

Designing Flyback Transformer for Discontinuous Mode

By Keith Billings, DKB Power Inc., Ontario, Canada

Let's return to the question, "Why Have an Air Gap?" initially posed in the Power Design article in the December 2002 issue of *Power Electronics Technology*. In that article, we found that although an air gap won't prevent saturation in true transformer applications, it does reduce the slope of the B/H loop, reducing permeability and inductance, and increasing the magnetizing current in the primary. We found even a small air gap benefits discontinuous forward designs, as the residual flux value will be nearer zero, allowing a larger working flux density range.

Now, let's examine one more example of the function of an air gap in a ferrite core, while providing a better understanding of flyback transformers. **Fig. 1** shows a typical off-line flyback converter found in a low-power multiple output application. A major advantage of this topology is its low cost and simplicity. In multiple output applications, the addition of a secondary winding, a single diode, and an output capacitor is all that's required for each additional isolated or common output. If one output is closed-loop

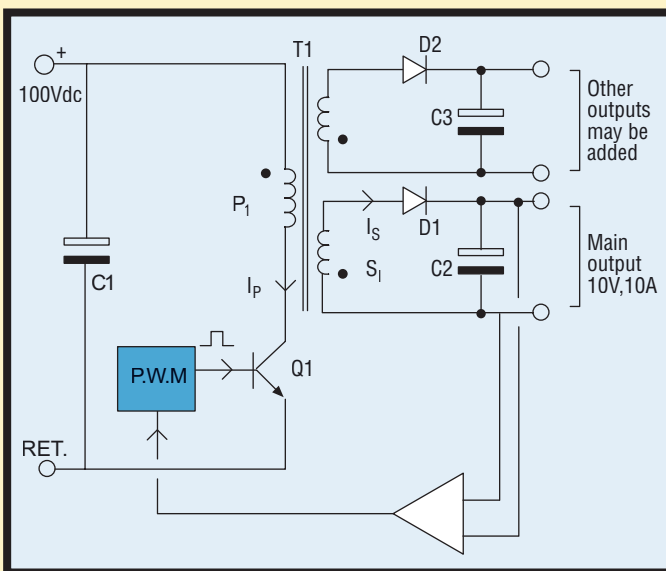


Fig. 1. Typical off-line flyback converter.

voltage stabilized, then all other outputs will be semi-stabilized, (within limits of loading and voltage typically required in many applications). A common difficulty in the design of such converters is the so-called "transformer." We can see how its design requires more care and understanding than its counterpart in the forward converter.

Design difficulties stem from the fact that flyback trans-

formers aren't really transformers. More accurately, they're inductors or chokes, depending on the mode of operation. They have additional windings called the *secondaries*. However, the secondary voltages aren't related to the primary to secondary turn's ratio, as they would normally be with a true transformer. Why is this?

Look again at the phasing of the windings and diodes in **Fig. 1**. When Q1 is ON the input voltage is applied to the start of the primary winding (the dot end). This is positive, and an increasing current is flowing in the primary (the magnetizing current). The phasing of the secondary windings and rectifier diodes is such that the secondaries aren't conducting during this period. Hence, the secondary windings effectively don't exist, or at least the primary can't "see" them. As a result, there's no usable transformer action during the ON period of Q1.

When Q1 turns OFF, the magnetizing current in the primary winding immediately stops. However, the core must now return to its previous condition of near-zero flux, and the voltage on all windings will reverse, creating the flyback action. This brings the secondary windings and diodes into conduction and a decreasing (demagnetizing current) now flows in the secondaries. In other words, the energy stored in the magnetic field during the ON period of Q1, spills out into the secondary during the OFF period of Q1. Once again, during this OFF period, the primary winding is not conducting and effectively does not exist. So, once again, there's no usable transformer action between primary and secondary. However, the secondary windings do "see" each other, because they're all conducting at the same time. As a result, the voltage on the controlled output defines the voltage on all other outputs. In this respect, there's effective transformer action between the secondaries.

