Analysis, design and implementation of a zero voltage switching two-switch CCM flyback converter

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Abstract: This study presents a two-switch continuous conduction mode pulse width modulation flyback converter that employs an LC snubber circuit. The snubber circuit is used to achieve zero voltage switching (ZVS) operation for the main switches during the turn-off transition and soft switching for power diodes. With the proposed LC snubber, the resonant circuit consists of a resonant inductor, a resonant capacitor and two diodes. The operating principles, theoretical analysis and the design methodology of ZVS two-switch continuous conduction mode (CCM) flyback converter are presented. A 200 W (50 V/4 A) laboratory prototype of the proposed converter, operating at a switching frequency of 300 kHz is built to verify the theoretical analysis. At full load the efficiency is 94%.

1 Introduction

Low element count, isolated, design simplicity and simple control circuit are advantages of the flyback converter. However the main drawbacks of the single switch flyback converter are:

- The high voltage spike on switch at turn-off state due to resonance of the transformer leakage inductance and the switch output capacitance that causes losses and electromagnetic interference (EMI) noise.
- The high voltage stress on switch at turn off, that is the sum of input voltage, reflected output voltage and voltage spike that cause selecting the high voltage switch, increases conduction losses.
- The switch is operated at hard switching.
- The rectifier diode is operated at hard switching and the reverse recovery loss is high.

To overcome these problems, it is desirable to utilise soft-switching techniques such as zero voltage switching (ZVS) or zero-current switching (ZCS) on the main switch and the passive elements, for damping the transformer leakage inductance effects, soft switching and reduce the losses.

An RCD snubber across the main switch or across the flyback transformer is introduced in [1, 2]. This snubber consists of a resistor, a diode and a capacitor. This configuration is simple, and limits the high voltage spikes that are caused by leakage inductance at switch turn-off state. However the energy stored in the capacitor dissipates on the snubber resistor, and the converter is operated at hard switching. Thus the efficiency suffers.

In [3], the RCD snubber is optimised, to minimise the dissipative energy in the snubber circuit. However the switching losses is hard as well.

To overcome the drawbacks of the RCD snubber, the quasi-resonant flyback converters have been presented in [4–7], to reduce the switching losses and increase the efficiency. At these converters, the variable frequency control (VFC) is used, and the switches are turned on under ZVS. The ZVS is generated by the resonance between the transformer magnetising inductance and the parasitic capacitance of the switch that the switching loss is reduced perfectly. However the quasi-resonant flyback converters have the following drawbacks:

- Under light load or high-input voltage, the switching frequency is high, which will increase the EMI noise.

To reduce switching losses and high voltage and current stresses on the switch, the PWM active clamp flyback converters are proposed [8–12]. The active clamp snubber consists of a switch and a capacitor parallel to the switch or the transformer. This snubber can recycle the magnetic energy in the leakage inductance and achieve ZVS for the main switch at turning on and off, without additional high voltage and current stresses. The converters in [8, 9] have good efficiency at full load, but the source of the auxiliary switch is float, and it is needed for the complex gate driver circuit. In [10], a two-transformer active clamp flyback converter is proposed to reduce the turn-on switching loss and achieve ZVS. By utilising the two transformers, the ZVS is achieved for the main and auxiliary switches. However, the ZVS at light load is poor. In [11], both series resonant and active clamp techniques are introduced to obtain high efficiency. The active clamp circuit is used for providing a ZVS at turn-on state, and clamps the switch voltage stresses. Moreover, the series resonant circuit provides a ZCS at turn-off. However two soft switching techniques also increase the cost and make the implementation of converter difficult.

In application the flyback converter output voltage is not too many times higher than the rectifier diode forward voltage drop, and the output current is high; the diode rectifier conduction losses are also high. Hence the diode is replaced with a low on-resistance MOSFET [13–15]. By utilising this technique, the rectification losses can be reduced and efficiency is increased. However, the rectifier switch gate drive voltage is proportional to the input voltage and the switching is hard.

