Implementing Flyback Transformer Design for Continuous Mode

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

In last month's Power Design column, we examined the functional principles of continuous mode (or incomplete energy transfer mode) of a flyback transformer. In this issue, let's apply the same design principles learned earlier in a transformer design example and look at the function of an air gap in a ferrite core.

Transformer Design Example

Fig. 1 shows a typical flyback converter as used for single output application. **Fig. 2** shows typical waveforms found in continuous mode operation. In this mode, energy is stored in the magnetic field of the core and air gap during the ON period of Q1. However, not all the stored energy is transferred to the secondary when Q1 turns OFF We have a variable (top) part of the waveform, and an effective dc component, (as there would be in an output smoothing choke in a forward converter). **Fig. 3** shows the B/H characteristic.

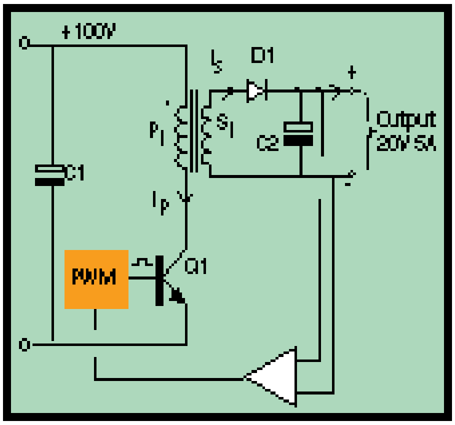

Fig. 1. *Typical flyback converter.*

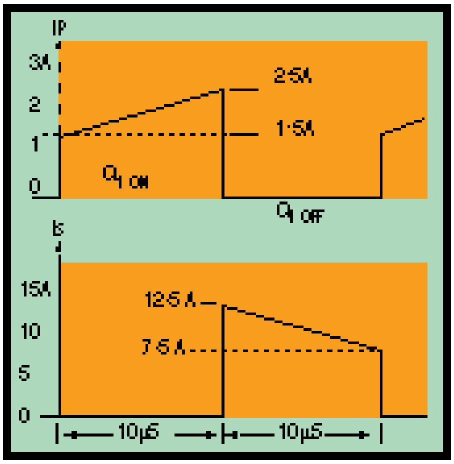

Fig. 2. *Typical waveforms found in continuous mode operation.*

For this transformer design example, we'll assume an output power requirement of 100W. The input voltage is 100V. The secondary voltage is 20V at a mean current of 5A. The frequency is to be 50 kHz, with a 50% ON period for Q1 (10µs). A gapped ferrite core with a pole area of 100mm² is to be used. The example assumes 100% efficiency and zero diode drop.

The first step is to decide the load current range over which the continuous mode of operation is to be maintained. The larger the load range, the more turns and inductance are required on the primary and the greater will be the copper loss. We'll assume a 75% load range providing 1.25A to 5A output. This choice results in a ripple current of half the mean value at maximum load. The waveforms in **Figs. 2** and **3** show this condition. We must choose the maximum flux density swing ΔB. The smaller the flux swing the lower the core loss, but the turns will be greater increasing the copper loss. The optimum choice is where total copper loss equals the core loss. This is an iterative process and can't be fully established at this stage. It must be checked in the final design, making required adjustments. For this example, let's assume a ΔB of 0.16 Tesla (about ½ of the saturating flux density). This provides a 50% flux margin for the dc polarizing force shown as *Bdc* on the vertical B scale in **Fig. 3**.

We can now calculate the minimum primary turns Np; (See **Equations.**)

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

Vp = 100V, t (the ON period of Q1) is 10µs, B max is 0.166 Tesla and Ae (the area of the core) is 100 mm²

Hence the primary turns (Np) will be 60 turns.

The primary inductance can be calculated as follows; Assuming 100% efficiency at a power of 100W and an input voltage of 100V the time averaged input current will be 1A .Hence, the mean current during the 50% "on" period of Q1 must be 2A. We have chosen to make the ripple current 50% (or 1A in this example). From inspection of the top part of **Fig. 2**, the primary current change is from 1.5A to 2.5A, a ΔI of 1A. The ON period is 10µs, and we can now calculate the primary inductance required to produce this result.

The primary inductance

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

This will be 1000 µH. (See **Equations.**)

While the OFF (flyback) period of Q1 is also 10µs, the secondary volts per turn during the flyback period will be the same as the primary volts per turn during the ON period of Q1. Thus, for 20V output, 12 turns will be required. The 50% duty ratio is a special case. If the duty ratio is not 50%, then this transformer like ratio won't apply, and you must obtain the secondary turns from the calculated secondary inductance, in the same way as it was in the discontinuous mode example in the April article. Also, we've provided no adjustment for rectifier diode drop.

