Lossless Snubber for GaN-Based Flyback Converter with Common Mode Noise Consideration

Y. T. Yau1,2, Member, IEEE, and Tsung- Liang Hung3

1 Department of Ph.D. Program, Prospective Technology of Electrical Engineering and Computer Science, National Chin-Yi University of Technology, Taichung, Taiwan
2 Department of Electrical Engineering, National Chin-Yi University of Technology, Taichung, Taiwan
3 Asian Power Device Inc., Taoyuan, Taiwan

Corresponding author: Y. T. Yau (e-mail: pabloyau@ncut.edu.tw).

This work was supported technically by Asian Power Device Inc.

ABSTRACT The previous lossless snubber for flyback converters, although with a small number of components, simple control and high power conversion efficiency, has no consideration for electromagnetic interference (EMI). This paper proposes a lossless snubber converter with a novel drain voltage rising slew rate control for the main switch of the flyback converter. Compared to traditional topologies, the proposed snubber topology has two advantages: first, only an additional inductor is used. Second, the proposed method can reduce EMI emissions significantly. Finally, a prototype of the proposed snubber is built and verified with two existing similar snubbers.

INDEX TERMS Lossless snubber, energy regenerative snubber, EMI, converter, common mode noise.

I. INTRODUCTION

Ideally, the transformer is perfectly coupled without leakage flux between the windings, but the actual transformer always has leakage inductance in its windings. Without a suitable voltage suppression mechanism, leakage inductance can induce a voltage spike over the rating voltage of the power switches, the power switch may be damaged. In order to prevent the power switch from being damaged, a snubber circuit is applied to suppress the voltage spike. The snubber is usually located on the drain of the power switch in Fig. 1. The typical dissipative circuit includes a resistor, a diode, and a capacitor called RCD snubber [1-2]. The RCD snubber has been widely used in industry for a long time.

Due to the low switching loss and high switching frequency of gallium nitride (GaN) power switch, there is a strong interest in emerging power electronics converters. However, one of the concerns is that the high switching speed (\(dv/dt\) or \(di/dt\)) of GaN devices can generate more EMI emissions [3] compared to conventional silicon-based power switches. The rapid drain voltage slew rate on the main switch is an important source of EMI common mode noise. The snubber can be passive [5]-[7] or active [8], or a mixed active and passive [9]. It can also be dissipative [10]-[12] or non-dissipative [5]-[9], [12]-[15]. According to the conclusions in
[4], the methods of controlling turned-off dv/dt have a better effect on reducing CM current of a flyback converter, such as [9], [13]-[15], but they use more active switches and passive components, so the circuits are too complicated.

The proposed circuit is shown in Fig. 2, which is a lossless snubber topology. Compared to the basic flyback topology, the proposed circuit has the following two features: first, it needs one additional inductor and one winding. Second, it can be driven by the commercial flyback control IC without any modification. Thus, the proposed circuit is easily commercialized for industrial applications.

There are two similar existing lossless snubber topologies shown in Fig. 3. Fig. 3(a) is the lossless LC snubber, which can recycle the energy of leakage inductance [16] to improve the efficiency. Compared with the first prior topology in Fig. 3(a), the proposed circuit adds an auxiliary winding to the main transformer. As shown in Fig. 3(b), the second prior circuit, named energy regenerative snubber [17]-[20], has the same components as the proposed circuit. The energy regenerative snubber is basically a lossless snubber. With the auxiliary winding of the transformer, the energy stored in the leakage inductor of the transformer can be recycled back to the input capacitor.

The above two existing snubber topologies are similar to the proposed circuits, and have similar operating principles, so these two circuits are described together in this paper. However, the difference from the previous literature is that the previous research did not discuss the behavior of reverse recovery of snubber diodes. However, it is found that the effect of reverse recovery cannot be ignored, and there is still valuable for discussion in this paper.

![FIGURE 2. The proposed lossless snubber.](image)

![FIGURE 3. Two types of conventional lossless snubber: (a) Lossless LC snubber [16]; (b) Energy regenerative snubber [17]-[20].](image)

The definition list of symbols and variables in this paper is shown in Table I.

| Symbol and variable | Definition |
|---------------------|------------|
| $V_{in}$            | Input voltage |
| $V_D$               | Output voltage |
| $Q_1$               | Main controller chip |
| $Q_2$               | Main switch with GaN FET |
| $Q_p$               | Synchronous rectifier with Si-based MOSFET |
| $T_1$               | Main transformer |
| $N_p$               | Primary winding of $T_1$ |
| $N_r$               | Secondary winding of $T_1$ |
| $T_{rr}$            | Resonant winding of $T_1$ |
| $L_m$               | Magnetizing inductance of $N_r$ of $T_1$ |
| $L_{LK1}$           | Leakage inductance of $N_r$ of $T_1$ |
| $L_{LK2}$           | Leakage inductance of $N_r$ of $T_1$ |
| $C_r$               | Resonance capacitor of snubber |
| $Z_r$               | Impedance of resonant tank of $L_r$ and $C_r$ |
| $Z_1$               | Impedance of resonant tank of $L_{LK1}$, $C_r$, and $C_{in1}$ |
| $Z_2$               | Impedance of resonant tank of $L_{LK2}$ and $C_r$ |
| $f_{res}$           | Resonance frequency of resonant tank of $L_r$ and $C_r$ |
| $\omega_1$         | Resonance frequency of resonant tank of $L_{LK1}$, $C_r$, and $C_{in1}$ |
| $\omega_2$         | Resonance frequency of resonant tank of $L_{LK2}$ and $C_r$ |
| $T_{res}$          | Resonant period of resonant tank of $L_r$ and $C_r$ |
| $T_1$              | Resonant period of resonant tank of $L_{LK1}$, $C_r$, and $C_{in1}$ |
| $C_{out}$           | Output capacitor |
| $D_1$              | Diode of snubber |
| $D_2$              | Diode of snubber |
| $C_{in1}$          | Output capacitance of $Q_1$ |
| $C_{in2}$          | Output capacitance of $Q_2$ |
| $D_{sw}$           | Body diode of $Q_1$ |
| $I_o$              | Switching frequency |
| $t_{sw}$          | Switching period |
| $D$               | Duty cycle |
| $n_m$             | Turns ratio of $N_r$ to $N_p$ |
| $n_{in}$          | Turns ratio of $N_r$ to $N_{in}$ |
| $n_{inr}$         | Turns ratio of $N_r$ to $N_{in}$ |
| $i_{Lm}$         | Current of $L_m$ |
| $i_{Cp}$         | Current of $C_{in}$ |
| $i_{Cp2}$        | Current of $C_{in2}$ |
| $i_{L_{LK1}}$   | Current of $L_{LK1}$ |
| $v_{ds}$        | Drain-source voltage of $Q_1$ |
| $v_{ds}^{(max)}$ | Maxima drain-source voltage of $Q_1$ |
| $i_{ds}$        | Drain-source current of $Q_1$ |
| $i_{ds}^{(max)}$ | Maxima drain-source current of $Q_1$ |
| $i_{C_p}$       | Current across $C_p$ |
| $v_{C_p}$       | Voltage across $C_p$ |
| $v_{C_{in}}$    | Maxima voltage of $C_{in}$ |
| $v_{C_{in}}^{(min)}$ | Minima voltage of $C_{in}$ |

Table I: Definition list of symbols and variables

After the first introduction section, the remainder of this paper is organized as follows. Section II describes the proposed circuit. Sections III and IV introduce the operation behavior of the lossless LC snubber and the energy regenerative snubber in the flyback converter. Then, in Section V, the specifications of these three prototypes are mentioned. The design process of the proposed circuit in
Section VI. The results of three prototype circuits are discussed and analyzed in Section VII. The performance of the proposed circuit is compared with the others to show its feasibility. Finally, the conclusions are drawn in Section VIII.

Before entering the following circuit discussion in Session II, III, and IV, there are several assumptions as follows.

1) These circuits operate in CCM.
2) The polarity of the $C_1$ is positive at initial voltage.
3) $D_1$ and $D_2$ have no initial current with non-ideal diode behavior of forward conduction voltages $V_{F,DI}$ and $V_{F,DO}$, reverse recovery times $t_{rv1}$, $t_{rv2}$ and reverse recovery charge $q_{rv1}$, $q_{rv2}$ respectively.
4) The turn-on resistance of $Q_2$ is so low that it can be regarded as an ideal rectifier.
5) $Q_1$ has parasitic output capacitor $C_{os1}$.
6) The leakage inductance of the winding $N_p$ is represented by $L_{LK1}$, and the transformer magnetizing inductance is expressed by $L_m$.
7) The leakage inductance of the winding $N_r$ is represented by $L_{LK2}$. $L_{LK2}$ can be ignored in the proposed circuit and lossless LC snubber, because the $L_r$ is much larger than $L_{LK2}$. However, there is no physical $L_r$ in energy regenerative snubber, and $L_{LK2}$ has to be considered.

II. PROPOSED CIRCUIT CONFIGURATION

As shown in Fig. 2, the proposed snubber stores leakage inductance energy when $Q_1$ is turned off. When $Q_1$ is turned on, the $N_r$, $L_r$, $D_2$, and $C_r$ constitute a resonant tank, which regenerates the energy of the leakage inductor $L_{LK1}$ to the input terminal. $N_r$ provides a reversed DC superposition voltage on $C_r$ for the soft switching function when $Q_1$ is turned off. In this section, a detailed analysis of the operation states will be carried out. Fig. 4 shows the waveform of each point of the circuit. It can be seen that in the state 1 ($t_r$–$t_0$) interval, $v_{C_r}$ will oscillate from positive polarity to negative polarity. Therefore, in state 4 ($t_r$–$t_4$), $v_{d1}$ can change from a low $v_{clamp}$ and gradually increase upward.