With all this in mind, we know not to try to design flyback transformers as if they were transformers, because they don't work that way. What, then, is the best design approach for them? The first step is to decide on the required mode of operation—either discontinuous (complete energy transfer or inductor type operation) or continuous (incomplete energy transfer or choke type operation). To help with this decision, let's look at the waveforms of **Fig. 2**, on page 45.

In the discontinuous mode, the secondary current falls to zero before the next Q1 ON period. The current waveform is sawtooth, as shown in **Figs. 2a** or **2b**, and the peak current is at least twice the mean current. The disadvantage of this mode of operation is the high peak ripple current in

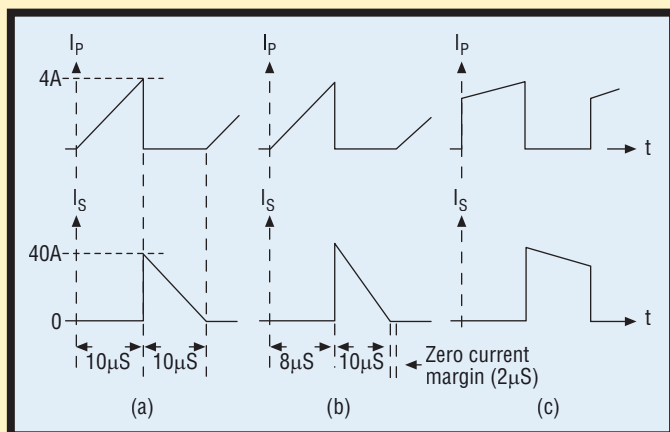


Fig. 2. (a) Discontinuous mode (boundary conditions); (b) Discontinuous mode (working condition with safety margin); (c) Continuous mode (higher power condition).

the transformer windings, diodes, transistors, and caps. Hence, this mode is more suitable for low power applications. Advantages are no dc component in the transformer; no reverse recovery problems in the output rectifier diodes (they're OFF before Q1 turns ON again) less turns; and no right-half-plane-zero in the transfer function ^[1].

In the continuous mode, the secondary current is greater than zero when Q1 turns ON for the next period. You can see

this in **Fig. 2c**. The main advantage is reduced ripple current. Unfortunately, the transformer must now support a dc current component. The output diodes are conducting when Q1 turns ON (presenting diode reverse recovery problems), and it requires more turns while there's a right-half-plane -zero in the transfer function. However, the advantage of much reduced ripple currents makes this mode more suitable for higher power applications.

You can define the mode of operation by adjusting the inductance. In this example, you can do this by adjusting the size of the air gap. Low inductance leads to the discontinuous mode, while large inductance leads to the continuous conduction mode. In general, the inductance may be adjusted by changing the turns, core material (permeability) or the size of the air gap, or a combination of these.

Design Example—Discontinuous Mode

At this stage, the design approach is best illustrated by considering a specific example. Assume a discontinuous mode of operation has been chosen (also known as the complete energy transfer mode) with 50 kHz operating frequency and a power of 100W. The ON period of Q1 is to be 40% of the total period or 8 μ s and the OFF period will be 12 μ s. This provides a 2 μ s margin to ensure the mode remains discontinuous. As shown in **Fig. 2b**, other values can be used.

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This 40% conduction period occurs at the minimum input voltage of 100Vdc and full load. A gapped ferrite core is used, which has a center pole area of say 100 mm².

At 100V input and 100W power, the time averaged primary input current will be 1A. The mean current in Q1 and the primary winding (P1), during the 8 μ s 40% ON period is 2.5A. The peak input current is twice the average or 5A in this example (see this waveform in Fig. 2b).

The required primary inductance can now be calculated using formula $L(di/dt) = V$, where $di = 5A$, $dt = 8 \mu s$ and $V = 100V$. L will be 160 μH (see equations, on page 48).

