Design Feature

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Spreadsheet simplifies switch-mode power-supply flyback-transformer design

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Designing flyback transformers for switch-mode power supplies involves many calculations. A spreadsheet can make short work of this otherwise-tedious task.

When you develop flyback power supplies, especially continuous-mode supplies, transformer design is usually the biggest stumbling block. **Reference 1** uses an Excel spreadsheet as the centerpiece of simple flyback-transformer design technique. With the spreadsheet, continuous or dis-continuous-mode flyback-transformer design takes less than 10 minutes. Compare this operation with the tedious, time-consuming task of manually number-crunching more than 50 independent parameters and dependent variables.

An improved spreadsheet incorporates many new features, including a means of accurately estimating magnet-wire gauge and current capacity. An energy equation, which uses loss allocation, now determines the continuous-mode primary inductance. The improved spreadsheet also calculates output-capacitor ripple current and estimates MOSFET switch-drain peak-stress voltage. To simplify design and iteration, the peak switch current and duty cycle are now dependent variables instead of independent parameters, as in **Reference 1**. For multiple-output supplies, the new spreadsheet calculates auxiliary-diode peak-inverse voltage and transformer auxiliary-winding turns. The remaining calculations remain unchanged.

Contact Power Integrations Inc for copies of **Reference 3**, which contains the following application notes: AN-14: "TOPSwitch tips, techniques, and troubleshooting guide," AN-16: "TOPSwitch flyback design methodology," AN-17: "Flyback transformer design for TOPSwitch power supplies," and AN-18: "TOPSwitch flyback transformer construction guide." You can also obtain copies of the Excel spreadsheet from Power Integrations or download it.

Figure 1 shows a typical flyback power supply using the TOP204 TOPSwitch. This device combines an integrated, high-voltage MOSFET switch with a complete switching-power-supply controller and protection circuitry in a single three-pin TO220 package. The power supply operates from 85 to 265V ac and delivers 30W at 15V output.

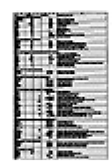
Figure 2 shows typical voltage and current waveforms taken with the flyback power supply delivering 30W from a 115V-ac input. The primary current, I_{PRI}, increases linearly with a rate of change (di/dt) that varies directly with the dc input voltage and inversely with the primary inductance. The ripple current, I_R, is defined as the incremental linear current rise (di) over the switch's entire on-time, t_{ON}. The peak primary current, I_P, is the final value that occurs just before the switch turns off. The transformer core's magnetic field stores energy proportional to I_P^2, just as if the primary winding were a simple inductor. The secondary winding carries a reflected voltage that is related to the primary voltage by the transformer turns ratio and has the same "dot" polarity.

While the switch is on, the output diode, D_2, and bias diode, D_3, are reverse-biased, preventing secondary-current flow. When the switch turns off, the collapsing magnetic field induces an abrupt voltage reversal on all transformer windings, so the dot side is now higher in potential than the no-dot side. Diodes D_2 and D_3 become forward-biased, and the secondary current rises quickly to a peak value of I_P/n, where n is the transformer turns ratio. I_{PRI} immediately drops to zero. The switch-drain voltage quickly rises to the sum of the dc input voltage and the reflected output voltage. The secondary-winding current, I_{SEC}, now linearly decreases at a rate that varies directly with the output voltage and inversely with secondary inductance. The duty cycle is defined as the ratio of t_{ON} to the switching period, T.

Figure 2 shows the drain and diode voltages as well as the trapezoidal current waveforms characteristic of continuous-mode operation. Secondary current is still flowing when the switch turns on at the beginning of the next cycle. The stored energy is not completely delivered to the load. Because of the nonzero magnetic field, energy remains in the core when the switch turns on again. This energy causes the initial step in the switch current.

The ripple-to-peak-current ratio, K_{RP}, determines how continuously the transformer operates. Transformers designed for discontinuous operation have a K_{RP} of 1. Practical continuous designs have lower peak currents and a K_{RP} that is less than 1 but usually greater than 0.4. K_{RP} is inversely proportional to the primary inductance, so a continuous design with lower K_{RP} has a higher inductance.