Fig. 5 is an explanation of the behavior of the proposed circuit, which can be divided into 7 operation states. The relative parameters of the proposed circuit are defined as follows.

$$\omega_0 = \frac{1}{\sqrt{L_r \cdot C_r}}$$  \hspace{1cm} (1)

$$T_0 = 2\pi \sqrt{L_r \cdot C_r}$$  \hspace{1cm} (2)

$$Z_0 = \frac{L_r}{\sqrt{C_r}}$$  \hspace{1cm} (3)

$$\omega_1 = \frac{1}{\sqrt{(C_r + C_{os1}) \cdot L_{LK1}}}$$  \hspace{1cm} (4)

$$T_1 = 2\pi \sqrt{L_{LK1} \cdot (C_r + C_{os1})}$$  \hspace{1cm} (5)

$$Z_1 = \frac{L_{LK1}}{\sqrt{C_r + C_{os1}}}$$  \hspace{1cm} (6)

$$\omega_2 = \frac{1}{\sqrt{C_r \cdot L_m}}$$  \hspace{1cm} (7)

$$Z_2 = \frac{L_m}{\sqrt{C_r}}$$  \hspace{1cm} (8)

$$i_{D2} = i_{Lr} = i_{N_r}$$  \hspace{1cm} (9)

$$i_{LK1} = i_{ps} + i_{Lm}$$  \hspace{1cm} (10)

$$D = \frac{V_o \cdot n_{ps}}{V_{in} + V_o \cdot n_{ps}}$$  \hspace{1cm} (11)

$$T_{sw} = D \cdot T_{sw}$$  \hspace{1cm} (12)

$$n_{ps} = \frac{N_p}{N_i}$$  \hspace{1cm} (13)

$$n_{pr} = \frac{N_p}{N_r}$$  \hspace{1cm} (14)

$$n_{pr} = \frac{N_i}{N_r}$$  \hspace{1cm} (15)

And, the initial voltage of $C_r$ is defined as (16).

$$v_{C_r}(t_0) = V_o \cdot n_{ps} \cdot \frac{Q_{rl}}{C_r} + i_{LK1} \sqrt{\frac{L_{LK1}}{C_{os1} + C_r}} - V_{F,DI}$$  \hspace{1cm} (16)

1) State 1: As shown in Fig. 5(a), $Q_1$ is turned on. Because $i_{Lm}$ has an initial value, $i_{di1}$ rises rapidly, and $V_{ps}$ continues to excite $L_m$ in the forward direction, causing $i_{Lm}$ to continue to rise as (18). At the same time, $C_r$ resonates with $L_r$. At this time, $D_2$ is forward-biased conduction, and reaches the lowest point of $v_{C_r}(t_1)$ after half a resonant period, as shown in (24). In addition, $i_{di1}$ and $i_{ps}$ are zero.

$$i_{Lm}(t) = i_{LK1}(t) = i_{Ls}(t_0) + \frac{V_{in}}{L_m}(t-t_0)$$  \hspace{1cm} (17)

$$i_{Lm}(t_1) = i_{LK1}(t_1) = i_{Ls}(t_1) + \frac{V_m}{L_m}(t_1-t_0)$$  \hspace{1cm} (18)
\[ i_{Lr}(t) = i_C(t) = i_{Lr}(t_0) \cdot \cos \omega_b(t - t_0) \]
\[ = \frac{V_m}{n_{pr}} \cdot \sin \omega_b(t - t_0) \]
(19)

\[ i_{Lr}(t) = i_C(t) = i_{Lr}(t_0) \cdot \cos \omega_b(t - t_0) \]
\[ = \frac{V_m}{n_{pr}} \cdot \sin \omega_b(t - t_0) \]
(20)

\[ v_{C}(t) = \frac{V_m}{n_{pr}} \cdot \sin \omega_b(t - t_0) \]
\[ + \left( \frac{Z_0 \cdot i_{Lr}(t_0)}{n_{pr}} \right) \cdot \sin \omega_b(t - t_0) \]
(21)

\[ v_C(t) = \frac{V_m}{n_{pr}} \cdot \sin \omega_b(t - t_0) \]
\[ + \left( \frac{Z_0 \cdot i_{Lr}(t_0)}{n_{pr}} \right) \cdot \sin \omega_b(t - t_0) \]
(22)

\[ i_{q_1} = \frac{1}{p} \cdot \frac{V_m}{n_{pr}} \cdot \sin \omega_b(t - t_0) \]
\[ + \left( \frac{Z_0 \cdot i_{Lr}(t_0)}{n_{pr}} \right) \cdot \sin \omega_b(t - t_0) \]
(23)

\[ i_{q_1} = \frac{1}{p} \cdot \frac{V_m}{n_{pr}} \cdot \sin \omega_b(t - t_0) \]
\[ + \left( \frac{Z_0 \cdot i_{Lr}(t_0)}{n_{pr}} \right) \cdot \sin \omega_b(t - t_0) \]
(24)

\[ i_{d_1}(t) = i_{Lr}(t) + i_{L}(t) \]
\[ = 0 \]
(25)

\[ i_{d_1}(t) = 0 \]
\[ = 0 \]
(26)

\[ i_{d_2}(t) = i_{Lr}(t) \]
\[ = 0 \]
(27)

\[ i_{N_1}(t) = 0 \]
\[ = 0 \]
(28)

\[ \Delta t_2 = t_2 - t_1 = \frac{T_m}{2} \]
(29)

2) **State 2:** As shown in Fig. 5(b), \( Q_1 \) is turned on, and \( V_{in} \) continues to magnetize \( L_m \) in the forward direction, causing \( i_{Lm} \) to continue to rise as (31). On the other hand, \( C_r \) resonates with \( L_r \) for half a resonant period. Due to the existence of non-ideal \( D_2 \) with \( Q_{r2} \) and \( r_{r2} \), after the resonance half cycle, \( i_C \) will be reverse discharged so that \( Q_{r2} \) of \( D_2 \) can be discharged and completely reverse biased off. \( V_{CC} \) can be expressed as (33). At this time, both \( i_{d_1} \) and \( i_{N_1} \) are zero.

\[ i_{Lr}(t) = i_{Lr}(t) = i_{Lr}(t_0) + \frac{V_m}{L_m} (t - t_1) \]
(30)

\[ i_{Lm}(t_2) = i_{Lr}(t_2) = i_{Lr}(t_1) + \frac{V_m}{L_m} (t_2 - t_1) \]
(31)

\[ i_{Lr}(t_2) = i_C(t_2) = i_{r_1} \]
(32)

\[ v_{C_r}(t_2) = v_C(t_1) + \frac{Q_{r2}}{C_r} \]
(33)

\[ i_{d_1}(t) = i_{Lm}(t) + i_{Lr}(t) \]
\[ = 0 \]
(34)

\[ v_{d_1}(t) = 0 \]
\[ = 0 \]
(35)

\[ i_{d_2}(t) = 0 \]
\[ = 0 \]
(36)

\[ i_{N_1}(t) = 0 \]
\[ = 0 \]
(37)

\[ \Delta t_2 = t_2 - t_1 = t_{r2} \]
(38)

3) **State 3:** As shown in Fig. 5(c), \( Q_1 \) remains on, and \( V_{in} \) continues to magnetize \( L_m \) in the forward direction, causing \( i_{Lm} \) to continue to rise. At this time, \( i_{Lr}, i_{D_1}, \) and \( i_{N_1} \) are zero.

\[ i_{Lm}(t) = i_{Lm}(t) = i_{Lm}(t_1) + \frac{V_m}{L_m} (t - t_1) \]
(39)

\[ i_{Lm}(t_3) = i_{Lm}(t_3) = i_{Lm}(t_1) + \frac{V_m}{L_m} (t_3 - t_2) \]
(40)

\[ i_{Lr}(t) = i_{Lr}(t) \]
\[ = 0 \]
(41)

\[ V_C(t_3) = V_{C}(t_2) \]
\[ = 0 \]
(42)

\[ i_{d_1}(t) = i_{Lm}(t) + i_{Lr}(t) \]
\[ = 0 \]
(43)

\[ v_{d_1}(t) = 0 \]
\[ = 0 \]
(44)

\[ i_{N_1}(t) = 0 \]
\[ = 0 \]
(45)

\[ \Delta t_3 = t_3 - t_2 = T_m - t_2 \]
(46)

4) **State 4:** As shown in Fig. 5(d), \( Q_1 \) is turned off. At this time, \( L_m \) charges \( C_{oss} \), and \( V_{d1} \) rises rapidly at the same time. \( V_{d1} \) can be expressed as (36). The voltage rising slew rate of \( V_{d1} \) can be described in (39). \( L_{LK1} \) charges \( C_{oss} \) and \( C_r \) at the same time to increase \( V_C \). Since \( C_r \) is much larger than \( C_{oss} \), \( V_{d1} \) in this interval rises slowly, so \( i_{d_1} \) and \( i_C \) are positive. In (60), \( V_{clamp} \) is the start point of clamping voltage by the proposed snubber on \( V_{d1} \) after \( Q_1 \) is turned off.

