The Over-Power Phenomenon in DCM/CCM-Operated Flyback Converters (Part 4): The Leakage Inductor Contribution

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In previous issues, we have shown that the output power runaway in a flyback converter operated at high line can be attributed to two variables: the propagation delay only for DCM-operated converters and a combination of the propagation delay plus the mode transition from continuous to discontinuous in CCM-operated power supplies. During the derivation of the power transfer equations, we have estimated various values for the total converter efficiency based on experience with real application circuits. Rather than guessing the efficiency, here in this fourth and final part of this article series, we calculate the theoretical maximum transmitted power by explaining the role of the transformer leakage inductance at the switch opening.

Power Transfer In A Flyback Converter

A flyback converter operates by storing and releasing energy on a cycle-by-cycle basis. Unlike other converters such as buck or boost, the energy taken from the source is not directly transmitted to the load during the power switch on-time: during this event, the current in the primary inductor L_p builds up while the secondary side is disconnected from the load by a blocking diode. When the controller instructs the power MOSFET to turn-off, a certain amount of energy has been stored in the transformer primary inductor. Through magnetic coupling, this energy will be transferred to the secondary side and will feed current to both the output capacitor and the load via the diode now conducting.

If all the stored energy is depleted during the off-time, the converter operates in discontinuous conduction mode. However, if some energy remains in the transformer at the end of the off-time, the converter operates in continuous conduction mode. Unfortunately, the conversion process involving the transformer is not perfect.

Fig. 1 portrays the so-called PI model of a transformer used in a flyback converter. You can identify the winding dots placed on opposite positions, typical of the flyback arrangement. As a current "entering" the primary dot must leave the secondary side by its dot, we will see in a few lines that the output diode presence imposes current circulation either in the primary or in the secondary but not simultaneously. For this reason, the magnetic arrangement in a flyback converter is often referred to as coupled inductors rather than as a true transformer.

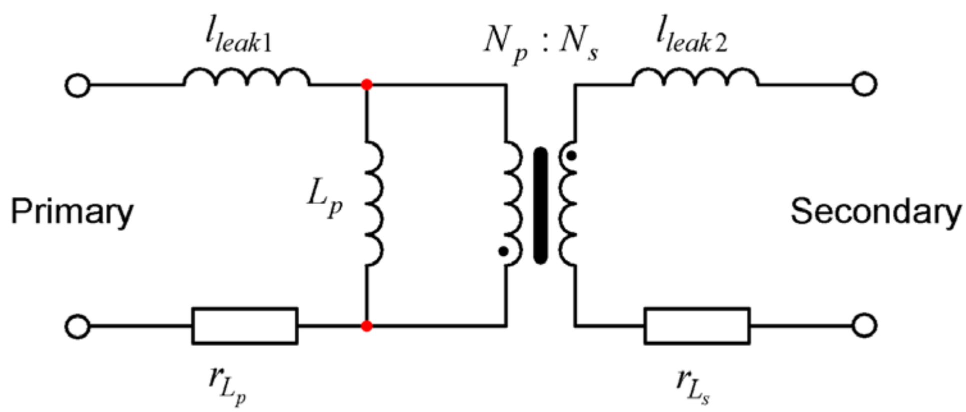

Fig. 1. The PI-model of a two-winding transformer shows leakage elements on both sides.

In this model, we can identify several elements:

- L_p is the magnetizing inductor, also called the primary inductor. The current that circulates in the primary energizes the magnetic core by aligning its ferrous particles. When the core is energized, both the primary and the secondary sides are coupled. In a flyback converter, despite the circulation of a primary current, there is no current in the secondary side because the windings arrangement blocks the secondary-side diode.