To overcome the high voltage stress on the switch and to reduce the size of the switch in the single switch flyback converter, the two-switch PWM flyback converter is introduced [16–19]. From this technique, the voltage across each switch is limited to the input voltage by two clamping diodes. Clamping diodes recycle the leakage energy to input source. However at two switch flyback converters [16, 17], the switching is hard. In [18], the ZVS is sensitive to parameter variations. Moreover the control circuit is complex. In [19], an auxiliary circuit consists of a switch and a
Fig. 1 Schematic of the proposed converter

The ZVS-CCM-PWM two-switch flyback converter operation stages and the equivalent circuits are analysed in Section 2. The design guideline equations are described in Section 3. In Section 4, the experimental results and waveforms for a 200 W (50 V/4 A) laboratory prototype of the proposed converter at a frequency of 300 KHz validate the theoretical analysis. In Section 5, the power losses are considered and estimated at full load. Finally, in Section 6, the conclusion is presented.

2 Proposed converter operation and analysis

The converter has six time intervals in a switching cycle at steady-state operation. Theoretical waveforms of the proposed converter are shown in Fig. 2, and the equivalent circuits for each time intervals are shown in Figs. 3a–f.

To simplify the steady-state analysis, the following assumptions are made:

- The switches and diodes are ideal.
- The inductors and the capacitors are ideal without any parasitic elements.
- The magnetising inductance is large enough to assume that the magnetising inductance current \( i_{L_m} \) is constant.
- The capacitor \( C_o \) is large enough to keeps the output voltage constant.

2.1 Interval-1 \([t_0 < t < t_1]\)

At the moments prior to \( t = t_0 \), the resonant capacitor voltage is \( v_{C_r} (t) = V_s \) and the resonant inductance current is \( i_{L_r} (t) = i_{L_1} (t) = 0 \). The switches \( Q_1 \) and \( Q_2 \) are off, and the energy at magnetising inductance \( i_{L_m} \) is delivered to the output by the ideal transformer and rectifier diode \( D_r \). At \( t = t_0 \), the switches \( Q_1 \) and \( Q_2 \) turn on under exactly ZCS by leakage inductance \( L_d \). The

Table 1 Comparison of reference converters and proposed ZVS flyback converter

<table>
<thead>
<tr>
<th>Conversion Method</th>
<th>Turn-on Switching</th>
<th>Turn-off Switching</th>
<th>Rectifier Diode Switching</th>
<th>Switch Voltage Stress</th>
<th>Switch Current Stress</th>
<th>Control System</th>
<th>Efficiency</th>
</tr>
</thead>
<tbody>
<tr>
<td>RCD flyback [1–3]</td>
<td>hard</td>
<td>hard</td>
<td>hard</td>
<td>limited</td>
<td>limited</td>
<td>simple</td>
<td>low</td>
</tr>
<tr>
<td>Resonant flyback [4–7]</td>
<td>almost ZVS</td>
<td>ZVS</td>
<td>hard</td>
<td>limited</td>
<td>limited</td>
<td>simple</td>
<td>low</td>
</tr>
<tr>
<td>Active clamp flyback [8–12, 18]</td>
<td>hard</td>
<td>ZVS</td>
<td>hard</td>
<td>high</td>
<td>high</td>
<td>complex</td>
<td>high</td>
</tr>
<tr>
<td>Synchronous flyback [13–15]</td>
<td>hard</td>
<td>hard</td>
<td>ZVS</td>
<td>limited</td>
<td>limited</td>
<td>complex</td>
<td>medium</td>
</tr>
<tr>
<td>Two-switch flyback [16, 17]</td>
<td>hard</td>
<td>hard</td>
<td>none</td>
<td>none</td>
<td>none</td>
<td>simple</td>
<td>medium</td>
</tr>
<tr>
<td>ZCT flyback [19]</td>
<td>ZCS</td>
<td>ZCS</td>
<td>ZCS</td>
<td>none</td>
<td>none</td>
<td>complex</td>
<td>high</td>
</tr>
<tr>
<td>Proposed ZVS flyback</td>
<td>ZCS</td>
<td>ZVS</td>
<td>ZCZVS</td>
<td>none</td>
<td>none</td>
<td>simple</td>
<td>high</td>
</tr>
</tbody>
</table>
During this resonance, $C_r$ and $L_d$ starts via the path $C_r = D_r - L_d - O_1$ under the $V_s$. During this resonance, $C_r$ discharges on $L_d$ and the resonant inductance current $i_{Q_r}(t)$ increases, $i_{Q_l}(t)$ and $i_{Q_r}(t)$ rise, and $D_r$ current falls simultaneously. The voltage across the leakage inductance is $V_s + nV_o$. The currents $i_{Q_l}(t)$ and $i_{Q_r}(t)$ can be written as