\[ i_{Lm}(t_4) = i_{Lm}(t_4) \]
\[ = i_{Lm}(t_3) \]
(48)

\[ i_{Lr}(t) = 0 \]
\[ = 0 \]
(49)

\[ V_C(t_4) = i_{Lm}(t_3) \left( \frac{C_r}{C_r + C_{oss}} \right) \]
\[ = 0 \]
(50)

\[ V_{d_1}(t_4) = i_{Lm}(t_3) \left( \frac{C_{oss}}{C_r + C_{oss}} \right) \]
\[ = 0 \]
(51)

\[ v_{C}(t_4) = i_{Lm}(t_3) \left( \frac{C_{oss}}{C_r + C_{oss}} \right) \]
\[ = 0 \]
(52)

\[ v_{d_1}(t_4) = V_o \cdot n_{ps} \]
\[ = 0 \]
(53)

\[ i_{d_1}(t_4) = i_{Lm}(t_3) \left( \frac{C_{oss}}{C_r + C_{oss}} \right) \]
\[ = 0 \]
(54)

\[ v_{d_1}(t_4) = \frac{i_{d_1}(t_4) \cdot (t - t_1)}{C_{oss}} \]
\[ = 0 \]
(55)

\[ v_{d_1}(t) = \frac{i_{d_1}(t) \cdot (t - t_1)}{C_{oss}} \]
\[ = 0 \]
(56)

\[ v_{d_1}(t) = V_o \cdot n_{ps} + V_m \]
\[ = 0 \]
(57)

\[ i_{N_1}(t) = 0 \]
\[ = 0 \]
(58)

\[ \frac{dV_{clamp}(prop)}{dt} = \frac{V_m + v_C(t_3) - V_{F_D1}}{C_{oss} + C_r + n_{ps} - v_{clamp}(prop)} \]
\[ = i_{Lm}(t) \]
(59)

\[ V_{clamp}(prop) = \frac{V_m + v_C(t_3) - V_{F_D1}}{C_{oss} + C_r + n_{ps} - v_{clamp}(prop)} \]
\[ = i_{Lm}(t) \]
(60)

\[ \Delta t_4 = t_4 - t_3 = \frac{C_{oss} + C_r}{V_m + v_C(t_3) - V_{F_D1} - v_{clamp}(prop)} \]
\[ = i_{Lm}(t) \]
(61)

5) **State 5:** As shown in Fig. 5(e), \( Q_1 \) remains off. But \( Q_2 \) is turned on. At this time, the stored energy of \( L_m \) starts to be released via the winding \( N_r \) to the output terminal. As \( V_C \) is
equal to $V_o/n_{ps}$. $L_{K1}$ starts to demagnetize and gradually decreases. $L_{K1}$ still charges $C_r$. The highest point of $v_C$ can be expressed as (68). When $L_{K1}$ is completely demagnetized, it enters state 6.

$$i_{Lm}(t) = i_{Lm}(t_4) + \left(-\frac{V_o}{n_{ps}}\right)(t-t_4)$$

(62)

$$i_{Lm}(t_4) = i_{Lm}(t_4) + \frac{-V_o \cdot n_{ps} (t_4 - t_4)}{L_m}$$

(63)

$$i_L(t_4) = 0$$

(64)

$$i_C(t) = i_{D1}(t) = i_{LK1}(t) \cdot \frac{C_r}{C_r + C_{os1}}$$

(65)

$$i_C(t) = i_{D1}(t_4) = 0$$

(66)

$$v_C(t) = v_C(t_4) + \left(Z_r \cdot i_C(t) \right) \cdot \sin \alpha (t-t_4)$$

(67)

$$v_C(t_4) = \frac{V_o \cdot n_{ps} - V_{F-D1} + i_{LK1} \cdot \sqrt{L_{K1}}}{C_{os1} + C_r} = v_{C_{max}}$$

(68)

$$v_{di1}(t) = v_C(t) + V_m + V_{F-D1}$$

(69)

$$i_{N}(t) = i_{Lm}(t) \cdot n_{ms}$$

(70)

$$\Delta t_5 = t_5 - t_4 = \frac{T_i}{2}$$

(71)

6) **State 6:** As shown in Fig. 5(f), $Q_1$ remains off. But $Q_2$ is turned on. When $L_{K1}$ is completely demagnetized, based on $D_1$ is not an ideal diode, there are $Q_{r1}$ and $i_{r1}$, so $C_r$ will generate a reverse current $i_{r1}$ to discharge $v_C$ and cut off, which will cause a slight discharge and drop. As $Q_{r1}$ is completely discharged and cut off, it enters state 7.

$$i_{Lm}(t) = i_{Lm}(t_5) + \left(-\frac{V_o}{n_{ps}}\right)(t-t_5)$$

(72)

$$i_{Lm}(t_5) = i_{Lm}(t_5) + \frac{-V_o \cdot n_{ps} (t_5 - t_5)}{L_m}$$

(73)

$$i_L(t_5) = 0$$

(74)

$$i_C(t) = i_{D1}(t) = i_{r1}$$

(75)

$$i_C(t_5) = i_{D1}(t_5) = 0$$

(76)

$$v_C(t) = v_C(t_5) + \left(Z_r \cdot i_C(t) \right) \cdot \sin \alpha (t-t_4)$$

(77)

$$v_C(t_5) = \frac{V_o \cdot n_{ps} + V_{F-D1}}{C_r}$$

(78)

$$i_{di1}(t) = 0$$

(79)

$$v_{di1}(t) = v_C(t) + V_m + V_{F-D1}$$

(80)

$$i_{n}(t) = i_{Lm}(t) \cdot n_{ms}$$

(81)

$$\Delta t_6 = t_6 - t_5 = T_i$$

(82)

7) **State 7:** As shown in Fig. 5(g), $Q_1$ remains off. But $Q_2$ turns on. As described in (83) and (84), $L_m$ continues to be magnetized via the winding $N_i$ to the output terminal. $L_{K1}$, $D_1$, $Q_2$, and $C_r$ all reach a steady state as (85) to (88). As $Q_1$ turns on, it returns to state 1.

III. **CIRCUIT CONFIGURATION OF LOSSLESS LC SNUBBER**

Fig. 6 shows the waveform of each point of the lossless LC snubber. It can be seen that in the state 1 ($t_o$–$t_4$) interval, the
voltage $v_{cr}$ will oscillate from positive polarity to negative polarity. Therefore, in state 4 ($t_{v~4}$), the voltage $v_{ds1}$ can rise from a low voltage and gradually increase upward.

As shown in Fig. 7(c), 3) State 3:

4) State 4: As shown in Fig. 7(d), $Q_1$ turned off. At this time, $L_{rK1}$ charges $C_{oss1}$, $L_{rK1}$ is still magnetized and $i_{Lr1}$ continues to rise, and $v_{ds1}$ rises rapidly at the same time. $v_{ds1}$ can be expressed as (97). The voltage rising slew rate of $v_{ds1}$ can be described in (101). $L_{rK1}$ charges $C_{oss1}$ and $C_r$ at the same time to increase $v_{Cr}$. Since $C_r$ is much larger than $C_{oss1}$, $v_{ds1}$, in this interval rises slowly, so $i_{ds1}$ and $i_{Cr}$ are positive. In (102), $v_{clamp}$(LC) is the start point of clamping voltage by the lossless LC snubber on $v_{ds1}$ after $Q_1$ is turned off. Compared to (24), (33), and (60), it shows the $v_{clamp}$ of the conventional lossless LC snubber is much higher than in the proposed method.

$$i_{ds1}(t) = i_{Lm}(t) \cdot \left( \frac{C_{oss1}}{C_r + C_{oss1}} \right)$$  \hspace{1cm} (96)

$$v_{ds1}(t) = i_{ds1}(t) \cdot \left( t - t_4 \right)$$  \hspace{1cm} (97)

$$i_{Cr}(t) = i_{D1}(t) = i_{Lm}(t) \cdot \left( \frac{C_r}{C_r + C_{oss1}} \right)$$  \hspace{1cm} (98)

$$v_{Cr}(t) = i_{Cr}(t) \cdot \left( t - t_3 \right)$$  \hspace{1cm} (99)

$$v_{Cr}(t_4) = V_o \cdot n_{ps}$$  \hspace{1cm} (100)

$$\frac{dv_{ds1}}{dt} = \frac{i_{ds1}(t)}{C_{oss1}}$$  \hspace{1cm} (101)

$$v_{clamp}(LC) = V_m + v_{Cr}(t_5) - V_{F-D1}$$  \hspace{1cm} (102)

$$\Delta t_4 = t_4 - t_3 = \left( \frac{C_{oss1} + C_r}{V_o \cdot n_{ps} - v_{ clamp(LC)}} \right)$$  \hspace{1cm} (103)

5) State 5: As shown in Fig. 7(e), $Q_1$ remains off, but $Q_2$ is turned on. At this time, the stored energy of $L_m$ starts to be released via the $N_s$ winding to the output terminal. As

$$v_{Cr} = \frac{V_o}{n_{ps}}$$, $L_{rK1}$ starts to demagnetize and gradually decreases. $L_{rK1}$ still charges $C_r$. The highest point of $v_{Cr}$ can be expressed as (105). When $L_{rK1}$ is completely demagnetized, it enters state 6.

$$i_{Cr}(t) = i_{D1}(t) = i_{Lk1}(t) \cdot \left( \frac{C_r}{C_r + C_{oss1}} \right)$$  \hspace{1cm} (104)

$$v_{cr}(t_5) = V_o \cdot n_{ps} - V_{F-D1} + L_{rK1} \cdot \left( \frac{L_{rK1}}{C_{oss1} + C_r} \right) = v_{Cr(max)}$$  \hspace{1cm} (105)

$$v_{ds1}(t) = v_{cr}(t) + V_m + V_{F-D1}$$  \hspace{1cm} (106)

$$\Delta t_5 = t_5 - t_4 = \frac{T_s}{2}$$  \hspace{1cm} (107)

