Investigation and Implementation of a Passive Snubber with a Coupled-Inductor in a Single-Stage Full-Bridge Boost PFC Converter

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Abstract

In this paper, an improved passive snubber is investigated in a single-phase single-stage full-bridge boost power factor correction (PFC) converter, by which the voltage spike across primary side of the power transformer can be suppressed and the absorbed energy can be transferred to the output side. When compared with the basic passive snubber, the two single-inductors are replaced by a coupled-inductor in the improved snubber. As a result, synchronous resonances in the snubber can be achieved, which can avoid the unbalance of the voltage and current in the snubber. The operational principle of the improved passive snubber is analyzed in detail based on a single-phase PFC converter, and the design considerations of both the snubber and the coupled-inductor are given. Finally, a laboratory-made prototype is built, and the experimental results verify the feasibility of the proposed method and the validity of the theoretical analysis and design method.

Key words: Coupled-inductor, Full-bridge, Passive snubber, Power factor correction (PFC), Single-stage

I. INTRODUCTION

In the research field of power factor correction (PFC), single-stage PFC integrates the functions of PFC and isolated DC/DC conversion into a single power converter, and it has advantages such as high efficiency, simplicity and low cost when compared with two-stage PFC [1], [2]. In recent years, many low power single-stage PFC converters have been investigated. However, fewer high power schemes have been proposed [3].

The isolated full-bridge boost topology is attractive in applications of medium and high power single-stage PFC. The reasons why this type of PFC has not been widely used can be attributed to: 1) an additional starting-up circuit is required to establish an initial output voltage, and 2) there is a voltage spike across the bridge leg caused by the transformer leakage inductor [4], [5]. For normal single-stage full-bridge boost PFC, starting-up has been realized through many efficient methods. For example, flyback starting schemes are proposed in [4]-[6], and a direct starting mode of the converter in the state of no load is presented in [7].

To suppress the voltage spike, a number of techniques have been proposed. Methods based on the basic active clamping technique are introduced in [8]-[10], and they has been the most widely investigated [11]. Two new active clamping techniques were proposed in [12], [13]. A two-switch clamping circuit was presented in [14]. Some active auxiliary circuits with a single-switch are adopted in [11], [15] and [16]. The voltage spike is efficiently suppressed after the adoption of each of the above active methods. However, the active methods above have a common drawbacks, that is: one (or two) additional switch is introduced, which increases the complexity of the control circuit and reduces the reliability of the whole system. Moreover, the switching frequency of the additional switch is two times as high as that of the main switches, so it is difficult to choose the switch. Besides the active methods, some passive methods have also been proposed. For example, LC resonance schemes have been studied in [17]-[19], which
can also achieve soft switching of the main switches. However, its resonance energy can not be transferred to the load. Instead, it is added to the conduction losses of the converter. A RCD snubber is used in [20], but the energy of the snubber circuit is released by the resister. A passive clamping technique is proposed in [21]. However, the problem of the magnetic bias of the power transformer appears after the adoption of the passive clamping circuit. A passive snubber is investigated in [22]. However, after the adoption of the passive snubber, a diode is connected in series with the bridge leg switches, which will reduce the efficiency of the converter. In [23], another passive snubber is investigated in three-phase single-stage PFC which overcomes the disadvantage of the snubber in [22]. However, an unbalance of the voltage and current appears between the two resonant circuits.

In this paper, based on the passive snubber in [23], an improved passive snubber is investigated in a single-phase single-stage full-bridge boost PFC converter. In the improved snubber, the two single-inductors are replaced by a coupled-inductor. Theoretical analysis and experimental results show that the voltage spike can be suppressed efficiently after adoption of the improved snubber. It can also overcome the drawback of the passive snubber in [23].

II. THE PFC CONVERTER AND ITS PRINCIPLE

A. The PFC Converter

A single-phase single-stage full-bridge boost PFC converter is shown in Fig.1, where the improved passive snubber is composed of $C_1$, $C_2$ ($C_1=C_3$), $D_{C1}$, $D_{C2}$, $D_C$ and the coupled-inductor ($L_1=L_2$ are equivalent inductances). $D_{S1}$-$D_{S4}$ and $C_{S1}$-$C_{S4}$ are the parasitic components of switches $S_1$-$S_4$, $L_R$ and $n$ are the equivalent leakage inductance and the voltage ratio of transformer $T$, respectively. The switching mode of $S_1$-$S_4$ is the same as that in [23]. The converter in Fig.1 operates in the continuous current mode (CCM). When the bridge leg switches are shorted ($S_1$ & $S_2$ or $S_1$ & $S_4$ are turning on), the boost inductor $L$ is charged by $u_o$ and the input current increases almost linearly. When the bridge diagonal-leg switches turn on ($S_2$ & $S_3$ or $S_2$ & $S_3$ are turning on), the output current is provided by $u_i$ and $L$, and the input current decreases. The process above is repeated periodically, the input current follows the input voltage, and both PFC and AC/DC conversion can be achieved.