$$i_{Q_l}(t) = i_{Q_r}(t) = \frac{V_s + nV_o}{L_d} (t - t_0)$$

where $n = n_1/n_2$ is turns ratio of the transformer.

At this interval, the following equations for snubber elements can be written

$$i_L(t) = V_s \sin \omega_s (t - t_0)$$

and

$$v_C(t) = V_s \cos \omega_s (t - t_0)$$

where

$$Z_r = \sqrt{L_d/C_r}$$

and

$$\omega_s = \frac{1}{\sqrt{L_d C_r}}$$

The $O_1$ current and $i_{Q_r}(t)$ can be written as

$$i_{Q_r}(t) = \frac{V_s + nV_o}{L_d} (t - t_0) + \frac{V_s}{Z_r} \sin \omega_s (t - t_0)$$

and

$$i_{Q_l}(t) = nI_{m_o} - \frac{nV_o + n^2 V_s}{L_d} (t - t_0)$$

At $t = t_1$, $i_{Q_l}(t)$ reaches $I_{m_o}$ and $D_r$ current drops to zero; thus the rectifier diode turns off under exactly ZCZVS, and this stage finishes. The duration of this stage can be written as

$$\Delta t_1 = t_1 - t_0 = \frac{I_{m_o} L_d}{V_s + nV_o}$$

### 2.2 Interval-2 $[t_1 < t < t_2]$

The resonance between $C_r$ and $L_d$ still occurs during this stage. Since $i_{Q_r}(t) = I_{m_o}$, the diode $D_r$ is off. The magnetic inductance is charged by the $V_s$. During this resonance, $C_r$ is discharged to zero first, then precharged to $-V_s$ and $i_{Q_r}(t)$ falls to zero at $t = t_2$. Therefore $D_r$ turns off under ZCZVS, and this stage finishes.

The time interval of this stage is given as follows

$$\Delta t_2 = t_2 - t_1 = \frac{\pi}{\omega_s} - \Delta t_1 = \frac{\pi}{2} - \Delta t_1$$

### 2.3 Interval-3 $[t_2 < t < t_3]$

This stage begins at $t = t_2$, when $v_C(t) = -V_s$ and $i_{Q_r}(t) = i_{Q_l}(t) = I_{m_o}$. Since $v_C(t) = -V_s$, the power switches are turned off under exactly ZVS condition at $t = t_3$. This stage finishes at $t = t_3$, when the snubber diode $D$ is forced on, under ZVZCS by $I_{m_o}$. The duration of this stage is given by

$$\Delta t_3 = D_s T_s - \Delta t_2 - \Delta t_1$$

where $D_s$ is the duty cycle of the control signal and $T_s = 1/f_s$ is the switching period.