6) State 6: As shown in Fig. 7(f), $Q_1$ remains off, but $Q_2$ is turned on. Because $D_1$ is not an ideal diode, it has $Q_{oss1}$ and $t_{oss1}$. When $L_{rK1}$ is completely demagnetized, there is a reverse current $i_{D1}$ to discharge $C_r$ and cut off after $t_{oss1}$, which causes a slight discharge and drop on $V_{O2}$. As $Q_{oss1}$ is completely discharged and cut off, it enters state 7.
\[ v_{C_2}(t_0) = v_{C_2}(t_0) - \frac{Q_{\text{off}}}{C_r} \]  
\[ v_{\text{di}}(t) = v_{C_2}(t) + V_{\text{in}} + V_{F,-D1} \]  

7) **State 7:** As shown in Fig. 7(g), \( Q_1 \) remains off, but \( Q_2 \) is turned on. \( L_m \) continues to be demagnetized with the winding \( N_t \) to the output terminal. The currents in \( L_{K1}, D_1, \) \( Q_1, C_r, D_2 \) and \( C_{\text{loss}} \) are zero. As \( Q_1 \) turns on, it returns to state 1.

\[ v_{C_2}(t_0) = v_{C_2}(t_0) \]  
\[ v_{\text{di}}(t) = v_{C_2}(t) + V_{\text{in}} + V_{F,-D1} \]

---

**FIGURE 7. Operation states of flyback with lossless LC snubber.**

**FIGURE 8. Waveform relevant to the flyback with energy regenerative snubber.**

The related parameters of the energy regenerative snubber circuit are defined as (4) to (15), and the initial conditions are as (115) and (116). Because there is no \( L_r \) in this topology, formulas (1), (2), and (3) are modified as (112), (113), and (114), respectively.

\[ a_0 = \frac{1}{\sqrt{C_r \cdot L_{K2}}} \]  
\[ T_0 = 2\pi \sqrt{C_r \cdot L_{K2}} \]  
\[ Z_0 = \sqrt{L_{K2} / C_r} \]  
\[ v_{C_2}(t_0) = V_o \cdot n_{ps} - V_{F,-D1} + i_{LK1} L_{K1} / \left( C_{\text{loss}} + C_r \right) \]  
\[ i_{L_r}(t_0) = i_{C_2}(t_0) = 0 \]  

1) **State 1:** As shown in Fig. 9(a), \( Q_1 \) is turned on. Because \( i_{L_{m1}} \) has an initial value, \( i_{d1} \) rises rapidly, and \( V_{\text{in}} \) continues to magnetize \( L_{m1} \) in the forward direction, causing \( L_{m1} \) to continue to rise. On the other hand, \( C_r \) resonates with \( L_{K2} \). At this time, \( D_2 \) is forward-biased conduction, and will reach the lowest point shown in (120) after half a resonant period. During this state, \( i_{D3} \) and \( i_{D4} \) are zero. In (121), \( V_{\text{clamp}(ER)} \) is the start point of clamping voltage by the lossless LC snubber on \( V_{\text{di}} \) after \( Q_1 \) is turned off.

\[ i_{L_2}(t) = i_{L_r}(t) = i_{d1}(t) \cdot \cos a_0(t - t_0) \]  
\[ \frac{V_{\text{in}}}{n_{ps}} - V_{C_2}(t_0) + \frac{Z_0}{i_{L_2}} \cdot \sin a_0(t - t_0) \]  
\[ i_{L_2}(t_1) = 0 \]  
\[ v_{C_2}(t) = \left( \frac{V_{\text{in}}}{n_{ps}} + V_{F,-D2} \right) + v_{C_2}(t_0) \cdot \cos a_0(t - t_0) \]  
\[ + (Z_0 \cdot i_{L_2}(t_1)) \cdot \sin a_0(t - t_0) \]
v_{CC}(t_i) = \left( \frac{V_{in} + V_{F-D1}}{n_p} \right) - v_{CC}(t_0) = v_{CC}(\text{min}) \tag{120}

v_{clamp(ER)} = V_{in} + v_{CC}(t_i) - V_{F-D1} \tag{121}

2) State 2: As shown in Fig. 9(b), \( Q_1 \) remains on, and \( V_{in} \) continues to magnetize \( L_m \) in the forward direction, causing \( L_m \) to continue to rise. During this state, \( i_{iss} \), \( i_{d1} \), and \( i_{d2} \) are zero.

\( i_{d2}(t_1) = i_{d2}(t_0) = 0 \tag{122} \)

\( i_{d2}(t_2) = i_{d2}(t_0) = 0 \tag{123} \)

\( v_{CC}(t_2) = v_{CC}(t_i) \tag{124} \)

3) State 3: As shown in Fig. 9(c), \( Q_1 \) turned off. At this time, \( L_{LK1} \) charges \( C_{oss1} \) and \( C_r \) to increase \( v_{CC} \). The \( V_{CC} \) can be expressed as (128). Since \( v_{CC} \) is always positive, \( v_{d11} \) in this interval rises quickly after \( Q_1 \) is turned off to induce high \( v_{clamp} \). Compared to (24), (33), and (60), (121) shows the \( v_{clamp} \) of energy regenerative snubber is also much higher than in the proposed method. At this interval, the stored energy of \( L_m \) starts to be discharged to the winding \( N_r \) to the output terminal. The highest point of \( v_{CC} \) can be expressed as (129). As \( L_{LK1} \) is completely demagnetized, it enters state 4.

\( i_{d1}(t_0) = 0 \tag{125} \)

\( i_{d2}(t_1) = i_{d1}(t) = i_{d2}(t) \left( \frac{C_r}{C_r + C_{oss1}} \right) \tag{126} \)

\( v_{CC}(t_1) = V_{CC}(t_1) + (Z_1 \cdot i_{d2}(t_1) \cdot \sin \omega t - t_2) \tag{127} \)

\( v_{CC}(t_2) = v_{CC}(t_1) + \sqrt{L_{LK1} \cdot \frac{L_{LK1}}{C_{oss1} + C_r}} \tag{128} \)

\( i_{d1}(t_2) = i_{d2}(t_2) \left( \frac{C_{oss1}}{C_r + C_{oss1}} \right) \tag{129} \)

\( \frac{dv_{d11}}{dt} = \frac{v_{d11}(t)}{C_{oss1}} \tag{130} \)

\( v_{d11}(t_3) = v_{d1}(t_3) + V_{F-D1} + V_{F-D1} \tag{131} \)

\( \Delta t_1 = t_3 - t_2 = \frac{(C_{oss1} + C_r)(V_{f} + n_p - V_{clamp(ER)})}{2} \tag{132} \)

\( v_{d11}(t_3) = V_{d1}(t_3) \tag{133} \)

4) State 4: As shown in Fig. 9(d), \( Q_1 \) remains off status, but \( Q_2 \) is turned on. Because \( D_1 \) is not an ideal diode, it has \( Q_{oss} \) and \( i_{oss} \). When \( L_{LK1} \) is completely demagnetized, there is a reverse current \( i_{d2} \) to discharge \( C_r \) and cut off after \( i_{oss} \), which can causes a slight discharge and drop on \( v_{CC} \). As \( Q_{oss} \) is completely discharged and cut off, it enters state 5.

\( i_{d2}(t_0) = 0 \tag{134} \)

\( i_{d2}(t_1) = i_{d2}(t) = i_{oss} \tag{135} \)

\( i_{d2}(t_2) = i_{d2}(t_0) = 0 \tag{136} \)

\( v_{CC}(t_2) = v_{CC}(t_2) - \frac{Q_{oss}}{C_r} \tag{137} \)

5) State 5: As shown in Fig. 9(e), \( Q_1 \) remains off, but \( Q_2 \) is turned on. \( L_m \) continues to demagnetize via the winding \( N_r \) to the output terminal. The currents on \( L_{LK1} \), \( D_1 \), \( Q_2 \), and \( C_{oss1} \) are zero. As \( Q_2 \) is turned on, it returns to state 1.

\( i_{d2}(t_0) = i_{d2}(t) = i_{oss}(t) = i_{LK1}(t) = 0 \tag{138} \)

\( v_{CC}(t_0) = v_{CC}(t_i) \tag{139} \)

\( v_{d11}(t) = v_{CC}(t) + V_{in} + V_{F-D1} \tag{140} \)

![FIGURE 9. Operation states of the flyback converter with energy regenerative snubber.](image-url)

Equations (60), (102), (121), (61), (103), and (133) can be organized as Table II. It is known that \( v_{clamp(ER)} > v_{clamp(LC)} > v_{clamp(proposal)} \). It is also known that \( \Delta t_1 \) rising time and \( v_{clamp} \) are positively correlated, so the \( \Delta t_1 \) rising time of the proposed snubber is related to \( v_{clamp} \). The slower rising slope of \( \Delta t_1 \) is helpful to reach better EMI performance.

| TABLE II COMPARISON OF \( v_{clamp} \) AND \( v_{d11} \) RISING TIME |
|---------------------------------|------------------|------------------|
| Topology                        | \( v_{clamp} \)   | \( v_{d11} \) rising time |
| Proposed circuit                | \( v_{clamp(proposal)} = V_{in} + v_{CC}(t_i) - V_{F-D} \) | \( \frac{(C_{oss1} + C_r)(V_{f} + n_p - V_{clamp(ER)})}{2} \) |
| Lossless LC snubber             | \( v_{clamp(LC)} = V_{in} + v_{CC}(t_i) - V_{F-D} \) | \( \frac{(C_{oss1} + C_r)(V_{f} + n_p - V_{clamp(ER)})}{2} \) |
| Energy regenerative snubber     | \( v_{clamp(ER)} = V_{in} + v_{CC}(t_i) - V_{F-D} \) | \( \frac{(C_{oss1} + C_r)(V_{f} + n_p - V_{clamp(ER)})}{2} \) |
V. CIRCUIT SPECIFICATIONS AND PARAMETERS

These three flyback prototype circuits are modified from an existing business 72 W flyback adapter product. Therefore, the design step of the main transformer is skipped to simplify the design process. The prototype circuits are built with the following specifications and parameters shown in Tables III and IV. The variables in the design process of the prototype in this paper are listed in Table V.