B. Analysis of the Operational Process

The working process of the snubber is closely related to the principle of the PFC converter, and the special control strategy has not been introduced for the snubber. The PFC converter in Fig.1 operates in CCM, which is different than the converter in [23]. As a result, its operational process is presented as followed for further analysis and design.

To simplify the analysis, it is assumed that: 1) all of the devices are ideal, 2) the capacitor $C$ is large enough, so that the output voltage $U_o$ can be considered as a constant value, and 3) during one charging period of $L$, the change of $u_i$ is negligible because the charging period is much shorter than the line period. The following analysis is during one charging period of $L$ when $u_i>0$. The theoretical waveforms and the equivalent circuits of the different stages are shown in Fig.2 and Fig.3, respectively.

Stage 1 (before $t_o$): $S_1$ and $S_3$ are turning on, while $S_2$ and $S_4$ are turning off. The voltage across the primary side of transformer $T$ is $U_{1k}=nU_o$, and $U_{CL1}=U_{CL2}=nU_o/2$, $U_{CL3}=U_{CL4}=nU_o/2$, and $U_{CS1}=U_{CS2}=0$. The current in $L$ ($i_L$) flows through $S_3$, $S_2$, and $T$ to the load, and it decreases. On the secondary side of $T$, $D_{O1}$ and $D_{O4}$ are turning on, and $D_{O2}$ and $D_{O3}$ are turning off.

Stage 2 ($t_o$-$t_1$): At $t_o$, $S_1$ turns on, and $S_3$ turns off. $L$ is charged by $u_o$, and $i_L$ increases linearly. In the snubber, $C_1$ is resonant with $L_1$ through $D_{C1}$, $S_1$, and $S_2$. Furthermore, $C_2$ is resonant with $L_2$ through $S_1$, $S_3$, and $D_{C2}$. The current of $L_R$ ($i_{L_R}$) can not be mutated, so $i_{L_R}$ decreases immediately. At $t_1$, $i_{L_R}$ reduces to zero (the excitation current of $T$ is not considered here), and $D_{O1}$ and $D_{O4}$ are turned off. The inductance of $L_R$ is very small, so the duration of this stage can be ignored.

Stage 3 ($t_1$-$t_2$): The resonances in stage 2 are continuous. The voltage of $C_1$ and $C_2$ and the current in $L_1$ and $L_2$ are:

$$u_{C1/2}(t)=\frac{nU_o}{2}\cos\frac{1}{\sqrt{L_1C_1}}(t-t_1)$$

$$i_{L1/2}(t)=\frac{nU_o}{2}\sqrt{C_1}\sin\frac{1}{\sqrt{L_1C_1}}(t-t_1)$$

At $t_2$, $U_{CL1}=U_{CL2}=0$, and the energy of $C_1$ and $C_2$ is transferred to $L_1$ and $L_2$ entirely. The duration of this stage is calculated as:

$$t_2=\frac{\pi}{2}\sqrt{L_1C_1}$$

Stage 4 ($t_2$-$t_3$): In this stage, $i_L$ still increases linearly and the output current is provided by capacitor $C$ alone. The voltage of $C_1$ or $C_2$ is zero, so that diode $D_2$ is turned on, $L_1$ is connected in series with $L_2$, and their current flows through $D_{C1}$, $S_1$, $S_2$, $D_{C2}$ and $D_C$. During stage 1-4, $U_{CS3}$ is still zero, so that $S_3$ turns off with zero voltage at $t_o$.