### 2.4 Interval-4 $[t_3 < t < t_4]$

During this stage the negative voltage across $C_r$ keeps the diode $D_r$ off. The snubber diode $D$ is on and $C_r$ is charged to $V_s$ by magnetic inductance current $I_{m_o}$ at $t = t_3$. However, clamping diodes $D_{m_1}$ and $D_{m_2}$ turn on under ZCZVS, and the voltages across the parasitic capacitance of the power switches are clamped at $V_s$, $v_C(t)$ for this interval is derived as

$$v_C(t) = \frac{I_{m_o}}{C_r} (t - t_3) - V_s$$

The time duration of this stage can be written as

$$\Delta t_4 = t_4 - t_3 = \frac{2V_s C_r}{I_{m_o}}$$
2.5 Interval-5 \([t_4 < t < t_5]\)

During this interval, the transformer starts to reset by \(D_{m1}\) and \(D_{m2}\). Thus \(i_L(t)\) decreases. Since \(v_C(t)\) equals \(V_s\), thus \(i_D(t)\) continues to increase. The important equations for this stage are:

\[
i_L(t) = i_{Dm1}(t) = i_{Dm2}(t) = I_{in} - \frac{(V_s - nV_o)}{L_d}(t - t_4) \quad (13)
\]

\[
i_D(t) = \frac{n(V_s - nV_o)}{L_d}(t - t_4) \quad (14)
\]

At \(t = t_5\), \(i_D(t)\) equals the reflected magnetising inductance current \(nI_L\); thus the clamping diode currents \(i_{Dm1}(t)\) and \(i_{Dm2}(t)\) are from null to zero. This moment, \(D_{m1}\) and \(D_{m2}\) turn off under ZCZVS, and this stage finishes. The duration of this interval is

\[
\Delta t_5 = t_5 - t_4 = \frac{I_{in}L_d}{(V_s - nV_o)} \quad (15)
\]

2.6 Interval-6 \([t_5 < t < t_6]\)

The proposed converter operates as the turn-off state of a conventional PWM flyback converter at this interval. However the magnetising energy of \(L_m\) discharges on load. At \(t = t_6\), one switching cycle is completed and another switching cycle begins.

3 Design procedure

3.1 Voltage gain

Referring to the voltage waveform \(v_{L_m}(t)\) in Fig. 2, the volt-second balance can determine the voltage gain for the proposed converter as

\[
V_s(\Delta t_4 + \Delta t_5 + \Delta t_6) = (V_o)(\Delta t_1 + \Delta t_5 + \Delta t_6) \quad (16)
\]

Assuming that the time duration of stages 1, 2 and 4 are very small in comparison with those of stages 3, 5 and 6, the DC voltage transfer
function is approximately

\[ M = \frac{V_o}{V_s} = \frac{D_s}{m(1 - D_s)} \]  

(17)

which is the same as that of the conventional flyback converter.

### 3.2 Maximum values of the power switches and diodes

At the end of stage 4, the maximum values of power switches off-state voltage are

\[ V_{Q_{1 \text{ (max)}}} = V_{Q_{2 \text{ (max)}}} = V_s \]  

(18)

During stage 2, the maximum value of the current through the power switch \( Q_1 \) is determined by

\[ I_{Q_1 \text{ (max)}} = I_{m} + \frac{V_s}{Z_r} \frac{P_{o \text{ (max)}}}{n V_s (1 - D_s \text{ (max)})} + \frac{V_s \text{ (max)}}{Z_r} \]  

(19)

The maximum value of the \( Q_2 \) current is

\[ I_{Q_2 \text{ (max)}} = I_{m} \]  

(20)

The maximum reverse voltage across the power diode \( D_c \) can be written as

\[ V_{D_c \text{ (max)}} = \frac{V}{n} + V_o \]  

(21)

The maximum value of the power diode current is

\[ I_{D_c \text{ (max)}} = \frac{I_{m \text{ (max)}}}{1 - D_s \text{ (max)}} \]  

(22)

The maximum reverse voltages across clamping diodes \( D_{m1} \) and \( D_{m2} \) is same as the maximum value of off-state voltage across power switches.