### Table III
**SYSTEM SPECIFICATION OF THE PROTOTYPES**

| Item                     | Specification |
|--------------------------|---------------|
| Rated high line input voltage $V_{in(h)}$ | 230 Vac/50 Hz |
| Rated low line input voltage $V_{in(l)}$ | 115 Vac/60 Hz |
| Output voltage $V_o$    | 12 V          |
| Rated output power $P_o$ | 72 W          |
| Rated output current $I_{out}$ | 6 A          |
| Expected efficiency $\eta$ | 90%          |
| Switching frequency $f_{sw}$ | 100 kHz     |

### Table IV
**SPECIFICATION OF KEY COMPONENTS**

| Item     | Detailed specifications, part number, and manufacturer |
|----------|-------------------------------------------------------|
| $U_r$, $Q_r$ | PowGoaN Switch IN363/70C, 750 V, $R_{on}=a=350$ mΩ IC operating current $I_{on}=2$ mA |
| $Q_2$     | AON56920 100 V, AOS $R_{on}=10.7$ mΩ, $C_{on}=485$ pF |
| $T_1$     | ECW25, JPF95, A-core, $L_{a1}=500$ μH, $L_{a2}=2.12$ μH, $N_a=36$ |
| $D_1$, $D_2$ | US1M, 1A, 1000 V, Panjit $V_{f}=V_{f}=0.7$ V |
| $C_1$     | 1 mF, 1206 MLCC, Murata 47 μH, RM8, JPF95, A-core |
| $L_r$     | 47 mH, RM8, JPF95, A-core, $R_{on}=0.2$ Ω, Resistance of $L_r$ 47 mH |
| $C_o$     | 2 x 1000 μF/16V, PSC, Nippon Chemi-Con ESRc=10 mΩ, Equivalent series resistance |
| BD        | 2 x TMBF310, Chongqing Pingwei Tech, $V_{f}=11.1$ V |
| $Y$ cap   | 1 nF/250 Vac, Y1 class safety capacitor |

### Table V
**TABLE V**

| Item                     | Specification |
|--------------------------|---------------|
| $D_{on(h)}$              | Maximum duty cycle at low line $V_o$ and high line $V_{in}$ respectively |
| $N_{pp}$                 | Calculated minimum turn of winding $N_p$ |
| $t_{on}$                 | Turn-on delay time $100$ns of $Q_r$ |
| $R_{ch1}$               | DC resistance of common chokes at low line and high line $I_{in}$ |
| $I_{F1}$                 | Peak forward current of $D_1$ at low line $V_o$ and high line $V_{in}$ respectively |
| $I_{F2}$                 | Peak forward current of $D_2$ at low line $V_o$ and high line $V_{in}$ respectively |
| $I_{F3}$                 | Average forward current of $D_2$ at low line $V_o$ and high line $V_{in}$ respectively |
| $I_{F4}$                 | RMS forward current of $D_2$ at low line $V_o$ and high line $V_{in}$ respectively |
| $V_m$                    | Reversed peak voltage of $D_1$ at low line $V_o$ and high line $V_{in}$ respectively |
| $V_{m1}$                 | Reversed peak voltage of $D_2$ at low line $V_o$ and high line $V_{in}$ respectively |
| $V_{m2}$                 | The start point of clamping voltage of proposed snubber at low line $V_o$ and high line $V_{in}$ respectively |
| $\Delta V_{m1}$         | Ripple current of $N_1$ at low line $V_o$ and high line $V_{in}$ respectively |
| $\Delta V_{m2}$         | Ripple current of $N_1$ at low line $V_o$ and high line $V_{in}$ respectively |
| $P_{R1}$                 | RMS current of $N_1$ at low line $V_o$ and high line $V_{in}$ respectively |
| $P_{R2}$                 | RMS current of $N_1$ at low line $V_o$ and high line $V_{in}$ respectively |
| $B_{max}$                | Maxima flux density of $T_1$ at low line $V_o$ and high line $V_{in}$ respectively |
| $V_{rms}$                | RMS ripple current of $C_1$ at low line $V_o$ and high line $V_{in}$ respectively |
| $P_{R3}$                 | The output power when boundary of CCM/DCM at low line $V_o$ and high line $V_{in}$ respectively |
| $P_{on}$                 | Switching loss of $Q_1$ at low line $V_o$ and high line $V_{in}$ respectively |
| $P_{cond}$               | Conduction loss of $Q_1$ at low line $V_o$ and high line $V_{in}$ respectively |
| $P_{cond+I_{on}}$        | Total loss of $Q_1$ at low line $V_o$ and high line $V_{in}$ respectively |
| $P_{cond+I_{off}}$       | Conduction loss of $Q_2$ at low line $V_o$ and high line $V_{in}$ respectively |
| $P_{cond+I_{on1}}$       | Recovery charge loss of $Q_2$ at low line $V_o$ and high line $V_{in}$ respectively |
| $P_{cond+I_{off1}}$      | Conduction loss of $Q_2$ at low line $V_o$ and high line $V_{in}$ respectively |
| $P_{cond}$               | Copper loss of $N_1$ at low line $V_o$ and high line $V_{in}$ respectively |
| $P_{cond}$               | Copper loss of $N_1$ at low line $V_o$ and high line $V_{in}$ respectively |
| $P_{cond}$               | Copper loss of $N_1$ at low line $V_o$ and high line $V_{in}$ respectively |
| $P_{cond}$               | Copper loss of $N_1$ at low line $V_o$ and high line $V_{in}$ respectively |
| $P_{loss}$               | Loss of common chokes at low line $V_o$ and high line $V_{in}$ respectively |
| $P_{loss}$               | Power loss of bridge loss at low line $V_o$ and high line $V_{in}$ respectively |
| $P_{loss}$               | Loss of $C_1$ at low line $V_o$ and high line $V_{in}$ respectively |
| $P_{loss}$               | Total loss at low line $V_o$ and high line $V_{in}$ respectively |
| $\eta_{est}$            | Estimative efficiency at low line $V_o$ and high line $V_{in}$ respectively |
The system blocks for the conventional lossless LC snubber, the energy regenerative snubber, and the proposed snubber are shown in Fig. 10.

![Diagram of system blocks]

**FIGURE 10.** System block of the flyback converters for verification with: (a) Lossless LC snubber; (b) Energy regenerative snubber; (c) The proposed snubber.

**VI. DESIGN OF PROTOTYPE**

The first design step is to calculate the duty cycle in CCM with high line and low line input voltages.

\[
D_{\text{max}}(LL) = \frac{N_p \left( V_o + R_{dl(2m)} \cdot i_{ns} \right)}{v_{\text{in}(LL)} + N_p \left( V_o + R_{dl(2m)} \cdot i_{ns} \right)} = 0.35 \quad (141)
\]

\[
D_{\text{max}}(HL) = \frac{N_p \left( V_o + R_{dl(2m)} \cdot i_{ns} \right)}{v_{\text{in}(HL)} + N_p \left( V_o + R_{dl(2m)} \cdot i_{ns} \right)} = 0.211 \quad (142)
\]

Then, the DCM/CCM boundary can be obtained. When the load current is above 43% of rated load at the low line input voltage and 63.3% of rated load at the high line input voltage, the converter operates in CCM.

\[
P_{\text{pD}}(LL) = \frac{v_{\text{in}(LL)}}{2 \cdot L_m} D_{\text{max}}^2(LL) = 31 \, \text{W} \quad (143)
\]

\[
P_{\text{pD}}(HL) = \frac{v_{\text{in}(HL)}}{2 \cdot L_m} D_{\text{max}}^2(HL) = 45.6 \, \text{W} \quad (144)
\]

When over 63% of rated load, the converter enters the CCM under high line input voltage. For the reasons above, it can be seen that it operates in CCM when the circuit is fully loaded even at the high line input voltage or low line input voltage.