Stage 5 ($t_3$-$t_4$): At $t_3$, $S_2$ turns off, and $S_3$ turns on with zero voltage. $C_{S2}$, $C_{S3}$, $C_1$ and $C_2$ are charged by $L_1$, $L_2$, and $L$. The voltage across the bridge leg increases from zero, so that $S_2$...
turns off with zero voltage. The capacitance of \( C_{S1} \sim C_{S4} \) is much small than that of \( C_1 \) and \( C_2 \). Therefore, only the charging process of \( C_1 \) and \( C_2 \) is considered in the following calculation. In this stage, the following relationships can be obtained:

\[
C_1 \frac{d u_{C1/C2}(t)}{dt} = i_{L1}(t_3) + i_{L1/L2}(t)
\]

\[
u_{C1/C2}(t) = -L_1 \frac{d i_{L1/L2}(t)}{dt}
\]

From (4) and (5), the following differential equation is obtained:

\[
L_1 C_1 \frac{d^2 u_{C1/C2}(t)}{dt^2} + u_{C1/C2}(t) = 0
\]

This equation (6) has the following initial data:

\[
u_{C1/C2}(t_3) = 0
\]

\[
i_{L1/L2}(t_3) = \frac{n U_o}{C_1} \frac{C_1}{\sqrt{L_1}}
\]

Therefore, the voltage expression of \( C_1 \) or \( C_2 \) and the current expression of \( L_1 \) or \( L_2 \) can be obtained:

\[
u_{C1/C2}(t) = \left[ \frac{C_1}{L_1} i_{L1}(t_3) \sin \frac{t}{2 \sqrt{C_1 L_1}} \right]
\]

\[
i_{L1/L2}(t) = i_{L1}(t_3) \cos \frac{t}{2 \sqrt{C_1 L_1}} - \frac{n U_o}{C_1} \frac{C_1}{\sqrt{L_1}} \cos \frac{t}{2 \sqrt{C_1 L_1}}
\]

At \( t_3 \), the charging process of \( C_1 \) and \( C_2 \) is over. Therefore, \( U_{S1} = n U_o \), \( U_{C1} = U_{C2} = n U_o / 2 \), \( U_{C3} = U_{C4} = 0 \), and \( U_{CS1} = U_{CS4} = n U_o \). In this stage, the output current is only provided by capacitor \( C \). The inductance of \( L \) is large enough, so that the change of \( i_L \) can be ignored during this stage. During the whole line period, the value of \( i_{L1}(t_3) \) is varying, so that the duration of this stage is different during the whole line period.

Stage 6 (\( t_2 - t_3 \)): In this stage, the current of \( L_1 \), \( L_2 \) and \( L \) flows through \( S_3 \), \( S_4 \) and \( T \) to the load, and then it decreases. On the secondary side of \( T \), \( D_{02} \) and \( D_{03} \) turn on. In this stage, the expression of \( i_{L1} \) and \( i_{L2} \) is:

\[
i_{L1/L2}(t) = i_{L1/L2}(t_4) - \frac{n U_o}{2 L_1} (t - t_4)
\]

Fig. 2. Theoretical waveforms.

Fig. 3. Equivalent circuit of each stage.
At \( t_s \), \( i_{t1} \) and \( i_{t2} \) reduce to zero. The inductance of \( L_b \) is so small that the rising process of \( i_{t,b} \) is ignored in this stage.

Stage 7 (\( t_o-t_b \)): In this stage, \( i_t \) flows through \( S_1 \), \( S_4 \) and \( T \) to the load, and it still decreases, which is similar to that in stage 1.

After \( t_o \), the converter operates in another charging period, and the switching state between \( S_1 \) and \( S_3 \), and \( S_2 \) and \( S_4 \) are exchanged.

C. Analysis of Parameters in the snubber

According to the analysis in [23] and the analysis above, the voltage and current stress of switches \( S_1 \sim S_4 \) in Fig.1 can be calculated:

\[
U_S = nU_o + i_L(t) = \frac{2L_{th}}{C_1} \sin \frac{\sqrt{2} \pi}{\sqrt{L_{th}C_1}} \left( \frac{\sqrt{2} \pi}{\sqrt{L_{th}C_1}} \right)
\]

(12)

\[
I_S = I_{L_{max}} + 2t_{L1/2}(t_2) = I_{L_{max}} + nU_o \left( \frac{C_1}{L_1} \right)
\]

(13)

The duration of stage 2 is so small that \( i_{t,b} \) is ignored in the calculation of the current stress of \( S_1 \) or \( S_3 \). It can be seen that \( U_S \) decreases as \( C_1 \) increases, and that \( I_S \) increases as \( C_1 \) increases or \( L_1 \) decreases.

According to the analysis in [23], the first limitation of \( L_1 \) and \( C_1 \) can be obtained:

\[
L_1 C_1 \leq 2.87D_{min}^2 T_r^2
\]

Where \( D = (t_f-t_o)/T \) is the duty cycle of the PFC converter, and \( T \) is the charging period of \( L \).