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- L_{leak1} and L_{leak2} are symbols that signify the imperfect coupling between the primary and the secondary sides. In other words, some induction lines on the primary side as well as on the secondary side, do not close in the opposite winding but in the air—this is a leak. This leak is expressed by the presence of the series inductors L_{leak1} and L_{leak2}. As L_{leak1} and L_p appear in series, a current crossing L_p is also flowing in L_{leak1}.
- N_p:N_s represents the effective turns ratio between both windings. For convenience, this ratio is often normalized to the primary and noted N. For instance, if $N = 0.05$, it implies a turns ratio of 1:0.05, meaning that the secondary winding has an amount of turns equal to that of the primary divided by 20.
- r_{L_p} and r_{L_s} are resistors representing the dc conduction losses of the copper wires.

This model can be further complicated by adding capacitors or extra terms such as core losses or non-linear behaviors. On the other hand, we will use a simpler version of this PI-model consisting of a single leakage term l_{leak} and one primary inductor L_p. We can further simplify the approach by considering a turns ratio of 1, without isolation at all. As such, Fig. 2 represents a flyback converter during the on-time where both L_p and l_{leak} clearly appear in series. V_{clamp} portrays the clamping voltage provided by the *RCD* network or the transient voltage suppressor (TVS) present in most commercial flyback designs.

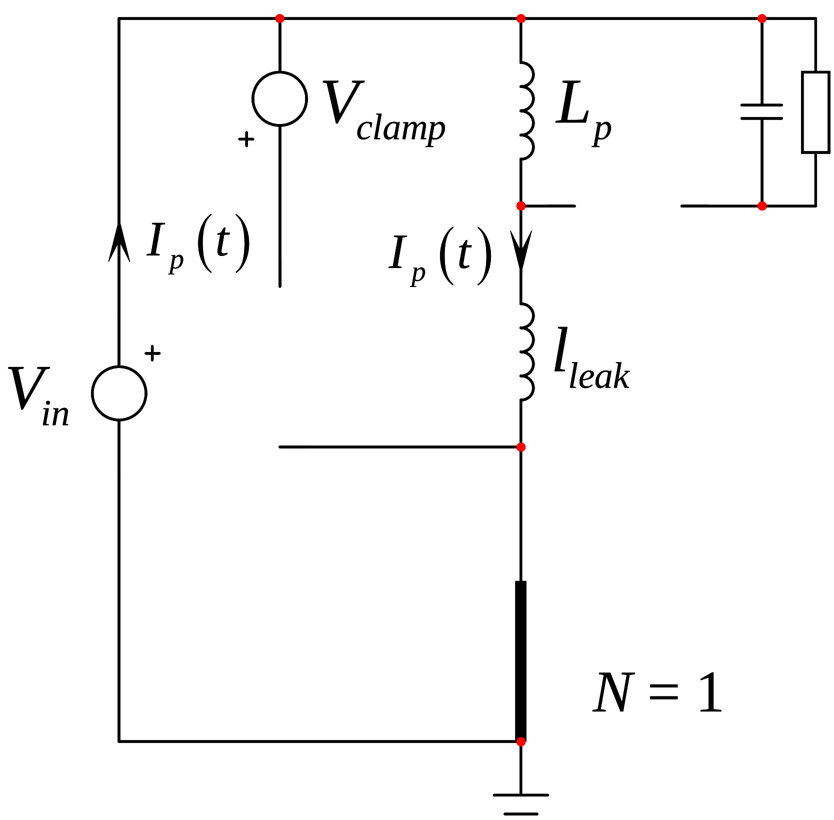

Fig. 2. The flyback converter during the on-time energizes the transformer primary inductor but also the leakage inductor.

During the on-time, a current I_p flows in the transformer primary and ramps up with a slope defined by:

$$S_{on} = \frac{V_{in}}{L_p + l_{leak}} \tag{1}$$

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No current circulates in the secondary side as the dot arrangement of the transformer reverse-biases the output diode. When the primary-side peak current reaches the target imposed by the controller, I_{peak}, the power MOSFET opens. At this time, the primary voltage reverses in an attempt to keep the amperes-turns constant. The secondary-side diode now conducts, offering a path to the primary current $I_p(t)$ through the coupling of both windings. Since the diode conducts, the output voltage V_{out} plus the forward voltage drop of the rectifier, V_f, are reflected across the primary inductor, forcing the current to decrease with a slope of:

$$S_{off} = \frac{V_{out} + V_f}{L_p} \tag{2}$$

However, the current suddenly interrupted in the leakage inductor also keeps flowing in the same direction. To avoid any discontinuity, besides the capacitance on the drain, a safe path has to be offered to this current. In a classical design, it will go to the low-impedance voltage source V_{clamp}, depleting the energy stored in the leakage inductor. As confirmed by Fig. 3, the leakage inductor undergoes a voltage equal to V_{clamp} minus the reflected voltage (V_{out} + V_f) imposing a downslope equal to:

$$S_{l_{leak}} = \frac{V_{clamp} - \left(V_{out} + V_f\right)}{l_{leak}} \tag{3}$$

As both inductors connect in series, a current circulating in the leakage inductor implies the same current flowing in the primary inductor. Since the primary inductor is feeding the secondary side, the leakage inductor diverts some of the stored energy, wasting it in heat through the clamping network. As indicated in Fig. 3, the secondary current is equal to:

$$I_{\sec}(t) = I_{L_p}(t) - I_{l_{leak}}(t) \tag{4}$$

Both currents being equal at the switch opening, the secondary-side current ramps from zero and grows until the leakage inductor current is fully reset. This reset time, Δt, delays the occurrence of the current on the secondary side of the transformer at the switch turn-off event.

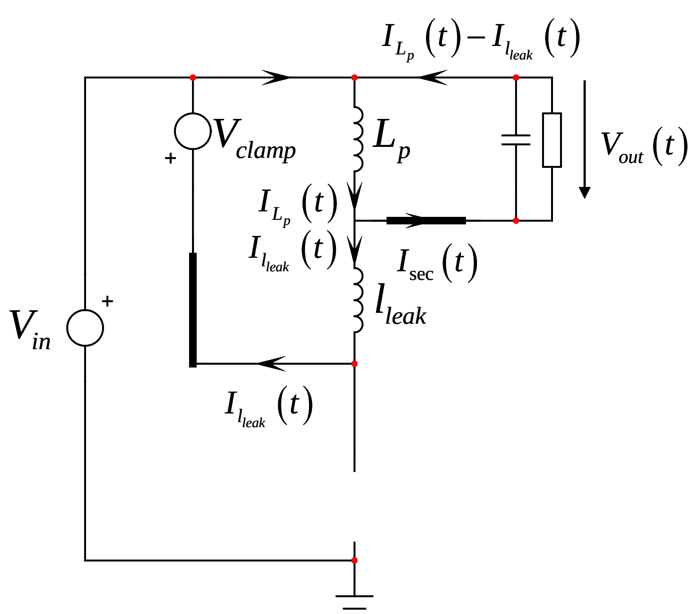

Fig. 3. During the off-time, as a current also flows in the leakage inductor, it diverts some of the energy stored in the primary inductor until the leakage inductor reset occurs.

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When the leakage inductor is reset, the secondary-side current is equal to the current in the primary inductor:

$$I_{\text{sec}}(t) = I_{L_p}(t) \tag{5}$$

Unfortunately, at this moment, because the leakage inductor dissipates in heat a portion of the primary energy, the primary current is no longer equal to I_{peak}, the value at which the MOSFET opens: it has decreased to a lower value, designated as point A in Fig. 4. This is the effective secondary-side peak current value. The off-time then lasts until the primary inductor current either reaches zero (DCM) or its valley value in the case of CCM. The downslope is calculated by equation (2).

