Coupled Inductor Design For A Flyback With Very Wide Input Voltage Range

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Flyback converters along with boost and SEPIC converters are based on the principle of storing energy in a magnetic field for one portion of the switching period in the magnetic core and then releasing during the remainder of the switching period. This is why such circuits use inductors for storing magnetic energy and not transformers.

The output voltage of such converters depends on the duty cycle of the pulse train that is used in its operation. Since the energy transfer has a delay, the transfer function of such converters has a right half-plane zero. This requires a special attention when designing the control system of these converters since these systems are much less stable.

Although flyback converters appear less sophisticated than other types of converters, there are specific difficulties in their design. That's because the inductor secondary winding transfers energy to the primary side of the inductor during the stage of the stored energy release, creating a step-up transformer with a very high turns ratio. Parameters of the snubber on the primary side are reflected to the secondary side with the square of the primary-to-secondary turns ratio. This causes a significant increase in the secondary voltage decay time and secondary current setup time. So, despite their popularity, flyback designs have their complications.

Flyback designs require more attention as their specifications become more challenging. For instance, there is an automotive and railroad class of flyback converters that demands operability over a very wide input voltage range—from 30 V to 800 V. These are backup power supplies that should operate when the traction battery decays. With such a wide input range, duty cycle will vary widely. Yet, many existing flyback converters' design processes do not assume the whole operable range of the duty cycle that is a reflection of the input voltage range. As an example, the design in reference [1] does not consider the maximum input voltage, which is where duty cycle goes low, perhaps even to pulse widths that are too short for the controller to generate.

Rather than assuming a duty cycle range, most flyback designs assume some average value for duty cycle. This can have serious implications for the performance of the coupled inductor, particularly in a flyback design with a very wide input voltage range like the one being addressed here. This article offers a procedure for designing the coupled inductor that takes into account the full duty-cycle range of the flyback. The procedure is demonstrated in a design example using the 30-V to 800-V input range.

The coupled inductor design presented here mainly concerns its magnetic design. Other inductor parameters like wire type, number of strands, placement in the core window, operating current RMS and peak values and wire thickness calculations are widely represented in many online and paperback sources. This approach is taken because the magnetic core design is the most important part of the whole inductor design since, if the core doesn't work properly, the rest is useless.

Defining Key Terms

To begin, let's denote some variables that will be needed to derive the equations for core selection as a function of a duty cycle range.

$\Delta\Phi_p$ = magnetizing flux of the primary winding

$\Delta\Phi_S$ = demagnetizing flux of the primary winding

V_{prim} = voltage applied to the primary winding when the main transistor is on

V_{sec} = voltage generated by the secondary winding

V_{AUX} = voltage generated by the auxiliary winding. This voltage is used in the regulation loop.

N_P = number of turns of the primary winding

N_S = number of turns of the secondary winding

N_{AUX} = number of turns of the auxiliary winding

t_{on} = turn-on duration of the main transistor

t_{off} = turn-off duration of the main transistor

α = input voltage range

β = duty-cycle range

T_C = period of repetitive pulses

Symbolic Design

From the Flux Conservation law we can figure out that the magnetizing flux $\Delta\Phi_P$ should be equal to the demagnetizing flux $\Delta\Phi_S$ since the lump flux for the period should be equal to 0.

$$\Delta\Phi_P = \Delta\Phi_S \quad (1)$$

For the primary winding, the following relationship holds true:

$$V_{prim} = \frac{\Delta\Phi_P}{t_{on}} \cdot N_P \quad (2)$$

Similarly, for the secondary winding, the following holds true:

$$V_{sec} = \frac{\Delta\Phi_S}{t_{off}} \cdot N_S \quad (3)$$

which yields:

$$\frac{V_{prim} \cdot t_{on}}{N_P} = \frac{V_{sec} \cdot (T_C - t_{on})}{N_S} \quad (4)$$

Denoting

$$N_{PS} = \frac{N_P}{N_S} \quad (5)$$

we get from (4):

$$V_{prim} \cdot t_{on} = V_{sec} \cdot (T_C - t_{on}) \cdot N_{PS} \quad (6)$$

Denoting duty-cycle as D = t_{on}/t_{off}, we get

$$\frac{V_{prim} \cdot D}{N_P} = \frac{V_{sec} \cdot (D-1)}{N_S} \quad (7)$$

and

$$N_{PS} = \frac{D \cdot V_{prim}}{V_{sec} \cdot (D-1)} \quad (8)$$

The coupled inductor under design has a wide primary winding voltage range from V_{prim_min} to V_{prim_max} while the output voltage V_{sec} remains stable due to operation of the control feedback loop.