In [23], the three-phase PFC converter operates in discontinuous current mode (DCM). Generally, its duty cycle is within 50%, so the current in \( L_1 \) or \( L_2 \) must reduce to zero during the phase when the bridge diagonal-leg switches are turning on. However, in this paper, the single-phase PFC converter operates in CCM. Therefore, the maximum duty cycle will be much more than 50%, so another constraint condition of \( L_1 \) and \( C_1 \) must be analyzed to make sure that the current in \( L_1 \) and \( L_2 \) can reduce to zero in order to avoid the magnetic saturation of \( L_1 \) and \( L_2 \) after several charging periods.

The current in \( L_1 \) and \( L_2 \) begins to decrease during stage 5, and it can be seen from (10) that \( i_{t1/2}(t) \) will decrease faster as \( i_{t1}(t) \) increases. It also needs to be made certain that \( i_{t1/2}(t) \) can reduce to zero when \( i_{t2}(t) \) is zero. In that case, the energy of \( L_1 \) and \( L_2 \) will be transferred to \( C_1 \) and \( C_2 \) entirely. Therefore, the following relationships can be obtained:

\[
\frac{\pi}{2} \sqrt{L_1 C_1} \leq (1 - D_{max}) T
\]

(15)

From (15), the second limitation of \( L_1 \) and \( C_1 \) can be obtained:

\[
L_1 C_1 \leq \frac{4(1-D_{max})^2 T_r^2}{\pi^2}
\]

(16)

From the analysis above, it can be seen that the voltage stress of \( D_{C1} \) or \( D_{C2} \) is equal to that of \( C_1 \) or \( C_2 \), which is \( U_o/2 \).

It can also be seen that the voltage stress of \( D_{C} \) is \( U_o \), the current stress of \( D_{C1} \) or \( D_{C2} \) is equal to that of \( L_1 \) or \( L_2 \), which can be obtained from (8), and the current stress of \( D_{C} \) is equal to that of \( S_1 \sim S_4 \), which can be obtained in (13).

III. DESIGN CONSIDERATIONS OF THE COUPLED-INDUCTOR

A. Mechanism Analysis of the Coupled-Inductor

In section II, it was proposed that \( C_{1}=C_{2} \) and \( L_{1}=L_{2} \). Therefore, the resonances between \( C_{1} \) (or \( C_{2} \)) and \( L_{1} \) (or \( L_{2} \)) are synchronous. However, there are some differences between two capacitors with the same capacitance. This is due to their tolerance features. For the same reason, the inductances of two inductors with the same configuration are not exactly the same. Therefore, the resonances between \( C_{1} \) (or \( C_{2} \)) and \( L_{1} \) (or \( L_{2} \)) are asynchronous, which would result in an unbalance of the voltage and current between \( C_{1} \) and \( C_{2} \), \( L_{1} \) and \( L_{2} \), and \( D_{C1} \) and \( D_{C2} \). In actual fact, to avoid over voltage on the capacitors \( C_{1} \) and \( C_{2} \) and the diodes \( D_{C1} \) or \( D_{C2} \) and to avoid saturation of the inductors \( L_{1} \) or \( L_{2} \), capacitors and diodes with a higher voltage or current stress and inductors with a larger magnetic core volume must be considered.

To resolve this problem, the coupled-inductor is adopted. Here, the two inductors \( L_{1} \) and \( L_{2} \) are made on a common magnetic core, and they have a common magnetic circuit and the same number of turns. Therefore, the difference in the inductances of the two inductors can be ignored. As a result, the asynchronous resonances brought from the difference between the two inductors have not been considered.

The resonances between \( C_{1} \) (or \( C_{2} \)) and \( L_{1} \) (or \( L_{2} \)) appear in stage 2, 3 and 5. The following analysis is based on the circuit models in Fig.4 and Fig.5, which show the equivalent circuit of the passive snubber in stage 2, 3 and 5. The model of the coupled-inductor is made up of \( L_{k1} \), \( L_{m1} \), \( L_{a2} \), \( L_{m2} \), and \( T_{ideal} \) where \( L_{k1} \) and \( L_{a2} \) are the leakage inductance, \( L_{m1} \) and \( L_{m2} \) are the excitation inductance (\( L_{k1}+L_{m1}=L_{k1} \), \( L_{a2}+L_{m2}=L_{a2} \)), and \( T_{ideal} \) is the ideal transformer. The two inductors have a common magnetic circuit and the same number of turns, so the difference in the inductance between them can be ignored. Therefore: \( L_{k1}=L_{a2} \) and \( L_{m1}=L_{m2} \). Here, it has been defined that:

\[
L_{m1} = aL_{1}, \quad L_{k1} = (1-a)L_{1}
\]

(17)

Where, \( 0<\alpha<1 \).