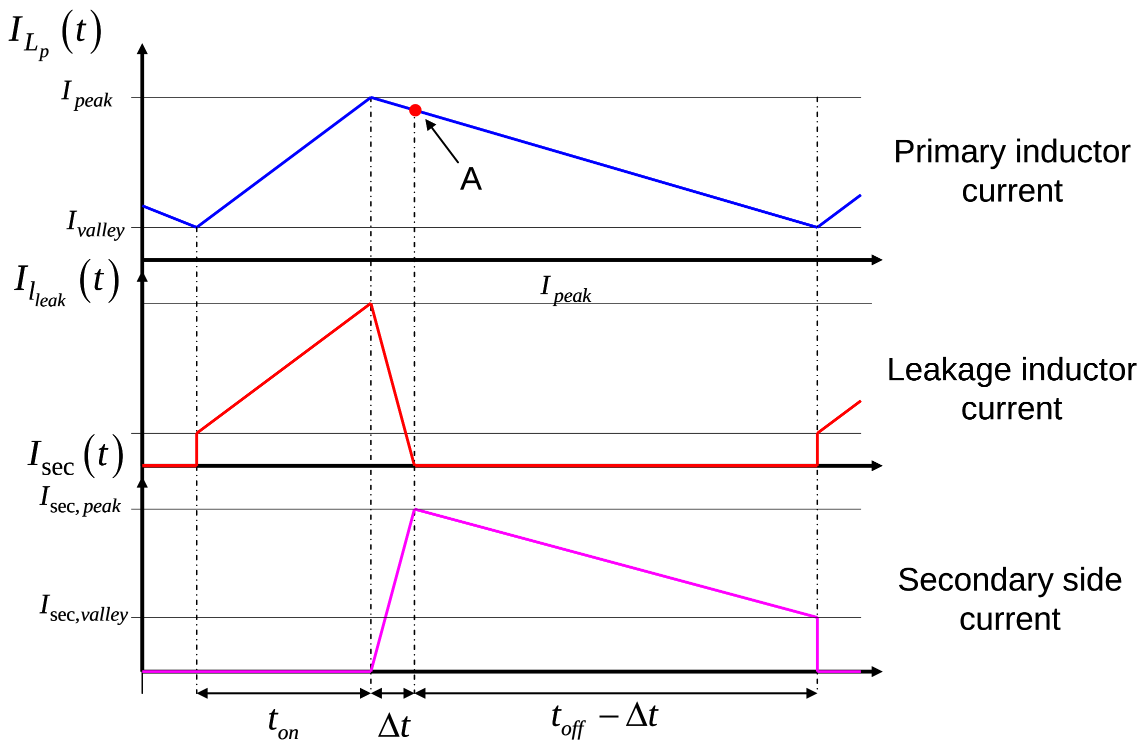

Fig. 4. The leakage inductor needs time to be reset. This time actually delays the current transfer from primary to secondary.

The Average Output Current

Capitalizing on Fig. 4 waveforms, we can now derive a few equations to estimate the maximum dc output current flowing in the output diode when the controller detects a fault condition. First off, we need to estimate the leakage inductor reset time, Δt. We know that the inductor peaks to I_{peak} before dropping to zero with a slope given by equation (3). Therefore, we have:

$$\Delta t = \frac{I_{peak}}{S_{l_{leak}}} = \frac{l_{leak} I_{peak}}{V_{clamp} - \left(V_{out} + V_f\right)} \tag{6}$$

If we now consider a transformer turns ratio N different from 1, this equation becomes:

$$\Delta t = \frac{l_{leak} I_{peak}}{V_{clamp} - \left(V_{out} + V_f\right)/N} \tag{7}$$

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Point A in Fig. 4 is the point reached by I_{peak} in Δt when falling down at a slope S_{leak}:

$$I_A = I_{peak} - S_{off}\Delta t \tag{8}$$

Using equations (7) and (3), we can update and derive an expression for the peak secondary current when the leakage inductor is reset:

$$I_A = I_{peak} - \frac{\left(V_{out}+V_f\right)}{NL_p}\frac{Nl_{leak}I_{peak}}{NV_{clamp}-\left(V_{out}+V_f\right)} \tag{9}$$

If we rearrange this equation and scale the current at point A by N to the secondary side, we obtain:

$$I_{\sec,peak} = \frac{I_{peak}}{N}\left(1-\frac{l_{leak}}{L_p}\frac{1}{\frac{NV_{clamp}}{V_{out}+V_f}-1}\right) \tag{10}$$

If the converter is operated in CCM, the secondary valley current does not change since there is no contribution from the leakage term anymore:

$$I_{\sec,valley} = \frac{I_{valley}}{N} \tag{11}$$

Having these two formulas on hand, we can now calculate the secondary-side ripple current ΔI_L:

$$\Delta I_{L,\sec} = I_{\sec,peak} - I_{\sec,valley} \tag{12}$$

To help calculate the average output current, Fig. 5 offers a magnified view of the secondary-side current during the off-time.