Flyback-transformer design now begins with the specification of the spreadsheet's three groups of independent variables (**Figure 3**).



Application variables

The output power, P_O; output voltage, V_O; ac-mains frequency, f_L; switching frequency, f_S (fixed at 100 kHz); and the minimum and maximum ac-mains voltage, (V_{ACMIN} and V_{ACMAX}), come directly from the application specification. For efficiency, *leta*, start with an estimate based on measurements in similar power supplies, or use a value of 0.8 if data is unavailable.

You can use efficiency to calculate the power supply's total power loss, P_L. Some power losses that occur in series primary components, such as the bridge rectifier, common-mode choke, and switch, are not directly associated with energy stored in the flyback-transformer core. The remaining power losses, which occur in the output rectifier and clamp zener diode when the flyback transformer releases energy, are now defined as secondary power loss P_{LS}. The loss-allocation factor, Z, defined as the ratio of P_{LS} to P_L, is a scale factor that distributes the power losses between the primary and secondary. Z is typically 0.5, which means P_{LS} is usually 50% of P_L.

$$P_L=P_O\cdot((1-leta)/leta);$$

$$Z=P_{LS}/P_L.$$

Bias voltage, V_B, depends on the feedback circuitry and is usually 10 to 30V. For a bridge rectifier, a conduction time, t_C, of 3 msec is typical.

For a universal or 115V-ac input, start with a standard input-filter-capacitor value, C_{IN} (in microfarads), of two to three times the output power in watts. For a 230V-ac input, use a standard value in microfarads equal to the output power in watts. For example, 60 to 90 µF is a suitable capacitance range for a universal-input, 30W supply; 68 and 82 µF are both standard values within this range.

The reflected output voltage, V_{OR}, appears across the transformer primary when the switch is off and current is flowing through the secondary and the output-rectifier diode. Transformers optimized for TOPSwitch applications should have a maximum V_{OR} of 60V or less for the TOP1xx series (115V-ac input) and 135V or less for the TOP2xx series (universal or 230V-ac input). V_{DS} is the on-state TOPSwitch voltage (typically, 10V).

Output-rectifier forward-voltage drop V_D depends on the type of diode you select. Use 0.4V for Schottky diodes and 0.7V for ultra-fast-recovery pn-junction diodes. Bias-winding-diode forward-voltage drop (V_{DB}) is also typically 0.7V.

K_{RP} determines how far into the continuous mode a flyback transformer operates. Continuous-mode transformers optimized for TOPSwitch applications that operate from 100 to 115V-ac or universal-input voltage should have a minimum K_{RP} of 0.4. Applications that operate from 230V ac should have a minimum K_{RP} of 0.6. Discontinuous-mode transformers optimized for TOPSwitch applications always have a K_{RP} of 1.0.

Core and bobbin manufacturers' data sheets specify the following effective parameters: cross-sectional area A_E (in centimeters squared), path length L_E (in centimeters), ungapped inductance A_L (specified in either millihenries/(1000 turns)^2 or nanohenries/T^2), and physical bobbin-winding width BW (in millimeters). Margin width M, determined by insulation methods and regulatory requirements, is usually set at 2.5 to 3.0 mm for margin-wound transformers or to zero for transformers wound with insulated wire.

Primary windings normally use one or two layers, L. Higher numbers of layers reduce coupling and increase cost, capacitance, and leakage inductance. The number of secondary turns, N_S, is a key iteration variable. One turn per volt of output voltage is a good starting value for N_S. (For example, start with five turns for a 5V output.)

Current-waveform shape parameters

D_{MAX} is the actual duty cycle that occurs when the power supply delivers the maximum output power from the minimum input voltage, V_{MIN}. D_{MAX} has an upper limit equal to the minimum value of the TOPSwitch data-sheet parameter, DC_{MAX} (64%).

$$D_{MAX}=(V_{OR})/(V_{OR}+V_{MIN}V_{DS}),$$

$$I_{AVG}=P_O/(leta\cdot V_{MIN}),$$

$$I_P=I_{AVG}\cdot(2/((2K_{RP})\cdot D_{MAX})), \text{ and}$$

$$I_{RMS}=I_P\cdot D_{MAX}\cdot(((K_{RP}^2)/3)K_{RP}+1)^{0.5},$$

where I_{RMS} is the rms value of the primary current.