For Fig.4, it has been defined that the time \( t_{54} \) (or \( t_{25} \)), before \( t_{54} \), \( U_{C1}=U_{C2} \) and \( i_{t1}=i_{t2} \). It is assumed that a difference appears between \( U_{C1} \) and \( U_{C2} \) after \( t_{54} \):

\[
U_{C2}(t_{54}) = U_{C1}(t_{54}) + \Delta U
\]

(18)

Where \( \Delta U>0. \)

After \( t_{54} \), the following relationships are obtained:
The following relationship can be obtained approximately:

\[
\begin{align*}
W & = \frac{1}{2} \left( L_1 + L_2 \right) M \left( i_{11} + i_{12} \right) \\
& + \int_{t_0}^{t} \frac{U_{C1}(t)}{L_1} \, dt + \int_{t_0}^{t} \frac{U_{C2}(t)}{L_2} \, dt
\end{align*}
\]

From (19) and (20), the following is obtained:

\[
U_{T1}(t) = \frac{1}{2} \left( U_{C1}(t) + U_{C2}(t) \right)
\]

(21)

From (18), (19), (21), the following expression is obtained:

\[
\begin{align*}
i_{11}(t) &= i_{11}(t_0) + \int_{t_0}^{t} \frac{(1-a)U_{C1}(t) - \frac{1}{2} \Delta U}{(1-a)L_1} \, dt \\
i_{12}(t) &= i_{12}(t_0) + \int_{t_0}^{t} \frac{(1-a)U_{C1}(t) + \Delta U - \frac{1}{2} a \Delta U}{(1-a)L_2} \, dt
\end{align*}
\]

(22)

From (22), it can be seen that after \( t_{0} \), \( i_{11} \) and \( i_{12} \) can help accelerate the discharging of \( C_1 \). Furthermore, the following expression can be obtained:

\[
\Delta I = i_{12}(t) - i_{11}(t) = \int_{t_0}^{t} \frac{(1+a)\Delta U}{(1-a)L_1} \, dt
\]

(23)

From (23), it can be seen that \( \Delta I \) increases as the value of \( a \) increases, and that the synchronous changing of between \( U_{C1} \) and \( U_{C2} \) can be achieved more easily.

After \( t_{0} \), the expression of \( i_{11} + i_{12} \) can be obtained from (20) and (21):

\[
i_{11}(t) + i_{12}(t) = i_{11}(t_0) + i_{12}(t_0) + \int_{t_0}^{t} \left[ \frac{U_{C1}(t)}{L_1} + \frac{U_{C2}(t)}{L_2} \right] \, dt
\]

(24)

It can be seen from (24) that \( i_{11} + i_{12} \) is independent of \( \Delta U \). Therefore, the magnetic core excitation of the coupled-inductor can not be affected by the value \( \Delta U \).

The same analysis procedure can also be suitable in Fig. 5, which does not have to be repeated.

**B. Design of the Coupled-Inductor**

According to the analysis above, the configuration scheme of the coupled-inductor is designed as shown in Fig. 6, where the two inductors are made on a common magnetic core, and they have the same number of turns.

From the coupled-inductor theory, the basic mathematical model of the coupled-inductor is:

\[
\begin{align*}
u_{L1} &= L_{11} \frac{di_{11}}{dt} + M \frac{di_{12}}{dt} \\
u_{L2} &= L_{22} \frac{di_{12}}{dt} + M \frac{di_{11}}{dt}
\end{align*}
\]

(25)

Where \( L_{11} = L_{22} \) are the self inductances and \( M \) is the mutual inductance.

The two inductors have a common magnetic circuit, so the following relationship can be obtained approximately:

\[
M = L_{11} = L_{22}
\]

(26)

If the asynchronous resonances in the snubber is ignored, the following can be obtained: \( u_{L1} = u_{L2} \) and \( i_{11} = i_{12} \). Therefore, the following relationship can be obtained approximately:
single
coupled

Under real conditions, the resonances in a snubber with two single-inductors are asynchronous, so the maximum currents of \( L_1 \) or \( L_2 \) increase. However, the maximum current of the coupled-inductor has not been changed. Therefore, under real conditions, the volume of the snubber with a coupled-inductor is smaller than that of the snubber with two single-inductors.