During stage 5, the maximum current through the clamping diodes are

\[ I_{D_{m1 \text{ (max)}}} = I_{D_{m2 \text{ (max)}}} = I_{m} \]  

(23)

The maximum values of snubber diodes \( D \) and \( D_r \) reverse voltage are

\[ V_{D_{\text{ (max)}}} = V_{D_{r \text{ (max)}}} = V_s \text{ (max)} \]  

(24)

The maximum value of current through \( D \) is \( I_{D_{\text{ (max)}}} = I_{m} \). During stage 2, the \( D_r \) maximum current is

\[ I_{D_r \text{ (max)}} = \frac{V_s \text{ (max)}}{Z_r} \]  

(25)

### 3.3 Design the \( C_r \) and \( L_r \)

During stage 4, the switches \( Q_1 \) and \( Q_2 \) are turned off, and the snubber capacitor discharges linearly by \( I_{m} \). To ensure switches turn off under exactly ZVS, the discharge time \( \Delta t_4 = \tau_{ZVS} \) should be three times larger than the fall time \( t_f \) of switches. However the snubber capacitor is derived as

\[ C_r > \frac{n I_{m}}{2 V_{m \text{ (min)}} (3 t_f)} \]  

(26)

To minimise the influence of the resonant parameters, the time interval \( t_2 - t_0 = T_r/2 = \pi/\omega_r \) should be less than \( D_{\text{ (min)}}/f_s \). The resonant frequency is derived as

\[ f_r \geq \frac{f_s}{2 D_{(\text{min})}} \]  

(27)

where \( D_{\text{min}} = nM_{\text{min}}/(nM_{\text{min}} + \eta) \) [20].

Fig. 4  Experimental circuit of the proposed ZVS flyback converter
Thus, $L_r$ can be obtained as

$$L_r = \frac{1}{(2\pi f_r)^2} \cdot C_r$$  

(28)

4 Experimental results

To verify the theoretical analysis and design procedure of the proposed converter, an experimental CCM-ZVS flyback converter for the following specifications is designed, implemented and some experimental results are measured.

$V_o = 50$ Vdc; maximum output power: $P_o = 200$ W; nominal input voltage: $V_s = 100$ Vdc; $V_{s(max)} = 110$ Vdc; $V_{s(min)} = 80$ Vdc; $f_s = 300$ kHz; maximum duty cycle: $D_{max} = 0.45$.

A complete circuit of the experimental proposed flyback converter is shown in Fig. 4. IRF3415 MOSFETs are adopted as the power switches. MUR820 diodes are selected as $D_m^1$, $D_m^2$ and $D_c$. The values of resonant components $C_r$ and $L_r$ can be designed using (23) and (25), respectively. They are chosen $C_r = 4.7$ nF and $L_r = 22 \mu$H. TL494L is a voltage-mode PWM controller. The maximum frequency and $D_{max}$ of TL494L are 300 KHz and 45%, respectively. Feedback loop is closed with TLP280 optocoupler.
Thus, the converter output is isolated from the control circuit. The gate drive circuit consists of the HIP2500. It is the high voltage, high speed power MOSFET driver with independent high and low side referenced output channels. The EI33/29/13 ferrite core is selected as a core of the power transformer. The transformer turns ratio is \( n = 1 \). The measured leakage inductance is \( L_d = 4 \mu \text{H} \). The output filter is designed as \( C_f = 24 \mu \text{F} \) for 1 per cent output voltage ripple.

The experimental results are depicted in Figs. 5–9. All waveforms are obtained at \( V_i = 105 \text{ V} \), and the regulated output voltage \( V_o = 50 \text{ V} \). The characteristic impedance is \( Z_r = 68.4 \Omega \). Figs. 5a–c show the power switch \( Q_1 \) current and voltage at full load, half-load and 10% full load, respectively. It can be seen from Fig. 5 that \( Q_1 \) commutates at ZCS turn-on and ZVS turn-off.