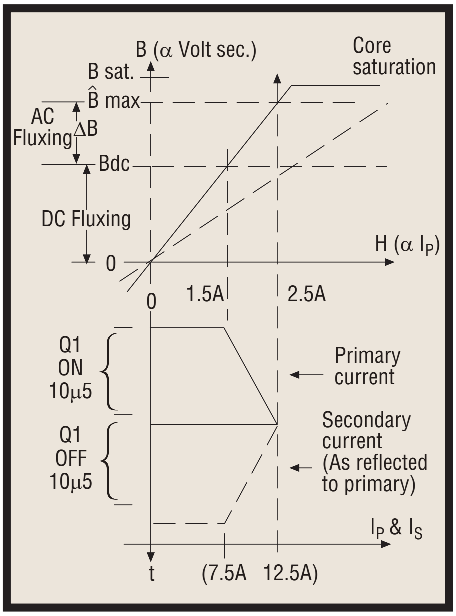

Fig. 3. *An ideal B/H characteristic.*

The secondary current ratio follows the turns ratio, since the primary ampere turns product must be maintained. Thus, Is = Ip × 60/12. As shown in **Figs. 2** and **3**, the secondary waveform starts at 12.5A and falls to 7.5A at the end of the OFF period. A mean value of 10A for the half period giving 5A average at 20V and power equality is maintained.

The "transformer" is now wound using a wire size for the 60 turns primary that will use less than 50% of the bobbin area to allow for insulation. The remaining space is used for the secondaries. To minimize skin effect, several parallel strands of a smaller wire size would normally be used. Although there's no direct transformer action (because the primary and secondary aren't conducting at the same time), the leakage inductance should still be minimized. As Q1 turns OFF, the primary current must be commutated from the primary winding to the secondary winding. Any leakage inductance opposes this commutation and will produce a voltage overshoot on the primary when Q1 turns OFF. Hence, the windings would normally be interleaved—½ primary, secondary and then ½ primary. (See reference.) At this stage, you normally fit the core, and adjust the gap to get the required primary inductance. However, it can also be calculated. (See **Equations**.)

The main advantage of the continuous mode is reduced ripple current. The disadvantages are that the "transformer" must support a dc current component, the output diodes are conducting when Q1 turns ON (presenting diode reverse recovery problems), more turns are required, and there's a "right half plane zero" in the transfer function leading to poor transient response. However, the reduced ripple current, makes this mode more suitable for higher power applications.

The mode of operation was defined by adjusting the inductance, (by using a smaller air gap). Low inductance (large air gap) leads to the discontinuous mode, while large inductance leads to the continuous mode. The inductance may be adjusted by changing the turns, core material, or the size of the air gap, or a combination of these.

For multiple outputs, 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 adjust depending on the loading applied to each output.

Although we've ignored the transformer action, as 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 the secondary turns increases secondary peak currents and increases voltage stress on Q1.

The following equations are dimensionally modified to yield convenient answers.

Equations

Primary Turns

$$N_P = \frac{V_P \times t}{\Delta B \times Ae}$$

Where:
N_P = Primary Turns
V_P = Primary Voltage
t = Q1 ON time (µs)
ΔB = Peak flux density (tesla)
Ae = effective area of center pole (mm^2)

Primary Inductance

$$L_P = \frac{V_P \times \Delta t}{\Delta I_P}$$

Where:
L_P =Primary inductance (µH)
Δt = Q1 ON time (µs)
ΔI_P = Current change I_1 to I_2 (A)

Inductance Factor

$$A_L = \frac{L_P}{N_P^2}$$

Where:
A_L = Inductance of a single turn (µH)

Secondary Inductance $L_S = N_S^2 \times A_L$

Where:
L_S = Secondary inductance (µH)
N_S = Secondary turns

Secondary Turns

$$N_S = \sqrt{\frac{L_S}{A_L}}$$

Air Gap Length

$$\text{Air Gap} = \frac{\mu o \times N_P^2 \times Ae}{L_P}$$

Where:
Air gap = (mm)
µo = 4 π × 10^{-7}
N_P = Primary turns
Ae = Area of core(mm^2)
L_P = Primary inductance (mH)
α = Air Gap (mm)

ac Flux Density

$$B_{ac} = \frac{V_P \times t}{N_P \times Ae}$$

dc Flux Density

$$B_{dc} = \frac{\mu o \times N_P \times I_{dc}}{\alpha \times 10^{-3}}$$

where α = Air Gap (mm)

In the next Power Design column, we'll cover a transformer design for a forward converter, including alternative core materials. **PETech**

References

1. *Switchmode Power Supply Handbook. Keith Billings, Published by McGraw Hill ISBN 0-07-006719-8.*
2. *Switching Power Supply Design. Abe Pressman, Published by McGraw Hill ISBN 0-07-052236-7.*

Keith Billings is president of DKB Power Inc. (dkbp@rogers.com) and author of the Switchmode Power Supply Handbook published by McGraw Hill ISBN 0-07-006719-8. He also presents the late Abe Pressman's three-day "Modern Switching Power Design Course." For more information visit www.apressman.com.