**A. Current calculation for \(N_p\) and \(Q_1\)**

The formulas of peak current of \(Q_1\) in CCM are also shown.

\[
i_{N_p-pk(LL)} = \frac{v_{\text{in}(LL)}}{2 \cdot L_m} D_{\text{max}}(LL) + \left( \frac{P_{\text{in}}}{v_{\text{in}(LL)} \cdot D_{\text{max}}(LL)} \right) = 1.96 \, \text{A} \quad (145)
\]

\[
i_{N_p-pk(HL)} = \frac{v_{\text{in}(HL)}}{2 \cdot L_m} D_{\text{max}}(HL) + \left( \frac{P_{\text{in}}}{v_{\text{in}(HL)} \cdot D_{\text{max}}(HL)} \right) = 1.829 \, \text{A} \quad (146)
\]

It can be ensured that the peak current in \(Q_1\) happens in CCM. Then both \(i_{N_p-pk(LL)}\) and \(i_{N_p-pk(HL)}\) can be obtained as (147) and (148).

\[
i_{N_p-pk(LL)} = 1.96 \, \text{A} \quad (147)
\]

\[
i_{N_p-pk(HL)} = 1.829 \, \text{A} \quad (148)
\]

Also, the peak current of the secondary winding and rectifier can be calculated as the following equation.

\[
i_{N_{p-min}} = \frac{L_m \cdot i_{N_p-pk}(LL)}{B_{5.T1} \cdot A_{e-T1} \cdot 85\%} = 31.678 \quad (149)
\]

In industrial design, the de-rating of flux density saturation is acceptable 90%. The turn number of winding \(N_{p-min}\) is 31.678 as (150). However, it is not a reasonable solution.

\[
N_{p-min} = \frac{L_m \cdot i_{N_p-pk}(LL)}{B_{5.T1} \cdot A_{e-T1} \cdot 85\%} = 31.678 \quad (150)
\]

Here, the turn number of \(N_p\) is determined with 36 and the maximum flux density is 0.276 T for the low line input voltage and 0.257 T for the high line input voltage.

\[
B_{\text{max}-T1(LL)} = \frac{L_m \cdot i_{N_p-pk}(LL)}{N_p \cdot A_{e-T1}} = 0.276 \, \text{T} \quad (151)
\]

\[
B_{\text{max}-T1(HL)} = \frac{L_m \cdot i_{N_p-pk}(HL)}{N_p \cdot A_{e-T1}} = 0.257 \, \text{T} \quad (152)
\]

Next, the change in the magnetizing currents of the primary winding \(i_{N_{p-min}(LL)}\) and \(i_{N_{p-min}(HL)}\) in CCM can be calculated as (153) and (154):

\[
\Delta i_{Lm(LL)} = \frac{v_{\text{in}(LL)} \cdot D_{\text{max}(LL)}}{L_m \cdot f_{sw}} \quad (153)
\]

\[
\Delta i_{Lm(HL)} = \frac{v_{\text{in}(HL)} \cdot D_{\text{max}(HL)}}{L_m \cdot f_{sw}} \quad (154)
\]

Take (145) into (153) to obtain (155) and take (146) into (154) to obtain (156):

\[
i_{N_{p-min}(LL)} = \sqrt{D_{\text{max}(LL)} \left( i_{N_p-pk(LL)} - \frac{\Delta i_{Lm(LL)}}{2} \right)^2 + \frac{\Delta i_{Lm(LL)}}{12}} \quad (155)
\]

\[= 0.856 \, \text{A} \]

This work is licensed under a Creative Commons Attribution 4.0 License. For more information, see https://creativecommons.org/licenses/by/4.0/
\[ i_{np-rms(HL)} = \sqrt{\frac{D_{\text{max}(HL)} \left( i_{np-peak(HL)} - \frac{\Delta i_{\text{Lm}(HL)}}{2} \right)^2 + \frac{\Delta i_{\text{Lm}(HL)}^2}{12}}{}} \]

\[ = 0.564 \text{ A} \]

**B. Current calculation for \( N_i \)**

Next, the peak-to-peak current \( \Delta i_{N_i(LL)} \) and \( \Delta i_{N_i(HL)} \) can be calculated as (157) and (158). Their RMS values are also obtained as (159) and (160). Finally, the average currents \( i_{\text{N-avg}}(LL) \) and \( i_{\text{N-avg}}(HL) \) are calculated as (161) and (162):

\[ \Delta i_{N_i(LL)} = \frac{N_{ps}^2 \cdot V_o}{L_m} \cdot \frac{1 - D_{\text{max}(LL)}}{f_{sw}} = 7.89 \text{ A} \]  

(157)

\[ \Delta i_{N_i(HL)} = \frac{N_{ps}^2 \cdot V_o}{L_m} \cdot \frac{1 - D_{\text{max}(HL)}}{f_{sw}} = 9.57 \text{ A} \]  

(158)

\[ i_{\text{N-avg}}(LL) = \left( 1 - D_{\text{max}(LL)} \right) N_{ps} \cdot i_{np-peak(LL)} - \Delta i_{N_i(LL)} \left( 1 - D_{\text{max}(LL)} \right) = 6.63 \text{ A} \]  

(161)

\[ i_{\text{N-avg}}(HL) = \left( 1 - D_{\text{max}(HL)} \right) N_{ps} \cdot i_{np-peak(HL)} - \Delta i_{N_i(HL)} \left( 1 - D_{\text{max}(HL)} \right) = 6.64 \text{ A} \]  

(162)

**C. RMS Current of \( I_{CO} \)**

The RMS current of \( C_r \) can be calculated as (163) and (164).

\[ i_{\text{rs-rms}(LL)} = \sqrt{i_{\text{rs-rms}(LL)}^2 - i_o^2} = 5.88 \text{ A} \]  

(163)

\[ i_{\text{rs-rms}(HL)} = \sqrt{i_{\text{rs-rms}(HL)}^2 - i_o^2} = 5.05 \text{ A} \]  

(164)

**D. Voltage calculation of \( C_r \) and \( V_{dc1} \)**

The maximum and minimum voltages on \( C_r \) are derived from the high line and low line input voltage as in the following equations.

\[ V_{Cr-max(LL)} = N_{ps} \left( V_o + R_{d2(20)} \cdot i_{N_i} \right) - V_{F-D1} \]  

(165)

\[ + \left( i_{np-peak(LL)} \cdot \frac{L_{k1}}{C_{o1} + C_r} \right) \cdot \frac{Q_{rel}}{C} = 137.28 \text{ V} \]

\[ V_{Cr-max(HL)} = N_{ps} \left( V_o + R_{d2(20)} \cdot i_{N_i} \right) - V_{F-D1} \]  

(166)

\[ + \left( i_{np-peak(HL)} \cdot \frac{L_{k1}}{C_{o1} + C_r} \right) \cdot \frac{Q_{rel}}{C} = 131.3 \text{ V} \]

\[ v_{Cr-min(LL)} = \frac{Q_{rel}}{C_r} - N_{ps} \left( V_o + R_{d2(20)} \cdot i_{N_i} \right) \]  

(167)

\[ + \left( i_{np-peak(LL)} \cdot \frac{L_{k1}}{C_{o1} + C_r} \right) \cdot - V_{F-D1} \]

\[ = -167.68 \text{ V} \]

\[ v_{Cr-min(HL)} = \frac{Q_{rel}}{C_r} - N_{ps} \left( V_o + R_{d2(20)} \cdot i_{N_i} \right) \]  

(168)

\[ + \left( i_{np-peak(HL)} \cdot \frac{L_{k1}}{C_{o1} + C_r} \right) \cdot - V_{F-D1} \]

\[ = -188.87 \text{ V} \]

The maximum peak value on \( v_{dc1} \) appears at the high line and when the circuit is operated at full load.

\[ v_{dc1-peak(LL)} = N_{ps} \left( V_o + R_{d2(20)} \cdot i_{N_i} \right) + v_{Cr-max(LL)} \]  

(169)

\[ = 543.4 \text{ V} \]

\[ v_{dc1-peak(HL)} = N_{ps} \left( V_o + R_{d2(20)} \cdot i_{N_i} \right) + v_{Cr-max(LL)} \]  

(170)

\[ = 380.4 \text{ V} \]

**E. The start point of clamping voltage \( V_{clamp} \)**

The clamping voltage \( V_{clamp(prop)} \) of the proposed circuit is calculated as (171) and (172). After \( Q_1 \) is turned off, \( V_{dc1} \) rises from \( V_{clamp(prop)} \). When \( V_{dc1} \) reaches \( V_{clamp(prop)} \), the slew rate control function will execute, so this voltage should be as low as possible. From (171), it can be seen \( V_{clamp(LL)} \) should be -12.8 V at the low line input voltage. However, the body diode of \( Q_1 \) can clamp the \( V_{dc1} \) at -0.9 V. The \( V_{clamp(LL)} \) is obtained as (172) at the low line input voltage.

\[ V_{clamp(LL)} = V_{min(LL)} - V_{Cr-min(LL)} = -12.8 \text{ V} \]  

(171)

\[ V_{clamp(HL)} = V_{min(HL)} - V_{Cr-min(HL)} = 123.8 \text{ V} \]  

(172)

**F. Voltage and current calculation of \( D_2 \)**

For a sinusoidal wave, the average value is 0.9 times of its RMS value. Therefore, the average value and RMS value of \( i_{D2} \) are obtained as (175) to (178):

\[ i_{D2-peak(LL)} = \sqrt{\frac{C_r \cdot V_{Cr-max(LL)}^2}{L_r}} = 0.651 \text{ A} \]  

(173)

\[ i_{D2-peak(HL)} = \sqrt{\frac{C_r \cdot V_{Cr-max(HL)}^2}{L_r}} = 0.531 \text{ A} \]  

(174)

\[ i_{D2-avg(LL)} = \frac{1}{\sqrt{2}} \sqrt{\frac{C_r \cdot V_{Cr-max(LL)}^2}{L_r}} \left( 1 + \frac{2 \cdot \pi \sqrt{L_r \cdot C_r}}{T_{sw}} \right) \]  

(175)

\[ = 20.3 \text{ mA} \]

\[ i_{D2-avg(HL)} = \frac{1}{\sqrt{2}} \sqrt{\frac{C_r \cdot V_{Cr-max(HL)}^2}{L_r}} \left( 1 + \frac{2 \cdot \pi \sqrt{L_r \cdot C_r}}{T_{sw}} \right) \]  

(176)

\[ = 19.7 \text{ mA} \]