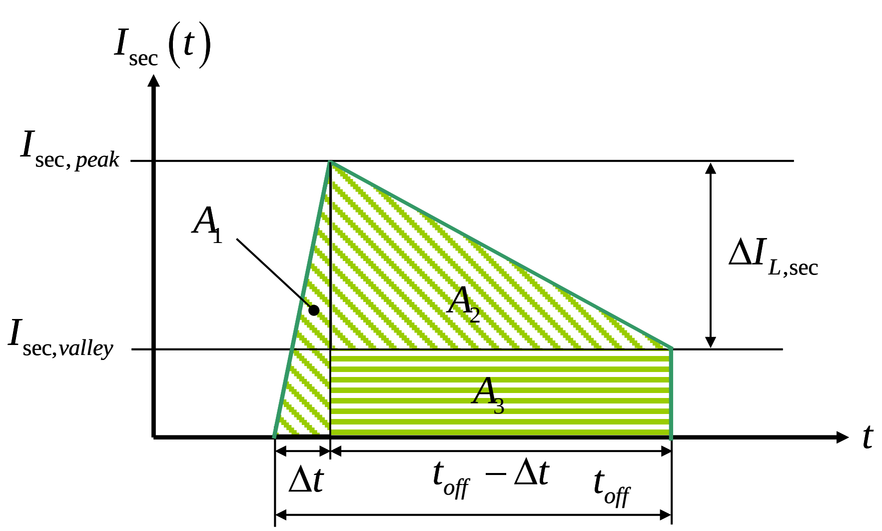

Fig. 5. The average current on the secondary side is obtained by dividing the current signal into surfaces, individually calculated.

To estimate the average value of such a waveform, the simplest way is to slice it into easily calculable surfaces and sum them together before dividing by the signal period:

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$$I_{\sec,avg} = \frac{A_1 + A_2 + A_3}{T_{sw}} \qquad (13)$$

Now using equations (7), (10), (11) and (12), we derive the following expression for the dc current circulating in the output diode:

$$I_{\sec,avg} = \frac{\frac{I_{\sec,peak}\Delta t}{2} + \frac{\Delta I_L\left(t_{off} - \Delta t\right)}{2} + I_{\sec,valley}\left(t_{off} - \Delta t\right)}{T_{sw}} \qquad (14)$$

This expression is the dc current flowing in the output diode and delivered to the load. The ac portion of the Fig. 5 signal goes to the output capacitor. In a fault condition (an overload situation, for instance), the controller pushes the primary-side peak current to its maximum. We have seen that if the maximum allowed sensed-voltage is V_{sense}, then the maximum peak current in the transformer at the switch opening is:

$$I_{p,\max} = \frac{V_{sense,\max}}{R_{sense}} + \frac{V_{in}}{L_p} t_{prop} \qquad (15)$$

where V_{sense} is fixed at 800 mV in an NCP1250 from ON Semiconductor, R_{sense} is the external sense resistor, V_{in} is the input voltage (low line, $V_{in,LL}$ and high line $V_{in,HL}$) and t_{prop} is the total propagation delay of the controller, including the driving path to the MOSFET. With all these elements on hand, we can now compute the maximum power in a numerical example.