Now, the minimum primary turns can be calculated. They are defined by the need to support the applied volt seconds ΔB —not by the inductance! The applied volt seconds equate to the parameter ΔB on the B/H loop shown in Fig. 3. This change in magnetic flux density ΔB will be maintained irre-

spective of the size of the air gap—it generates within the primary winding the back mmf required to offset the applied volt seconds.

We'll choose a maximum core flux density of 0.2 Tesla to provide a good working margin on the saturation level of 0.35 Tesla. (Higher flux values will increase core loss but will result in less copper loss because less turns are required, and conversely). The optimum choice is where copper and core loss are equal. This is an iterative process, and cannot be finalized at this stage; it must be checked in the final design. The following formula will yield the minimum primary turns for the chosen parameters:

$$N_{\min} = \frac{V \times t}{B \times Ae}$$

(See equations)

Using the chosen values in the above formula, we find the minimum primary turns are 40.

Similarly, the secondary turns are defined by the need for the secondary

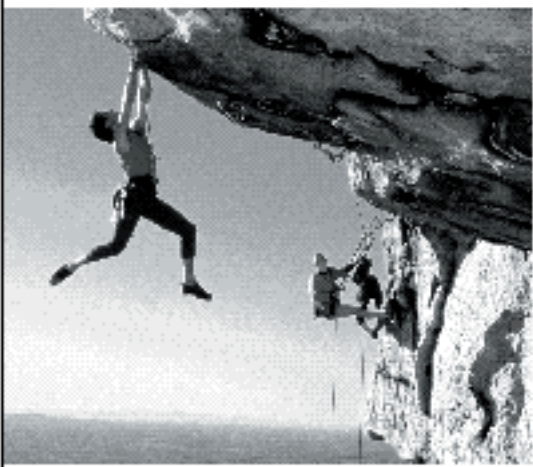
current to fall to zero before the next ON period of Q1. Once again, the turns are not calculated by transformer action, although they are a function of the required secondary voltage (see Figs. 2a and 2b). Fig. 2a shows the boundary condition where the secondary current just falls to zero just before the next ON period. Fig. 2b shows a better condition with more working margin where the current falls to zero in 10 μ s, 2 μ s before the next ON period. This allows for some load and tolerance variations but also increases the peak current slightly.

Assume the secondary voltage is to be 20V. The flyback period is 60% (12 μ s), and the time averaged secondary current for 100W output will be 5A. We have chosen to make secondary conduction period 10 μ s. Hence, the mean current during the secondary conduction period of 10 μ s is 10A, and the peak current is 20A. In this simplified example, we've not made any allowance for diode drop and efficiency loss. Note the high output ripple current, peaking at 20A. Clearly, this isn't a practical design and shows why the discontinuous mode is normally limited to low power applications. However, if the output voltage was to be larger—1000V, for example (you can do this with the flyback topology)—then the peak current would be only 400mA, which is fine.

The secondary inductance required for 20V output and 10 μ s conduction period may now be calculated in the same way as the primary inductance. V_s is now 20V and di/dt is 2A/ μ s. Evaluating, we get 10 μH (see appendix). Since the gap has already been defined by the primary design, we must now select the secondary turns to get this value of inductance in the secondary.

The primary inductance (L_p) is 160 μH for 40 turns, and inductance ratios as N^2 so 10 turns will give 10 μH (see appendix). Less turns may be used, resulting in the current dropping to zero in a shorter flyback period, moving more into the discontinuous mode, increasing the peak secondary current, and providing a wider zero