As can be noted from Fig. 5a, the maximum current through the \( Q_1 \) is 8.9 A. In addition the maximum voltage \( V_{Q_1} \) is limited to 105 V. Thus voltage and current waveforms agree with (18) and (19), respectively. In Figs. 6a–c is shown the current and voltage waveforms for power switch \( Q_2 \) at full load, half-load and 10% full load, respectively. It can be observed that \( Q_2 \) is turned on under ZCS and turned off under ZVS condition. From Fig. 6a, it is seen that the \( Q_2 \) maximum current is 7.3 A, and \( V_{Q_2} = 105 \text{ V} \).

Fig. 7 shows the experimental waveforms \( v_c(t) \), \( i_{Lr}(t) \) and the reverse voltage across the snubber diode \( D_r \). The \( D_r \) maximum current is 1.4 A, and \( D_r \) maximum reverse voltage is 105 V. They are verified in (24) and (25). The voltage and current waveforms of the rectifier diode is shown in Fig. 8. The results shown in Fig. 8 demonstrate that ZCZVS is achieved for \( D_r \) at turn-on and turn-off transitions. Fig. 9 represents the reverse voltage and current waveforms of clamping diodes. It can be seen that the clamping diodes are turned on and turned off under ZCZVS. It can be observed that the maximum current through the clamping diodes equals \( I_{D_m} = 7.3 \text{ A} \).

From the experimental results, it can be revealed that the theoretical analysis of the proposed ZVS two-switch CCM flyback converter are exactly verified.

### 5 Power losses consideration

The proposed ZVS flyback converter is designed to achieve ZVS and ZCS operation for the main switches during the turn-off and turn-on transition, respectively, and soft switching for diodes. Therefore the switching losses and the reverse recovery loss of the rectifier diode are negligible. The conduction losses are determined here. The on-state resistor of the power switches from the datasheet is 42 m\( \Omega \). The conduction losses of \( Q_1 \) is

\[
P_{C(Q_1)} = r_{d(on)} \left( \frac{V_i}{2sZ_r} \sqrt{\frac{f_s}{f_i}} + \frac{I_o}{n(1-D)} \left( D + \frac{f_s}{2s} \right) \right) \tag{29}
\]

<table>
<thead>
<tr>
<th>Table 2</th>
<th>Calculated losses of proposed ZVS flyback converter and conventional flyback converter</th>
</tr>
</thead>
<tbody>
<tr>
<td>Losses</td>
<td>Proposed ZVS flyback converter</td>
</tr>
<tr>
<td>( Q_1 ) turn-off loss [17]</td>
<td>–</td>
</tr>
<tr>
<td>( Q_1 ) turn-on loss [17]</td>
<td>–</td>
</tr>
<tr>
<td>( Q_2 ) turn-off loss [17]</td>
<td>–</td>
</tr>
<tr>
<td>( Q_2 ) turn-on loss [17]</td>
<td>–</td>
</tr>
<tr>
<td>( Q_2 ) conduction loss (29)</td>
<td>2.9 W</td>
</tr>
<tr>
<td>( Q_2 ) conduction loss (30)</td>
<td>1 W</td>
</tr>
<tr>
<td>( Q_2 ) conduction loss (31)</td>
<td>3.6 W</td>
</tr>
<tr>
<td>( D_r, D_m ) conduction loss (32)</td>
<td>( 2s \times 1 \text{ W} )</td>
</tr>
<tr>
<td>( D_1 ) conduction loss (33)</td>
<td>0.24 W</td>
</tr>
<tr>
<td>( D_2 ) conduction loss (34)</td>
<td>0.23 W</td>
</tr>
<tr>
<td>Transformer copper losses (35)</td>
<td>1</td>
</tr>
<tr>
<td>Transformer core losses [21]</td>
<td>2</td>
</tr>
<tr>
<td>efficiency</td>
<td>94%</td>
</tr>
</tbody>
</table>
Since the time $\Delta t_2$ is quite smaller than $D_2T_s$, the $Q_2$ conduction loss can be calculated by