Computing the Transmitted Power

Let's assume we have built a 65-W converter featuring the following component values and operating voltages:

$V_{in,HL}$ is the dc input voltage at high line, 370 V

$V_{in,LL}$ is the dc input voltage at low line, 120 V

V_{out} is the dc output voltage, 19 V

L_p is the transformer primary inductance, 600 µH

l_{leak} is the transformer leakage inductance, 2 µH

t_{prop} is the total propagation delay, 350 ns

R_{sense} is the sense resistor, 0.33 Ω

V_{sense} is the maximum authorized sense value, 0.8 V

T_{sw} is the switching period for a 65-kHz frequency, 15.4 µs

N is the turns ratio between the secondary and the primary sides, 0.25

V_{clamp} is the selected clamping level on the *RCD* network, 107 V

η_{LL} is the converter efficiency at low line, 85%

η_{HL} is the converter efficiency at high line, 89%

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First, we can calculate the peak current in a fault condition at low line and high line:

$$I_{peak,\max,LL} = \frac{V_{sense,\max}}{R_{sense}} + \frac{V_{in,LL}}{L_p} t_{prop} = \frac{0.8}{0.33} + \frac{120}{600u} \times 350n = 2.49\,\text{A} \tag{16}$$

$$I_{peak,\max,HL} = \frac{V_{sense,\max}}{R_{sense}} + \frac{V_{in,HL}}{L_p} t_{prop} = \frac{0.8}{0.33} + \frac{370}{600u} \times 350n = 2.64\,\text{A} \tag{17}$$

Next is the valley-current calculation, changing between the two input-line conditions. It has been derived in part 2 of this series of articles:

$$I_{valley,LL} = I_{peak,\max,LL} - \frac{T_{sw} V_{in,LL} \left(V_f + V_{out}\right)}{L_p \left(V_f + V_{out} + N V_{in,LL}\right)} = 2.49 - \frac{15.4u \times 120 \times (19 + 0.5)}{600u \times (19 + 0.5 + 0.25 \times 120)} = 1.28\,\text{A} \tag{18}$$

$$I_{valley,HL} = I_{peak,\max,HL} - \frac{T_{sw} V_{in,HL} \left(V_f + V_{out}\right)}{L_p \left(V_f + V_{out} + N V_{in,HL}\right)} = 2.64 - \frac{15.4u \times 370 \times (19 + 0.5)}{600u \times (19 + 0.5 + 0.25 \times 370)} = 0.99\,\text{A} \tag{19}$$

Equation (14) requires the value of the off-time at both input conditions. It can easily be calculated using primary values as follows:

$$t_{off} = \frac{N\left(I_{peak,\max} - I_{valley}\right)}{\left(V_{out} + V_f\right)} L_p \tag{20}$$

At the two extremes, t_{off} becomes:

$$t_{off,LL} = \frac{N\left(I_{peak,\max,LL} - I_{valley,LL}\right)}{\left(V_{out} + V_f\right)} L_p = \frac{0.25 \times (2.49 - 1.28)}{19 + 0.5} \times 600u = 9.3\,\mu\text{s} \tag{21}$$

$$t_{off,HL} = \frac{N\left(I_{peak,\max,HL} - I_{valley,HL}\right)}{\left(V_{out} + V_f\right)} L_p = \frac{0.25 \times (2.64 - 0.99)}{19 + 0.5} \times 600u = 12.7\,\mu\text{s} \tag{22}$$

We now need to calculate the leakage inductor reset time Δt under both input-line conditions:

$$\Delta t_{LL} = \frac{N l_{leak} I_{peak,\max,LL}}{N V_{clamp} - \left(V_{out} + V_f\right)} = \frac{0.25 \times 2u \times 2.49}{0.25 \times 107 - (19 + 0.5)} = 171\,\text{ns} \tag{23}$$

$$\Delta t_{HL} = \frac{N l_{leak} I_{peak,\max,HL}}{N V_{clamp} - \left(V_{out} + V_f\right)} = \frac{0.25 \times 2u \times 2.64}{0.25 \times 107 - (19 + 0.5)} = 182\,\text{ns} \tag{24}$$