Solving (8) with respect to D we get

$$D = \frac{N_{PS} \cdot V_{sec}}{V_{prim} + N_{PS} \cdot V_{sec}}$$

Therefore, since at the maximum input voltage V_{prim_max}, the duty-cycle is minimal (D_{min}) and at the minimal input voltage V_{prim_min} the duty-cycle has a maximum value of D_{max}, we can write:

$$\left.\begin{aligned} N_{PS} &= \frac{D_{min} \cdot (V_{prim\_max})}{V_{sec} \cdot (D_{min} - 1)} \\ V_{prim\_max} \cdot D_{min} &= V_{sec} \cdot (1 - D_{min}) \cdot N_{PS} \\ V_{prim\_min} \cdot D_{max} &= V_{sec} \cdot (1 - D_{max}) \cdot N_{PS} \end{aligned}\right\} \tag{9}$$

Next, designate:

$$V_{prim\_max} = \alpha \cdot V_{prim\_min} \tag{10}$$

and

$$D_{max} = \beta \cdot D_{min} \tag{11}$$

Then, divide the first equation of (9) over the second one:

$$\frac{V_{prim\_max} \cdot D_{min}}{V_{prim\_min} \cdot D_{max}} = \frac{V_{sec} \cdot (1 - D_{min}) \cdot N_{PS}}{V_{sec} \cdot (1 - D_{max}) \cdot N_{PS}} \tag{12}$$

And plugging in (10) and (11) into (12) we get:

$$\frac{\alpha \cdot V_{prim\_min} \cdot D_{min}}{V_{prim\_min} \cdot \beta \cdot D_{min}} = \frac{V_{sec} \cdot (1 - D_{min}) \cdot N_{PS}}{V_{sec} \cdot (1 - \beta \cdot D_{min}) \cdot N_{PS}}$$

Simplifying, we obtain:

$$\frac{\alpha}{\beta} = \frac{1 - D_{min}}{1 - \beta \cdot D_{min}} \tag{13}$$

A reminder: α is a given value. It is the input (supply) voltage ratio:

$$\alpha = \frac{V_{prim\_max}}{V_{prim\_min}}$$

So, from (13) above, we can find an expression for β, which defines the duty-cycle range:

$$\beta = \frac{\alpha}{D_{min} \cdot \alpha - D_{min} + 1} \tag{14}$$

Now, based on equation (18) for core volume given in reference [2], we can define the minimum core volume Vol that would work at the specified power as

$$Vol = \frac{2 \cdot P_{in} \cdot D_{max} \cdot \mu_0 \cdot \mu_r}{{B_m}^2 \cdot 2f_{SW}} \tag{15}$$

With a value for core volume in hand, we are now able to select a magnetic core (as will be demonstrated further in the design example below.)

Then, from equation (9), the primary-to-secondary winding number of turns is:

$$N_{PS} = \left[-\frac{D_{min} \cdot V_{prim\_max}}{V_{sec} \cdot (D_{min} - 1)}\right]$$

Finally, to complete the magnetic design, we need to determine the primary, secondary and auxiliary winding inductances and numbers of turns:

$$L_p = \frac{{B_m}^2 \cdot Vol}{{I_{max}}^2 \cdot \mu_0 \cdot \mu_r} \tag{16}$$

(The derivation for this equation is given in the appendix.)

$$N_P = \sqrt{\frac{L_P}{A_L}} \tag{17}$$

$$L_S = \frac{L_P}{(N_{PS})^2} \tag{18}$$

$$N_S = \frac{N_P}{N_{PS}} \tag{19}$$

$$L_{AUX} = \frac{L_P}{(N_{PAUX})^2} \tag{20}$$

$$N_{AUX} = \frac{N_P}{N_{PAUX}} \tag{21}$$

A Physical Design Example

To illustrate the application of the procedure described above, we begin by assigning the following numbers to the given parameters:

T_C = 10 µs

V_{prim_max} = 800 V

V_{prim_min} = 30 V

V_{sec} = 20 V

Efficiency $\eta = P_{out}/P_{in}$, assume η = 0.8

P_{out} = 60 W

μ_r = 90

B_m = 0.4 T

f_{SW} = 100 kHz

$AL = 100 \times 10^{-9}$ H

First, we will calculate input voltage range:

$$\alpha = \frac{V_{prim\_max}}{V_{prim\_min}} = 26.667$$

Based on the author's previous experience, we will set

$$D_{min} = 0.15$$

Therefore

$$\beta = \frac{\alpha}{D_{min} \cdot \alpha - D_{min} + 1} = 5.498$$

$$D_{max} = \beta \cdot D_{min} = 0.825$$

$$N_{PS} = \left[-\frac{D_{max} \cdot V_{prim\_min}}{V_{sec} \cdot (D_{max} - 1)} \right] = 7.059$$

To double-check that result:

$$N_{PS} = \left[-\frac{D_{min} \cdot V_{prim\_min}}{V_{sec} \cdot (D_{min} - 1)} \right] = 7.059$$

In both cases—at the minimum primary voltage and at its maximum—the primary-to-secondary turns ratio is the same. This confirms the correctness of the method.

Having a value for N_{ps} allows us to calculate maximum primary current:

$$I_{max} = \frac{P_{out}}{\eta \cdot V_{sec} \cdot D_{min} \cdot N_{PS}} = 3.542 \text{ A}$$

which we'll need shortly to calculate primary inductance.

But first, with a substitution for P_{in}, we can apply equation 15 from above to determine the minimal permissible magnetic core volume:

$$Vol = \frac{2 \cdot \frac{P_{out}}{\eta} \cdot D_{max} \cdot \mu_0 \cdot \mu_r}{{B_m}^2 \cdot 2f_{SW}} = 437 \times 10^{-6} L$$

This value is the minimum size dictated by core saturation as a magnetic core having smaller volume wouldn't be able to produce enough power to the load.

Let's replace with a standard model 2510-E of Magnetics[3] having volume

$Vol = 1870$ mm^3

This value will not only ensure that the core does not saturate, but also will allow enough of a winding window so wires fit well with sufficient clearance and creepage. Keep in mind that the maximum voltage seen on the primary will be 800 V operating voltage plus 200 V reflected.

Next, we'll use that value to calculate primary inductance:

$$L_p = \frac{{B_m}^2 \cdot Vol}{{I_{max}}^2 \cdot \mu_0 \cdot \mu_r} = 210.908 \text{ x } 10^{-6} H$$

from which we will assume a standard value of

Lp = 220 µH

This yields the numbers of turns for the primary and secondary windings:

$$N_p = \text{ceil}\left(\sqrt{\frac{L_p}{A_L}}\right) = 47$$

and

$$N_S = \text{ceil}\left(\frac{N_P}{N_{PS}}\right) = 7$$

And for the auxiliary winding at V_{AUX} = 16 V, the turns ratio primary-to-auxiliary windings is

$$N_{PAUX} = -\frac{D_{max} \cdot V_{prim\_min}}{V_{AUX} \cdot (D_{max} - 1)} = 8.824$$

$$N_{PS} = -\frac{D_{max} \cdot V_{prim\_min}}{V_{SEC} \cdot (D_{max} - 1)} = 7.059$$

Therefore, the number of turns for the auxiliary winding is

$$N_{AUX} = \text{ceil}\left(\frac{N_P}{N_{PAUX}}\right) = 6$$

Inductance of the auxiliary winding:

$$L_{AUX} = \frac{L_p}{(N_{AUX})^2} = 2.826\, x\, 10^{-6} H$$

Recalling the calculation of N_{PS}:

$$N_{PS} = -\frac{D_{max} \cdot V_{prim\_min}}{V_{SEC} \cdot (D_{max} - 1)} = 7.059$$

We can now calculate the inductance of the secondary winding:

$$L_S = \frac{L_p}{(N_{PS})^2} = 4.415\, x\, 10^{-6} H$$

Finally, we can roundup these inductances to integer values:

L_{AUX} = 4 µH

L_S = 5 µH

Conclusion

By designing a coupled inductor that takes into account duty cycle range rather than average value, we can eliminate a cumbersome adjustment of the inductor's number of turns and avoid usage of oversized windings.