$$P_{C(Q_2)} = P_{CQ_2} \cdot \left( \frac{I_o}{n(1-D)} \right)^2$$

The MUR820 is used as a rectifier diode and its forward voltage $V_{F(on)} = 0.9$. Since the time $\Delta t_4$ is quite smaller than $(1-D)T_s$, $D_4$ can be calculated by

$$P_{C(D_4)} = V_{F(on)} \cdot I_o$$

The conduction losses of clamping diodes $D_{m1}$ and $D_{m2}$ can be given by

$$P_{C(D_{m1})} = P_{C(D_{m2})} = V_{F(on)} \left( \frac{I_o}{n(1-D)} \frac{\Delta t_5}{T_s} \right)$$

The forward voltage of MUR420 as a snubber diode is $V_{F(on)} = 0.8$. The conduction losses of $D_3$ and $D_5$ can be calculated by

$$P_{C(D_3)} = V_{F(on)} \cdot \frac{V_{F(on)}}{nD_1I_1}$$

$$P_{C(D_5)} = V_{F(on)} \cdot (2V_sC_sI_s)$$

The copper losses of the primary and secondary windings of the transformer are

$$P_{Cu(T)} = P_{Cu(pro)} + P_{Cu(sec)}$$

$$P_{Cu(T)} = \left[ r_{(pro)} \left( \frac{I_o}{n(1-D)} \right)^2 \right] + \left[ r_{(sec)} \left( \frac{I_o}{\sqrt{(1-D)}} \right)^2 \right]$$

where $r_{(pro)} = 15$ m$\Omega$ and $r_{(sec)} = 10$ m$\Omega$ are the measured resistor of the transformer primary and secondary windings respectively.

Transformer core losses for EI33/29/13 ferrite core are obtained from [21]. The maximum flux density is $B_{m} = 0.13$ T.

Fig. 10 Efficiency of the proposed converter (continuous line) and conventional flyback (broken line) versus output power

The comparison of calculated losses for the proposed converter and the conventional two-switch flyback converter operating at full load condition are summarised in Table 2.

Fig. 10 shows the measured efficiency versus various output power for the proposed converter and the conventional flyback converter.

It can be seen that the maximum measured efficiency of ZVS flyback converter is 94% at full load. Thus the measured efficiency is verified by the calculated efficiency at full load. Note that the measured efficiency of the conventional two-switch flyback converter is 91% at full load, and it does not agreed with calculated efficiency. This is due to the diodes reverse recovery losses.

6 Conclusion

In this paper an LC snubber circuit, consisting of a resonant inductor, a resonant capacitor and two diodes, is applied to a two-switch PWM flyback converter, to increases the overall efficiency. The snubber circuit is used to achieve ZVS operation for the main switches during the turn-off transition and soft switching for all-passive semi-conductor devices. With the proposed LC snubber, the magnetic energy in the transformer leakage inductance can be fully recycled and transferred to the input side. Thus the switching frequency can be increased. The proposed converter is presented, and its operating principle is described in detail. For 100 V input and 50 V, 200 W output, a prototype of the proposed ZVS two-switch flyback converter with CCM operation, operating at 300-kHz, is implemented. The theoretical analysis of the proposed converter has been verified with experimental results.

Note: the experimental results and the features of the proposed converter can be summarised as follows:

(i) Both switches of the ZVS flyback converter are exactly turned off with ZVS, and are turned on under ZCS. Snubber diodes $D_s$ and $D_t$ are turned on and off under ZVZCS.

(ii) The leakage inductance magnetic energy is absorbed by the resonance capacitor.

(iii) The control system is pulse width modulation, hence it is very easy and cheap.

(iv) The converter acts as a conventional two-switch PWM flyback converter during most of the time, because the resonant cycle is very short during both the turn-on and turn-off transitions.

7 References