The flyback-transformer energy equation defined below determines the primary inductance, L_P, in microhenries. The flyback transformer stores energy proportional to the square of the primary current. When the switch is on, the primary current linearly ramps up over a current range, I_R, and increases the energy stored in the flyback-transformer core. When the switch turns off, the stored-energy increment associated with I_R transfers to the load and the secondary losses in the rectifier and clamp diode. You can now calculate L_P from P_O; *leta*, Z; I_P; switching frequency f_S; and K_{RP}, which determines I_R.

$$P_O\cdot((Z\cdot(1\mathit{leta}))+\mathit{leta})/\mathit{leta}$$
$$(f_S I_P^2\cdot K_{RP}\cdot 1(K_{RP}/2))$$

The number of primary turns, N_P, depends on N_S, V_O, V_D, the effective primary voltage ($V_{MIN}V_{DS}$), and D_{MAX}:

$$N_P=N_S\cdot(V_{MIN}V_{DS})\cdot D_{MAX}/((V_O+V_D)\cdot(1D_{MAX})).$$

Determine the primary insulated-wire diameter, OD, in millimeters from the effective bobbin width, BW_E, and N_P:

$$OD=BW_E/N_P.$$

The actual magnet-wire outside diameter is slightly larger than the diameter, DIA, of the bare-copper conductor. Insulation thickness varies inversely with the AWG size of the bare-copper conductor, which means that smaller diameter conductors have thinner insulation. Data from several manufacturers yields an empirical expression for total insulation thickness, INS, in millimeters, as a function of heavy insulated-magnet-wire outside diameter in millimeters.

$$INS=(0.0594\cdot\log(OD))+0.0834.$$
$$DIA=ODINS.$$

Another empirical equation determines the AWG for magnet wire with a given DIA. Integer AWG values are standard sizes, so before proceeding with the current-capacity or circular-mils-per-ampere (CMA) calculation, you should always round up the calculated AWG value to the next integer or standard value, the next-smaller standard-conductor diameter.

$$AWG=9.97\cdot(1.8277(2\cdot\log(DIA))).$$

The cross-sectional area of magnet wire for transformer windings is usually specified in circular mils (CM). A circular mil is the cross-sectional area of a wire that has a diameter of 1 mil (0.0254 mm). The following simple expression relates a bare-conductor wire's effective cross-sectional area in circular mils to its standard AWG size.

$$CM=2^{(50AWG)/3}.$$

Circular mills per ampere is a convenient way to specify a winding's current capacity. CA, which is the inverse of current density, is simply the ratio of cross-sectional area in circular mils to I_{RMS}. CA should be 200 to 500.

$$CMA=CM/I_{RMS}.$$

This calculation completes all necessary primary-winding calculations. You must now calculate the secondary peak current, rms current, average output current, output-capacitor ripple current, and the minimum and maximum conductor diameter of the secondary windings.

The peak secondary current, I_{SP}, is a simple function of I_P, N_P, and N_S.

$$I_{SP}=I_P\cdot N_P/N_S.$$

You can find the secondary rms current, I_{SRMS}, from D_{MAX}, I_{SP}, and K_{RP}. (K_{RP} is identical for the primary and secondary.)

$$I_{SRMS}=I_{SP}\cdot((1D_{MAX})\cdot(((K_{RP}^2)/3)K_{RP}+1))^{0.5};$$
$$I_O=P_O/V_O.$$

The ripple current, I_{RIPPLE}, in output capacitor C_2 is not a true transformer parameter, but you need to know it for capacitor selection. You can easily calculate I_{RIPPLE} from other transformer parameters: I_{SRMS} and I_O.