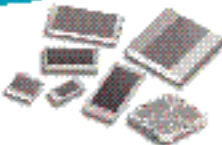
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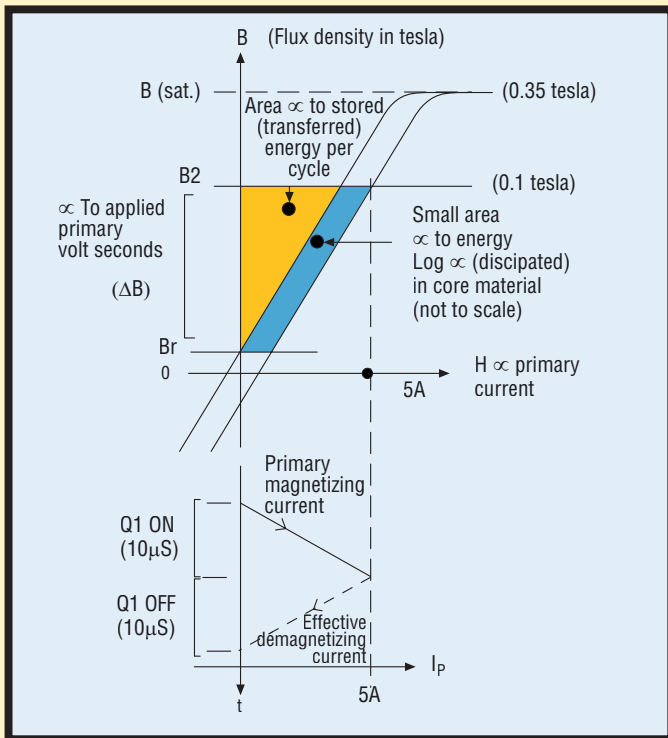


Fig. 3. Top right quadrant of the B/H loop of the gapped core.

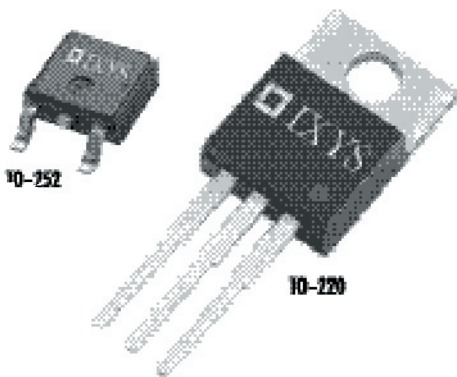
current margin. Conversely, more secondary turns may not allow the secondary current to reach zero during the flyback period, and the continuous mode of operation will be entered. Neither the primary or secondary turns are defined by transformer action. However, if additional secondary windings are to be provided, their voltages will be defined by the transformer ratio between the various outputs, because they all conduct at the same time.

To specify the output voltage, the control loop is closed to the chosen output and adjusts the duty ratio to maintain the output voltage at the chosen value for variations in input voltage and reduced loading. Increasing the loading beyond the designed 100W moves the action into the continuous mode. While this is possible, it complicates the design of both transformer and control loop because it introduces a RHP zero into the transfer function and introduces a dc component into the windings. If the control loop fails, the output voltage goes to a high and uncontrolled value, thus, overvoltage protection is a wise precaution.

To optimize the design, calculate the core and copper loss at full load. They should be near equal for optimum design. If necessary, adjust the chosen flux value and turns to obtain this equality. Remember: Increasing the flux will increase core loss and decrease copper loss, since there will be less turns.

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We can now wind the transformer. The primary is wound using a wire gauge to fill less than 50% of the bobbin area to allow for insulation. The rest of the space is used for the secondaries. Several smaller diameter strands may be used in parallel to reduce skin effects.

At the moment of turn OFF of Q1, the primary current (more correctly, ampere turns product) must commute to the secondary. Any leakage inductance opposes this action, and a large voltage overshoot will be generated on the primary needing aggressive snubbing action. This problem is reduced by interleaving primary and secondary as you would in a forward transformer design.

When finished, the core is fitted using an elastic band to hold it together, and the core gap adjusted to obtain the required primary inductance of 160 μ H. For a better understanding, consider the action of the core gap in

a somewhat different way. Energy is stored in the magnetic field during the ON period of Q1 and transferred to the secondary during the OFF period of Q1. Fig. 3 shows the top right quadrant of the B/H loop of the gapped core, with the working value of flux density swing ΔB imposed on it. The area to the left of the B/H loop is proportional to the stored energy per cycle, while the much smaller area contained within the B/H loop is an indication of core loss (not to scale). In general, power = $f \times V_e \int H dB$