And the reverse voltage \( V_{D2-peak(LL)} \) and \( V_{D2-peak(HL)} \) of \( D_2 \) are shown as (179) and (180), respectively.
The calculation formulas of the voltage and current of \( D_1 \) are shown as (181) to (185):

\[
v_{\text{D1-pk}(LL)} = v_{\text{int}(LL)} + N_s v_o \left( v_o + R_{ds(2\text{on})} \cdot i_{N(t)(LL)} \right) = 176.5 \text{ V}
\]

\[
v_{\text{D1-pk}(HL)} = v_{\text{int}(HL)} + N_s v_o \left( v_o + R_{ds(2\text{on})} \cdot i_{N(t)(HL)} \right) = 339.5 \text{ V}
\]

\[
i_{\text{D1-avg}(LL)} = \frac{2}{\pi} \sqrt{i_{\text{L1}}(LL)} \cdot C_r \cdot i_{Np-pk(LL)} = 20.05 \text{ mA}
\]

\[
i_{\text{D1-avg}(HL)} = \frac{2}{\pi} \sqrt{i_{\text{L1}}(HL)} \cdot C_r \cdot i_{Np-pk(HL)} = 18.71 \text{ mA}
\]

**H. Loss estimation of \( Q_1 \)**

Since the manufacturer of \( Q_1 \) GaN FET does not provide detailed parameter information such as gate charge, gate capacitance, and threshold voltage, it is unavailable to estimate the switching loss. This paper can only calculate the conduction loss and the switching loss of \( C_{\text{oss}} \) as (187) and (188), respectively. The conduction losses are shown as (189) and (190). Finally, the total loss of \( Q_1 \) is shown as (191) and (192):

\[
P_{\text{oss}(LL)} = \frac{C_{\text{oss}} \cdot v_d^{2}}{2} \cdot f_{sw} = 0.145 \text{ W}
\]

\[
P_{\text{oss}(HL)} = \frac{C_{\text{oss}} \cdot v_d^{2}}{2} \cdot f_{sw} = 0.295 \text{ W}
\]

**I. Loss estimation of \( T_1 \)**

Since the \( B_{\text{max}} \) values of \( T_1 \) provided by the supplier under high line and low line conditions are too close, the core loss is calculated to be the same result. \( B_{\text{max}} \) values of \( T_1 \) under 100 kHz can be found in the data sheet.

Since the switching frequency is 100 kHz and multi-stranded wire has been used, in order to simplify the design, the skin effect is not considered here, and the measured DC resistance of each winding is directly used for calculation.

The copper loss of each winding is shown as (194) to (199). The auxiliary winding is only used for powering PWM IC \( U_1 \), and its operation current is only 1.5 mA, so the winding loss is not included in the calculation:

\[
P_{\text{core-}T_1} = V_{c-T_1} \cdot P_{CV-T_1} = 3.56 \text{ W}
\]

\[
P_{\text{copp-Np}(LL)} = i_{Np-pk(LL)}^2 \cdot ACR_{Np} = 0.097 \text{ W}
\]

\[
P_{\text{copp-Np}(HL)} = i_{Np-pk(HL)}^2 \cdot ACR_{Np} = 0.042 \text{ W}
\]

\[
P_{\text{copp-Ns}(LL)} = i_{Ns-pk(LL)}^2 \cdot ACR_{Ns} = 0.561 \text{ W}
\]

\[
P_{\text{copp-Ns}(HL)} = i_{Ns-pk(HL)}^2 \cdot ACR_{Ns} = 0.489 \text{ W}
\]

\[
P_{\text{copp-Nr}(LL)} = i_{Np-pk(LL)}^2 \cdot ACR_{Nr} = 31 \mu\text{W}
\]

\[
P_{\text{copp-Nr}(HL)} = i_{Np-pk(HL)}^2 \cdot ACR_{Nr} = 28 \mu\text{W}
\]

**J. Loss estimation of \( L_1 \)**

Since \( i_{L1} \) is equal to \( i_{Ls} \), the copper loss can be calculated as (200) and (201). The core loss can also be calculated as (202) according to the core loss density volume \( P_{\text{CL}} = 3.2 \text{ kW/m}^3 \) of the JPP95 data sheet under the operation conditions of 100 kHz, 50 mT and 100 °C:

\[
P_{\text{copp-Lr}(LL)} = i_{Np-pk(LL)}^2 \cdot ACR_{Lr} = 69 \mu\text{W}
\]

\[
P_{\text{copp-Lr}(HL)} = i_{Np-pk(HL)}^2 \cdot ACR_{Lr} = 69 \mu\text{W}
\]

\[
P_{\text{core-Lr}(LL)} = P_{\text{core-Lr}(HL)} = V_{c-Lr} \cdot P_{CL-Lr} = 34 \text{ mW}
\]

**K. Loss estimation of \( D_1 \) and \( D_2 \)**

The formulas of power loss analysis of \( D_1 \) and \( D_2 \) are estimated as (203) to (206):

\[
P_{\text{D1}(LL)} = \frac{Q_{1r1} \cdot v_{\text{D1-pk}(LL)} \cdot f_{sw} + i_{\text{D1-avg}(LL)} \cdot V_{F-D1}}{2} = 0.693 \text{ W}
\]

\[
P_{\text{D1}(HL)} = \frac{Q_{1r2} \cdot v_{\text{D1-pk}(HL)} \cdot f_{sw} + i_{\text{D1-avg}(HL)} \cdot V_{F-D1}}{2} = 1.302 \text{ W}
\]

\[
P_{\text{D2}(LL)} = \frac{Q_{1r2} \cdot v_{\text{D2-pk}(LL)} \cdot f_{sw} + i_{\text{D2-avg}(LL)} \cdot V_{F-D2}}{2} = 1.281 \text{ W}
\]

\[
P_{\text{D2}(HL)} = \frac{Q_{1r2} \cdot v_{\text{D2-pk}(HL)} \cdot f_{sw} + i_{\text{D2-avg}(HL)} \cdot V_{F-D2}}{2} = 1.972 \text{ W}
\]

**L. Loss estimation of \( Q_2 \)**

The conduction loss of \( Q_2 \) is shown as (207) and (208). The loss of parasitic capacitor \( C_{\text{oss2}} \) of \( Q_2 \) is shown as (209) and (210). In (211) and (212), the switching loss from \( Q_{2r1} \) of \( Q_2 \) is shown as (207) and (208). The gate driving loss of \( Q_2 \) is shown as (213).

\[
P_{\text{cond-Q2}(LL)} = \frac{R_{\text{ds(2on)}} \cdot I_{\text{ds(2on)}}^2 + R_{\text{ds(2on)}} \cdot i_{N(t)(LL)}^2 \cdot i_{\text{Q2}} \cdot f_{sw}}{12} = 0.552 \text{ W}
\]

\[
P_{\text{cond-Q2}(HL)} = \frac{R_{\text{ds(2on)}} \cdot I_{\text{ds(2on)}}^2 + R_{\text{ds(2on)}} \cdot i_{N(t)(HL)}^2 \cdot i_{\text{Q2}} \cdot f_{sw}}{12} = 0.487 \text{ W}
\]

\[
P_{\text{Qoss2}(LL)} = \frac{C_{\text{oss2}}}{2} \left( \frac{v_o + v_{\text{int}(LL)}}{N_{pi}} \right)^2 \cdot f_{sw} = 29 \text{ mW}
\]

\[
P_{\text{Qoss2}(HL)} = \frac{C_{\text{oss2}}}{2} \left( \frac{v_o + v_{\text{int}(HL)}}{N_{pi}} \right)^2 \cdot f_{sw} = 79 \text{ mW}
\]
\[ P_{QrQ2(LL)} = Q_{rQ2} \left( V_a + \frac{V_{in(LL)}}{N_{ps}} \right) \cdot f_{sw} = 0.587 \text{ W} \]  
(211)

\[ P_{QrQ2(HL)} = Q_{rQ2} \left( V_a + \frac{V_{in(HL)}}{N_{ps}} \right) \cdot f_{sw} = 0.971 \text{ W} \]  
(212)

\[ P_{dvQ2} = Q_{dvQ2} \cdot V_{dvQ2} \cdot f_{sw} = 7.5 \text{ mW} \]  
(213)

The total power loss of \( Q_2 \) can be estimated as (214) and (215):

\[ P_{tot-Q2(LL)} = P_{cond-Q2(LL)} + P_{dv-Q2(LL)} + P_{loss-2(LL)} \]  
(214)

\[ P_{tot-Q2(HL)} = P_{cond-Q2(HL)} + P_{dv-Q2(HL)} + P_{loss-2(HL)} \]  
(215)

\[ P_{tot-Q2} = P_{QrQ2(LL)} + P_{QrQ2(HL)} \]  
(216)

\[ P_{PQrQ2} = \frac{P_o}{N_{ps}} \]  
(217)

\[ P_{tot loss} = P_{tot loss HL} + P_{tot loss LL} \]  
(218)

\[ P_{tot loss HL} = \frac{P_{PQrQ2}}{N_{ps}} \]  
(219)

\[ P_{tot loss LL} = \frac{P_{PQrQ2}}{N_{ps}} \]  
(220)

\[ \eta_{(HL)} = \frac{P_o}{P_o + P_{tot loss}} = 87.6\% \]  
(225)

The experimental results of the energy regenerative snubbers are shown in Figs. 14 and 15. It can be observed that there is no significant change in the rising slope of \( v_{ds} \). Compared with the \( v_{ds} \) rising slope of the energy regenerative snubber in Figs. 15 and 16, the \( v_{ds} \) rising slope of the proposed circuit in Figs. 17 and 18 are significantly slower, which proves that the proposed circuit can indeed achieve a significant switching speed reduction.

After \( Q_1 \) is turned off and \( V_{ds} \) rises to the highest point, it can be seen that \( i_C \) will have a negative current, which is due to the current of the diode recovery charge \( Q_o \) to discharge \( C_r \) reverse. This behavior occurs in all these three topologies.