APPENDIX

Calculate Inductance Of Any Inductor For Maximum Current And Core Datasheet Parameters

If an inductor is connected to a current source whose current changes from the minimum to the maximum value, I_{max}, , the magnetic flux density B_m in the inductor magnetic core depends on the current I_{max} in the inductor winding and its turns number N_1 this way:

$$B_m = \mu_0 \cdot \mu_r \cdot \frac{I_{max}}{l_m} \cdot N_1 \qquad (1)$$

It is conventional to define inductor winding inductance L_1 by means of inductance factor A_L as:

$$L_1 = A_L \cdot {N_1}^2 \qquad (2)$$

where

$$A_L = \frac{\mu_0 \cdot \mu_r \cdot S_m}{l_m} \qquad (3)$$

And S_m and l_m are the magnetic core cross-sectional area and magnetic path length, respectively. These are the core datasheet parameters.

From (2) we can easily obtain using (3)

$$N_1 = \sqrt{\frac{L_1 \cdot l_m}{\mu_0 \cdot \mu_r \cdot S_m}} \qquad (4)$$

Thus, from (1), plugging in (4) for N_1, we get:

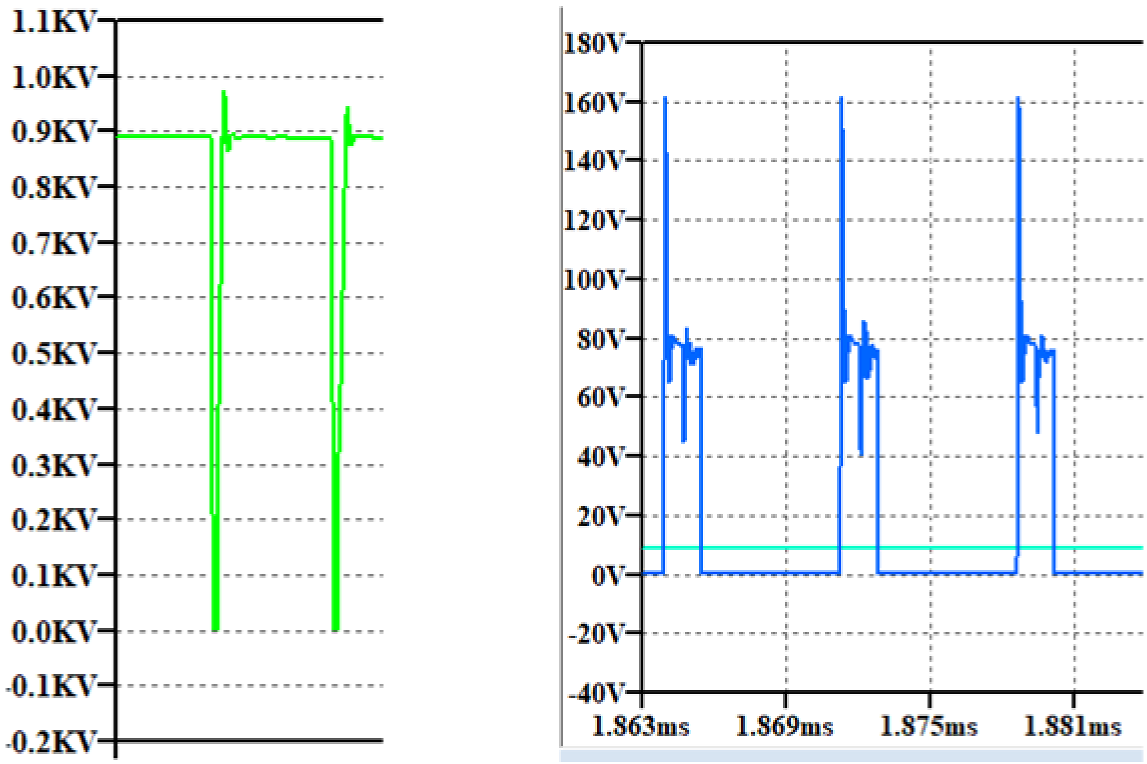
$$B_m = \mu_0 \cdot \mu_r \cdot \frac{I_{max}}{l_m} \cdot \sqrt{\frac{L_1 \cdot l_m}{\mu_0 \cdot \mu_r \cdot S_m}}$$

that gives, after simplification,

$$B_m = I_{max} \cdot \sqrt{\frac{L_1 \cdot \mu_0 \cdot \mu_r}{S_m \cdot l_m}} \qquad (5)$$

