is almost insensitive to the reverse-recover speeds of the rectifiers. The maximum difference in efficiency for the three diodes, which occurs at high line, is less than 0.5%, which is within the measurement error range. It also should be noted that the minimum ef-
Fig. 8: Measured efficiency as a function of input voltage of converter without snubber.

Fig. 9: Measured efficiency as a function of input voltage of converter with active snubber.
Fig. 10: Comparison of measured efficiencies as functions of input voltage of converters with passive, active, and without snubber.

Efficiency that occurs at low line is increased to 92%. This represents approximately 3% increase with respect to the circuit with BYT79-500. The maximum efficiency is not affected by the active snubber. At 98%, it is the same as for the converter without the snubber.

Finally, Fig. 10 shows the measured efficiency of the converter without a snubber and with active and passive snubber. In this evaluation, the HEXFRED10 rectifier is used in all the three circuits. It should be noted that the converter with the passive snubber requires a rectifier with at least 800-V rating (twice the output voltage). Although the HEXFRED10 device is rated at 600 V, we could use this device in our evaluations because our samples exhibited the breakdown voltages in excess of 800 V.

The converter with the passive snubber is obtained from the circuit shown in Fig. 7 by removing the active-snubber components and inserting the following passive-snubber components according to Fig. 1: $C_1 = C_2 = 3.3 \text{ nF}$, $L_D = 2.5 \mu\text{H}$, and $D_1 - D_3 = \text{MUR450}$.

As can be seen, at low line the efficiency of the converter with the passive snubber is only 0.5% lower than that of the converter with the active snubber. At high line the efficiency of the passive-snubber converter is 1% lower than the efficiency of the active-snubber converter. This is mainly due to the snubber loss. However, since the minimum efficiency dictates the thermal design, the same thermal design is required for both the active- and passive-snubber converters.

Although efficiency-wise the ZVT circuit has a slight edge over the converter with the passive snubber, its major advantages are in a much lower voltage stress on the rectifier and virtually ringing-free waveforms. These two features are usually translated into a more reliable operation and smaller-size EMI filters of the ZVT converter.

The number of power-stage components in the ZVT converter and in the converter
with passive snubber is almost the same. As a result, the sizes of the two circuits are similar.

Finally, the cost of the auxiliary switch and its associated gate drive in the ZVT circuit is more than justified by the overall performance improvements at high-power levels. This is especially true for the power levels in the kW range and beyond.

5 Summary

Comparisons between the ZVS passive-snubber technique and the active-snubber technique based on the newly introduced zero-voltage-transition (ZVT) principle were presented. The merits and limitations of these techniques were evaluated with respect to their performance, current and voltage stresses of their semiconductor components, and complexity.

It was shown that both techniques improve the conversion efficiency. The active-snubber approach results in a slightly higher efficiency. However, the main advantages of the active-snubber technique is a much smaller voltage stress on the rectifier and very smooth, ringing-free waveforms. These features make this approach very attractive at high power levels, and more than justify the cost of the auxiliary switch and its associated gate drive.

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References