IV. EXPERIMENTAL VERIFICATIONS

To verify the theoretical analysis and the evaluations mentioned above, a laboratory-made prototype of the converter was built, in which the average current mode control strategy was adopted. The basic circuit parameters and the main utilized component types are: \( u_i=110V \text{rms} \pm 10\% \), \( U_{dc}=100V \text{dc} \), \( P_{out}=500W \), \( L=0.58mH \), \( C=1000\mu F \), \( L_s=6\mu H \), \( n=2 \), and \( S_1-S_4 \); 24N60C3 (Infineon, \( R_{on}=0.16\Omega \), with a switching frequency of about 37kHz). In the snubber, \( C_1=C_2=44\mu F \pm 10\% \) (the two capacitors CBB223K are connected in parallel), and \( L_{11}=L_{22}=38\mu H \) (EL28, \( L_1=L_2=76\mu H \), the value of \( L_{k1} \) or \( L_{k2} \) is within 2\( \mu H \), and \( a>0.97 \)).

Fig. 7 shows the input waveforms of the PFC converter. Table 1 shows the experimental data for the power factor (PF), the efficiency (\( \eta_c \)) and the efficiency when two single-inductors are adopted (\( \eta_s \)) according to the output power (\( P_o \)) variation. When two single-inductors are adopted, asynchronous resonances occur in the snubber. As a result, the experiment is only within 300W to protect the circuit. It can be seen that a good PFC effect has been achieved, and that the prototype shows good performance in conversion efficiency. Fig.8 shows the experimental waveforms of the switches \( S_1 \) and \( S_2 \). It can be seen that \( S_1 \) turns off with zero voltage, and \( S_2 \) turns on and off with zero voltage. The switching states of \( S_1 \) and \( S_2 \) are the same as those of \( S_1 \) and \( S_2 \). Therefore, the related experimental results are not presented here.

To verify the above analysis, two single-inductors with the same structure have been constructed (El25, where the inductance \( L_1=L_2=76\mu H \)). Fig.9 and 10 are the experimental results (the efficiency data is shown in Table 1). To protect the circuit, the results of Fig.9 and 10 are obtained under a relative low bus voltage.

Fig. 9 shows the voltage waveforms of \( C_1 \) and \( C_2 \), when the two single-inductors are adopted. It can be seen that regardless of whether \( C_1=C_2 \) or \( C_1 \neq C_2 \), the voltage between \( C_1 \) and \( C_2 \) is unbalanced, which is the same as saying that the resonances between \( C_1 \) (or \( C_2 \)) and \( L_1 \) (or \( L_2 \)) are asynchronous. Furthermore, the following conclusion can be obtained from Fig.9: the asynchronous resonances in the snubber become more serious as the difference of between \( C_1 \) and \( C_2 \) increases.

Fig.10 shows the voltage waveforms of \( C_1 \) and \( C_2 \), when the coupled-inductor is adopted. It can be seen that regardless of whether \( C_1=C_2 \) or \( C_1 \neq C_2 \), the voltage between \( C_1 \) and \( C_2 \) is balanced, which is the same as saying that the resonances between \( C_1 \) (or \( C_2 \)) and \( L_1 \) (or \( L_2 \)) are synchronous after the coupled-inductor is adopted. Fig.9 and 10 prove the analysis in section III A. Furthermore, comparing the experimental results in Fig.9 (c) with those in Fig.10 (c), it can be seen that the charging and discharging time of the capacitors are approximately equal when two single-inductors and a coupled-inductor are adopted, respectively. From this, it can

![Fig. 7. Input voltage and current.](image)

![TABLE I EXPERIMENTAL DATA](image)

![Fig. 8. Soft switching waveforms.](image)
be seen that the analysis of the self inductance of the coupled-inductor in section III B is verified.

V. CONCLUSIONS

In this paper, an improved passive snubber with a coupled-inductor is investigated based on a single-phase single-stage full-bridge boost PFC converter. The theoretical analysis and experimental results show that: 1) The adoption of this snubber can realize both the suppression of the voltage spike across the primary side of the power transformer and the energy transfer from the snubber itself to the load, and 2) The use of a coupled-inductor in the improved snubber to replace the two single-inductors can help achieve the synchronous resonances in the snubber. It can also help avoid the unbalance of the voltage and current among the devices of the snubber.

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REFERENCES