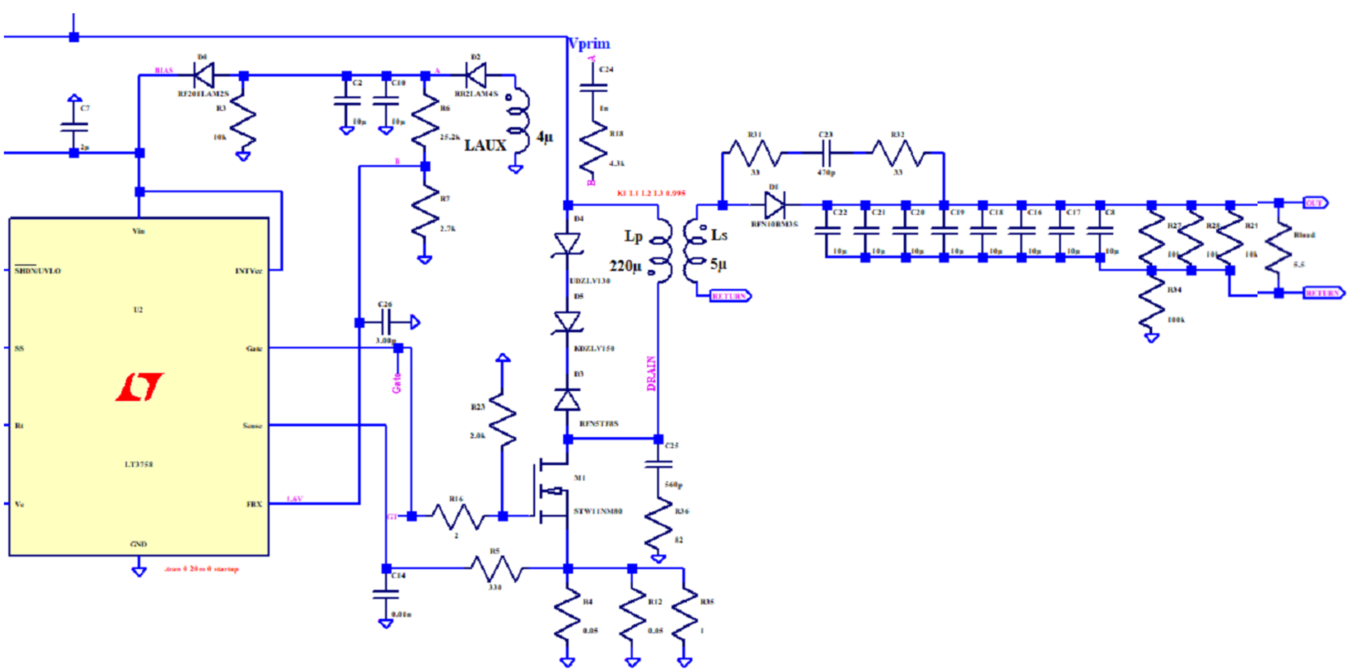
From (5) we obtain an expression for L_1:

$$L_1 = \frac{{B_m}^2 \cdot S_m \cdot l_m}{{I_{max}}^2 \cdot \mu_0 \cdot \mu_r} \qquad (6)$$



(a) (b)

Appendix Fig. 1. V_{DRAIN} with respect to ground at 800-V supply (Vprim_max) and minimum duty cycle (Dmin = 0.15) (a). V_{DRAIN} with respect to ground at 30-V supply (Vprim_min) and maximum duty cycle (Dmax = 0.825) (b). Keep in mind that the duty cycle here is the ratio of the notch-to-period duration.



Appendix Fig. 2. Partial schematic of a flyback converter.

References

1. "Automotive 40-V to 1-kV Input Flyback Reference Design Supporting Regenerative Braking Test," Texas Instruments Design Guide: TIDA-01505, August 2019.
2. "Designing Energy Storing Inductors Properly" by Gregory Mirsky, How2Power Today, January 2019.
3. Magnetics' Advanced Part Number Finder, search results for 2510-E cores.

About The Author

Gregory Mirsky is a design engineer working in Deer Park, Ill. He currently performs design verification on various projects, designs and implements new methods of electronic circuit analysis, and runs workshops on MathCAD 15 usage for circuit design and verification. He obtained a Ph.D. degree in physics and mathematics from the Moscow State Pedagogical University, Russia. During his graduate work, Gregory designed hardware for the high-resolution spectrometer for research of highly compensated semiconductors and high-temperature superconductors. He also holds an MS degree from the Baltic State Technical University, St. Petersburg, Russia where he majored in missile and aircraft electronic systems.

Gregory holds numerous patents and publications in technical and scientific magazines in Great Britain, Russia and the United States. Outside of work, Gregory's hobby is traveling, which is associated with his wife's business as a tour operator, and he publishes movies and pictures about his travels online.

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