$$I_{RIPPLE}=(I_{SRMS}^2 I_O^2)^{0.5}.$$

The minimum secondary bare-conductor diameter, DIA_S, (in millimeters) is based on the previously calculated circular mils per ampere and secondary rms current. You can calculate the minimum secondary bare-conductor circular mils per ampere from the primary circular mils per ampere and the secondary I_{SRMS}:

$$CMS=CMA\cdot I_{SRMS}.$$

You can then calculate the minimum secondary wire gauge, AWG_S, from another empirical equation. Always round AWG_S down to the next integer value. Doing so selects the next larger standard wire size.

$$AWG_S=9.97\cdot(5.017\log(CMS)).$$

Now, determine DIA_S in millimeters.

$$DIA_S=((4\cdot(2^{(50AWG)/3})/(1.27\cdot 3.14))^{0.5})\cdot 25.4/1000.$$

You must also calculate the maximum outside wire diameter, OD_S, in millimeters for a single layer based on the number of secondary turns and the bobbin width:

$$OD_S=(BW(2\cdot M))/N_S.$$

You can now calculate the secondary-wire insulation thickness from the outside diameters of the bare conductor (determined by circular mils per ampere) and the insulated wire (determined by the number of turns and the effective bobbin width). Note that the secondary insulation thickness, INS_S, in millimeters is the insulation-wall thickness rather than the total insulation thickness you use in the primary-winding calculation.

$$INS_S=(OD_S DIA_S)/2.$$

Obviously, if INS_S is not a positive number, you must perform another transformer-design iteration with either more secondary layers, a smaller number of secondary turns, or a transformer core with a wider bobbin.

For insulated-wire secondaries, INS_S must equal or exceed the selected wire's insulation thickness. Parallel combinations of wire with half the diameter may be easier to wind and terminate, but the effective secondary circular mils per ampere will be half the value of the single winding.

The maximum drain voltage, V_{DRAIN}, is the sum of the maximum dc-input voltage, V_{MAX}, an estimated drain clamp-voltage term based on V_{OR}, and an estimated voltage term related to typical blocking-diode forward recovery. (Refer to application note AN-14 for more detail.)

$$V_{DRAIN}=V_{MAX}+(2.1 \cdot V_{OR})+20.$$

You can easily calculate the parameters of auxiliary windings for additional outputs. V_X is the output voltage, V_{DX} is the diode voltage drop, N_X is the number of turns, and PIV_X is the peak-inverse voltage associated with the auxiliary winding.

$$N_X=N_S \cdot (V_X+V_{DX})/(V_O+V_D).$$
$$PIV_X=V_X+(V_{MAX} \cdot (N_X/N_P)).$$

You can now use spreadsheet iteration to reach a final and acceptable solution for the flyback-transformer design. Iterate N_S or K_{RP} until the maximum flux density, B_M, is between indicated limits (2000 to 3000 gauss), and check that the gap length, L_G, is higher than the indicated minimum value (0.051 mm). B_M decreases and L_G increases as you increase N_S or K_{RP}.

Examine circular mils per ampere. If circular mils per ampere is less than the specified lower limit of 200, consider increasing the number of primary layers from one to two or using the next-larger core size and performing a new iteration. If CMA is greater than 500, consider using the next-smaller core size. (A greater-than-500 CMA simply means that the wire diameter is oversized for the expected rms current.)

The transformer design is now complete. The improved spreadsheet contains all transformer-design parametric information that a transformer vendor requires.

References

1. Leman, Brooks, "**Finding the keys to flyback power supplies produces efficient design**." *EDN*, April 13, 1995, pg 101.
2. Bisci, J, "Magnet wire: Selection determines performance," PCIM, October 1994, pg 37.
3. 1996/97 data book and design guide, Power Integrations Inc, Sunnyvale, CA, 1996.

Author's biography

Brooks Leman is a principal application engineer at Power Integrations Inc (Sunnyvale, CA), where he has worked for eight years. He develops low-cost flyback and power-factor-correction circuits using highly integrated power-supply devices. Leman holds a BSEE and MSEE from Santa Clara University (Santa Clara, CA) and is a member of Tau Beta Pi, the EE honor society. His leisure activities include soccer and skiing.