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From these numbers, we can evaluate the secondary-side peak current when the leakage inductor is reset:

$$I_{\sec,peak,LL} = \frac{I_{peak,\max,LL}}{N}\left(1-\frac{l_{leak}}{L_p}\frac{1}{\frac{NV_{clamp}}{V_{out}+V_f}-1}\right) = \frac{2.49}{0.25}\times\left(1-\frac{2u}{600u}\times\frac{1}{\frac{0.25\times107}{19+0.5}-1}\right) = 9.9\,\text{A} \quad (25)$$

$$I_{\sec,peak,HL} = \frac{I_{peak,\max,HL}}{N}\left(1-\frac{l_{leak}}{L_p}\frac{1}{\frac{NV_{clamp}}{V_{out}+V_f}-1}\right) = \frac{2.64}{0.25}\times\left(1-\frac{2u}{600u}\times\frac{1}{\frac{0.25\times107}{19+0.5}-1}\right) = 10.4\,\text{A} \quad (26)$$

From the above equations, we can derive the secondary-side ripple current:

$$\Delta I_{L,\sec,LL} = I_{\sec,peak,LL} - \frac{I_{valley,LL}}{N} = 9.9-\frac{1.28}{0.25} = 4.78\,\text{A} \quad (27)$$

$$\Delta I_{L,\sec,HL} = I_{\sec,peak,HL} - \frac{I_{valley,HL}}{N} = 10.4-\frac{0.99}{0.25} = 6.44\,\text{A} \quad (28)$$

Now, it is time to apply the average current formula given by equation (14):

$$I_{\sec,avg,LL} = \frac{\frac{I_{\sec,peak,LL}\Delta t_{LL}}{2}+\frac{\Delta I_{L,\sec,LL}\left(t_{off,LL}-\Delta t_{LL}\right)}{2}+I_{\sec,valley,LL}\left(t_{off,LL}-\Delta t_{LL}\right)}{T_{sw}} \quad (29)$$

$$I_{\sec,avg,LL} = \frac{\frac{9.9\times171n}{2}+\frac{4.78\times(9.3u-171n)}{2}+\frac{1.28}{0.25}\times(9.3u-171n)}{15.4u} = \frac{846n+21.8u+46.7u}{15.4u} = 4.5\,\text{A} \quad (30)$$

$$I_{\sec,avg,HL} = \frac{\frac{I_{\sec,peak,HL}\Delta t_{HL}}{2}+\frac{\Delta I_{L,\sec,HL}\left(t_{off,HL}-\Delta t_{HL}\right)}{2}+I_{\sec,valley,HL}\left(t_{off,HL}-\Delta t_{HL}\right)}{T_{sw}} \quad (31)$$

$$I_{\sec,avg,LL} = \frac{\frac{10.4\times182n}{2}+\frac{6.44\times(12.7u-182n)}{2}+\frac{0.99}{0.25}\times(12.7u-182n)}{15.4u} = \frac{946n+40.3u+49.6u}{15.4u} = 5.9\,\text{A} \quad (32)$$

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To check our calculations, we built a 65-W power supply using the new NCP1250. The adopted schematic appears in Fig. 6.