The experimental results of conversion efficiency are shown in Fig. 19. Under 80% to 100% of rated load, the efficiency of the proposed topology is about 0.1% higher than the efficiency of the energy regenerative snubber.

\[ \eta_{(LL)} = \frac{P_o}{P_o + P_{tot loss}} \]  
(224)
FIGURE 13. Waveforms on $i_{NP}$, $v_{ds}$, $i_{Cr}$, and $v_{Cr}$ with the lossless LC snubber under 25% of rated load and high line input voltage:
(a) Full time scale at 115 Vac; (b) Rising edge of $v_{ds}$ at 115 Vac; (c) Full time scale at 230 Vac; (d) Rising edge of $v_{ds}$ at 230 Vac.

FIGURE 14. Waveforms on $i_{NP}$, $v_{ds}$, $i_{Cr}$, and $v_{Cr}$ with the lossless LC snubber under rated load and high line input voltage:
(a) Full time scale at 115 Vac; (b) Rising edge of $v_{ds}$ at 115 Vac; (c) Full time scale at 230 Vac; (d) Rising edge of $v_{ds}$ at 230 Vac.

FIGURE 15. Waveforms on $i_{NP}$, $v_{ds}$, $i_{Cr}$, and $v_{Cr}$ with the energy regenerative snubber under 25% of rated load and high line input voltage:
(a) Full time scale at 115 Vac; (b) Rising edge of $v_{ds}$ at 115 Vac; (c) Full time scale at 230 Vac; (d) Rising edge of $v_{ds}$ at 230 Vac.

FIGURE 16. Waveforms on $i_{NP}$, $v_{ds}$, $i_{Cr}$, and $v_{Cr}$ with the energy regenerative snubber under rated load and high line input voltage:
(a) Full time scale at 115 Vac; (b) Rising edge of $v_{ds}$ at 115 Vac; (c) Full time scale at 230 Vac; (d) Rising edge of $v_{ds}$ at 230 Vac.

FIGURE 17. Waveforms on $i_{NP}$, $v_{ds}$, $i_{Cr}$, and $v_{Cr}$ with the proposed snubber under 25% of rated load and high line input voltage:
(a) Full time scale at 115 Vac; (b) Rising edge of $v_{ds}$ at 115 Vac; (c) Full time scale at 230 Vac; (d) Rising edge of $v_{ds}$ at 230 Vac.
Table VI lists $v_{ds1}$ turned-off rising slew rate of these three prototypes. The proposed circuit slows down the turned-off rising slew rate compared to the energy regenerative snubber, but similar to the lossless LC snubber.

The experimental EMI results can also verify the previous waveforms in Fig. 20 to Fig. 22. Even the energy regenerative snubber provides higher efficiency, but its EMI performance is worse than the proposed circuit and the lossless LC snubber. The proposed method significantly reduces conduction EMI emissions in the frequency range of 500 kHz to 2 MHz and radiation EMI emissions up to 6 dBμV in the frequency range of 60 MHz to 230 MHz.

**VIII. CONCLUSION**

In view of the high frequency and wide bandgap switching devices brought by the miniaturization of power converters, faster switching speeds and EMI problems that come with them, this paper proposes a slope control snubber. It can be applied to the flyback converter. The experimental results show that compared to the previous methods [13]-[15], the proposed snubber significantly improves both conduction and radiation EMI emissions with the same components.
REFERENCES

[1] H. Chen, W. Dong, Y. He and Z. Qian, “Secondary side post regulation application in multiple outputs flyback converter,” in Proc. IEEE PEDS Conf., Kuala Lumpur, Malaysia, 2005, pp. 1273-1277.

[2] C.C. Wen and C.L. Chen, “Magamp application and limitation for multiwinding flyback converter,” IEEE Electric Power Applications, vol. 152, pp. 517-525, May 2005, 10.1049/ip-epa:20040829.

[3] D. Han, S. Li, W. Lee, W. Choi and B. Sarlioglu, “Trade-off between switching loss and common mode EMI generation of GaN devices—analysis and solution,” in Proc. IEEE APEC Conf., Tampa, FL, USA , 2017, pp. 843-847.

[4] Alenka Hren, Joze Korelic and Miro Milanovic, “RC-RCD clamp circuit for ringing losses reduction in a flyback converter,” IEEE Trans. Circuits and systems—II:Express Briefs, vol. 53, no. 5, pp. 369-373, May 2006, 10.1109/TCSII.2006.870547.

[5] Alexander Abramovitz, Chih-Sheng Liao, and Keyue M. Smedley, “State-plane analysis of regenerative snubber for flyback converters,” IEEE Trans. Power Electronics, vol. 28, no. 11, pp. 5005–5015, Nov. 2013, 10.1109/TPEL.2013.2243845.

[6] A. Alganidi and G. Moschopoulos, “A Comparative study of two Passive Regenerative Snubbers for flyback Converters,” in Proc. IEEE ISCAS Conf., Florence, Italy, 2018, pp. 1-4.

[7] K. L. Hwu, Y. T. Yau and Li-Ling Lee, “Powering LED using high-efficiency SR flyback converter,” IEEE Trans. Industrial Applications, vol. 47, no. 1, pp. 376-386, Jan.-Feb. 2011, 10.1109/TIA.2010.2091234.

[8] Rahnamaee, A., Milimonfared, J., Malekian, K., Abroushan, M., “Reliability consideration for a high power zero-voltage-switching flyback power supply,” in Proc. IEEE EPE-PEMC Conf., Poznan, Poland, 2008, pp. 365-371.

[9] M. A. Rezaei, K. Lee and A. Q. Huang, “A high-efficiency flyback micro-inverter with a new adaptive snubber for photovoltaic applications,” IEEE Trans. Power Electronics, vol. 31, no. 1, pp. 318-327, Jan. 2016, 10.1109/TPEL.2015.2407405.

[10] Song-Yi Lin and Chern-Lin Chen, “Analysis and design for RCD clamped snubber used in output rectifier of phase-shift full-bridge ZVS converters,” IEEE Trans. Industrial Electronics, vol. 45, no. 2, Apr. 1998, pp. 358-359, 10.1109/41.681236.

[11] Patel, H. K., “Voltage transient spikes suppression in flyback converter using dissipative voltage snubbers,” in Proc. IEEE ICIEA Conf., Singapore, 2008, pp. 897-901.

[12] Chih-Sheng Liao and Keyue M. Smedley, “Design of high efficiency flyback converter with energy regenerative snubber,” in Proc. IEEE APEC Conf., Austin, TX, USA, 2008, pp. 796-800.

[13] Y. T. Yau, W. Z. Jiang and K. I. Hwu, “Light-load efficiency improvement for flyback converter based on hybrid clamp circuit,” in Proc. IEEE ICICT Conf., Taipei, Taiwan, 2016, pp. 329-333.

[14] M. Mohammadi, E. Adib and H. Farzanchfard, “Passive lossless snubber for double-ended flyback converter,” IET Power Electronics, vol. 8, no. 1, pp. 56-62, 2015, 10.1049/iet-pel.2013.0862.

[15] M. Mohammadi and M. Ordonez, “flyback lossless passive snubber,” in Proc. IEEE ECCE Conf., Montreal, QC, Canada, 2015, pp. 5896-5901.

[16] T.-H. Ai, “A novel integrated nondissipative snubber for flyback converter,” in Proc. IEEE ICSSC Conf., Sousse, Tunisia, 2005, pp. 66–71.

[17] A. Abramovitz, C. Liao and K. Smedley, “State-plane analysis of regenerative snubber for flyback converters,” IEEE Trans. on Power Electronics, vol. 28, no. 11, pp. 5323-5332, Nov. 2013, 10.1109/TPEL.2013.2243845.

[18] R. Lin and S. Hsu, “Design and Implementation of Self-oscillating flyback converter with efficiency enhancement mechanisms,” IEEE Trans. on Industrial Electronics, vol. 62, no. 11, pp. 6955-6964, Nov. 2015, 10.1109/TIE.2015.2436880.

[19] A. Alganidi, A. Abosnina and G. Moschopoulos, “A comparative study of DC-DC flyback converters for telecom applications,” in Proc. IEEE INTELEC Conf., Broadbeach, QLD, Australia, 2017, pp. 424-431.

[20] E. Dzhunusbekov and S. Orazbayev, “A new passive lossless snubber,” IEEE Trans. on Power Electronics, vol. 36, no. 8, pp. 9263-9272, Aug. 2021, 10.1109/TPEL.2021.3056189.

Y. T. YAU was born in Taiman, Taiwan, in 1980. He received the B.S. and M.S. degrees in electrical engineering from Tamkang University, Tamsui, Taiwan, in 2002 and 2004, respectively, and the Ph.D. degree in electrical engineering from the National Taipei University of Technology, Taipei, Taiwan, in 2012. In 2002, he was with ABeL Company for six months. From 2005 to 2011, he was a Researcher with the Industrial Technology Research Institute, Hsinchu, Taiwan. From 2011 to 2014, he was a Senior Engineer with Leadtrength Technology Corporation and Advanced Analog Technology. From 2014 to 2020, he was an Engineer Asian Power Devices Inc., Taoyuan, Taiwan. He is currently an assistant professor in National Chin-Yi University of Technology, Taichung, Taiwan. His research interests include power electronics, converter topology, and digital control.

Tsung-Liang Hung received the B.S. and M.S. degree in electrical engineering from the National Yulin University of Science and Technology, Yunlin, Taiwan, in 2000 and 2002. From 2002 to 2013, He was a supervisor and manager with Ambit, Foxconn, Ampower and Leadtrength Technology Corporation. He is a manager in Technology Center with Asian Power Devices Inc. His research interests include power electronics, converter topology and control behavior.

Taichung, Taiwan. His research interests include power electronics, converter topology and control behavior.