Equations

Primary Inductance

$$L_p = \frac{V_p \times \Delta t}{\Delta I}$$

Where:

L_p = Primary inductance (μ H)

V_p = Primary voltage (V)

Δt = Q1 ON time (μ s)

ΔI = Peak primary current (A)

Primary turns

$$N_p = \frac{V_p \times t}{B \times A_e}$$

Where:

t = Q1 ON time (μ s)

B = Peak flux density (tesla)

A_e = effective area of center pole (mm^2)

N_p = Minimum primary turns

Inductance factor

$$A_L = \frac{L_p}{N_p^2}$$

Where:

A_L = Inductance of a single turn (μ H)

Secondary inductance $L_s = N_s^2 \times A_L$

Where:

L_s = Secondary inductance (μ H)

N_s = Secondary turns

Secondary turns

$$N_s = \sqrt{\frac{L_s}{A_L}}$$

Primary stored/transferred energy per cycle $j = \frac{1}{2} L_p I_p^2$ (joules)

Where:

I_p = Peak primary current (A)

L_p = primary inductance (H)

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Transferred power $P = f \times (\frac{1}{2} L_p I_p^2) = (W)$

Where:

f = Frequency (Hz)

$$\text{Air Gap} = \frac{\mu r \times N_p^2 \times A_e}{L_p}$$

Where:

Air gap = (mm)

$\mu r = 4 \pi \times 10^{-7}$

N_p = Primary turns

A_e = Pole area (mm²)

L_p = Primary inductance (mH)

Increasing the Air Gap

Increasing the air gap will swing the B/H loop more to the right, increasing the area and hence the stored energy. Most of the energy is stored in the air gap because this is the lowest permeability part of the magnetic path, getting something-for-nothing.

The remnant flux (Br) also reduces as the core gap increases, slightly increasing the usable flux working range. The peak primary current increases as the inductance reduces. Since stored

energy is proportional to $L_p \times I_p^2$, energy still increases with the larger gap—even though inductance falls.

Below the B/H loop is shown the primary magnetizing current during the ON time of Q1. In pecked form, it shows the effective secondary demagnetizing current as would be reflected to the primary winding in terms of ampere-turns. Of course, it does not actually flow in the primary. The ac flux swing ΔB remains constant irrespective of the core gap: It is a function of applied primary volt seconds. The peak current and stored energy are the dependant variables set by the size of the air gap. So now we see one more application for the air gap.

For multiple outputs, simply sum the total output power and use this value in the equations when calculating the primary and secondary inductance of the main controlled output. The secondary currents will sort themselves out, depending on the loading applied

to each output.

Even while ignoring the transformer action, (because it's not a design parameter), it still exists. The flyback voltage is reflected back to the primary winding during the OFF period of Q1. Hence, reducing secondary turns increases secondary peak currents and increases the voltage stress on Q1 during the OFF period.

Look for the follow-up article in the next issue of *PETech*. It will cover the transformer design for the continuous mode (or incomplete energy transfer choke mode). **PETech**

Keith Billings is president of DKB Power Inc., dkbp@rogers.com.

Reference

1. Keith Billings, "Switchmode Power Supply Handbook," McGraw-Hill ISBN 0-07-006719-8.

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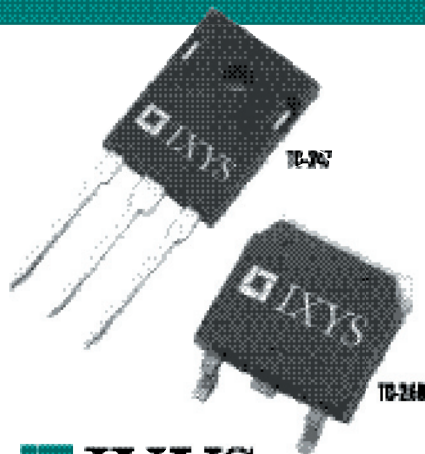
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IXL 40470K	1700V	20A	1.2V	10	30A	100	TO-247, 170-300
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