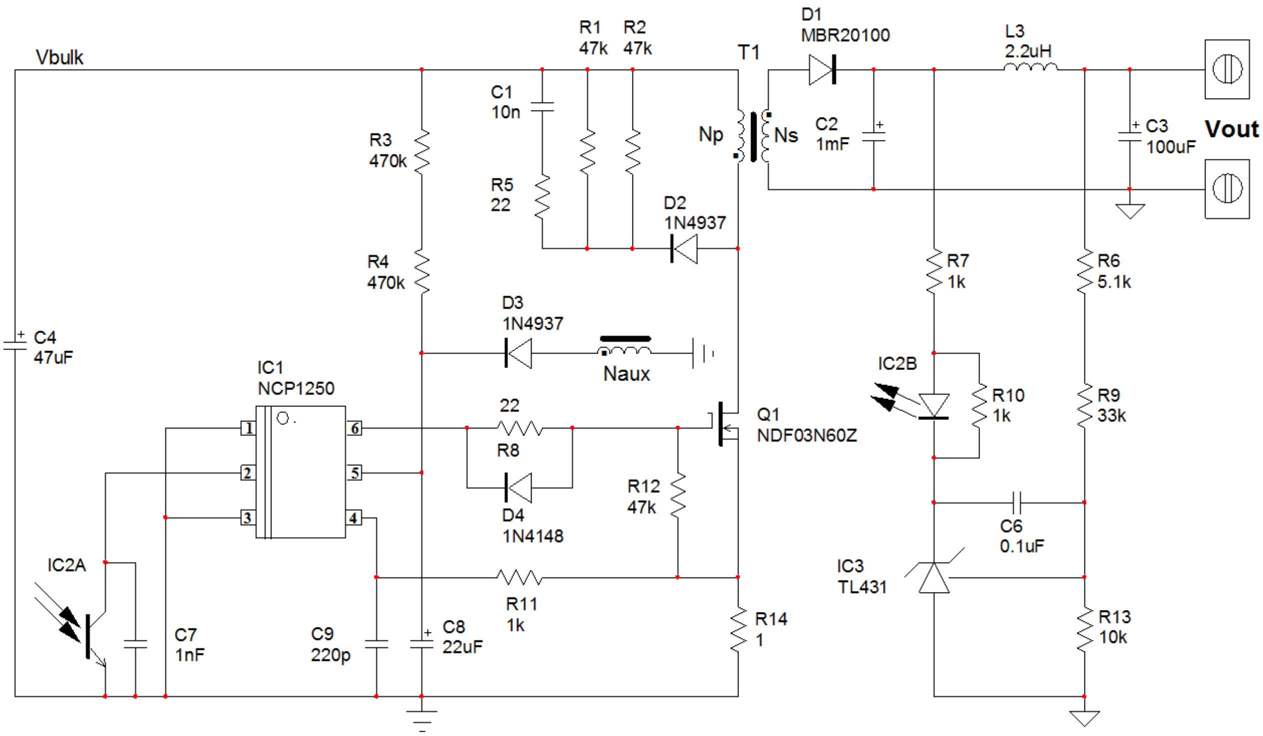

Fig. 6. A prototype adapter has been built using the NCP1250 where the OPP pin (pin 1) is grounded.

Pin 1, which is the Over-Power Protection pin, has purposely been grounded. To check the maximum power that could be delivered before the protection circuit actually trips, we powered the converter at two dc input voltages (120 V and 370 V) while applying a variable load at its output. At low line, we found a maximum dc current of 4.2 A, which rose to 5.9 A at high line.

If we compare these measured values with the results given by equations (30) and (32), a 7% error occurs at low line only. The discrepancy in the numbers can be explained by several variables such as a reduction of the primary inductance as the transformer heats up at low line or by a slight change of the *RCD* clamping network voltage. This voltage could influence the leakage inductor reset time and affect the average output current. Nevertheless, these calculations instruct the designer that a means of limiting the output power has to be implemented. Through this analytical approach, OPP can be incorporated during the initial flyback design stage and then verified when the prototype is assembled.

Conclusion

This final part of the article series on over-power protection explains the influence of the leakage inductor in the power transfer of a flyback converter. It is important to understand the role of the leakage inductor, which delays the rise of the secondary-side current and reduces the available output current. By diverting a portion of the primary-side stored energy, it affects the total efficiency of the converter.

References:

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Christophe Basso is an application engineering director at ON Semiconductor in Toulouse, France. He has originated numerous integrated circuits among which the NCP120X series has set new standards for low standby power converters. SPICE simulation is also one of his favorite subjects and he authored a second book, "Switch-Mode Power Supplies: SPICE Simulations and Practical Designs", published by McGraw-Hill in 2008. His work was positively reviewed in several magazines and in a recent PELS newsletter. Christophe holds a BSEE-equivalent from the Montpellier University, France and a MSEE from the Institut National Polytechnique de Toulouse, France. He holds 17 patents on power conversion and often publishes papers in conferences and trade magazines.

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