

OPTIMIZE FLYBACK MAGNETICS TO EMPOWER THE PD

Introduction

In PoE systems, performance of the flyback converter used to step down the 48-V supply in the powered device depends on careful transformer design.

A growing number of electronic devices require both data connectivity and a power source to function effectively. The IEEE 802.3af standard outlines a method for delivering power and data over LAN cables using the existing Ethernet infrastructure. A key element of this standard is that the power delivery must be isolated on the user or powered device (PD) side. This isolation requires the use of an isolated power topology within the PD.

The most cost-effective power topology to provide this isolation with acceptable efficiency is the continuous-mode flyback. To implement the flyback effectively in PDs, designers must understand the basic operation of this topology and the equations needed to design the flyback transformer.

In Power-over-Ethernet (PoE) applications, the power-sourcing equipment (PSE) supplies a voltage of 36 Vdc to 57 Vdc to a PD. However, very few, if any, devices will require voltages in this range. As a result, a power supply inside the PD must convert the available power to the regulated voltage required by the electronics within the PD. The designer of the power supply for the PD must find the most cost-effective, space- and energy-efficient means of accomplishing this task.

Due to the requirement for 1500- V_{RMS} isolation, an isolated power topology must be used. For the power levels required for the existing PoE specification (less than 13 W), the flyback converter is the simplest and least-expensive isolated topology. Traditionally, for non-PoE applications at these power levels, a discontinuous-mode flyback is generally used because it usually requires a smaller transformer. It also provides faster transient response due to a lower primary inductance, and the feedback loop is easier to stabilize than a continuous-mode flyback.

However, for PoE applications where efficiency is critical, the respective sizes of the transformer needed for the continuous and discontinuous modes are essentially the same and the lower peak currents of the continuous flyback improve overall efficiency. Furthermore, the right half-plane zero in the continuous-mode flyback is sufficiently high so as not to affect the stabilization of the feedback loop.

As a result, the vast majority of published PD power reference designs use a continuous-mode flyback topology. Reference designs from several power IC vendors demonstrate that this topology is capable of efficiencies in excess of 90%, which is sufficient for most PoE applications.

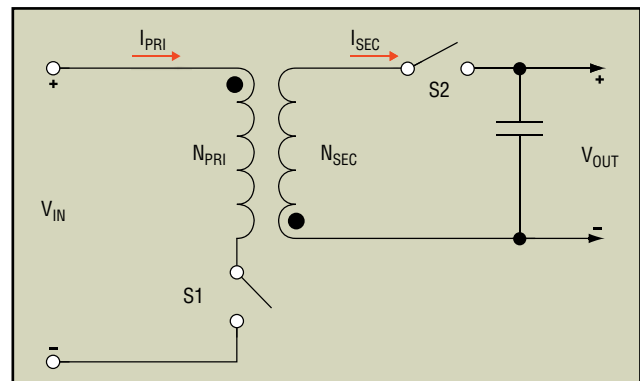


Fig. 1. Operation of the flyback converter can be modeled with this simplified schematic where switches S1 and S2 can represent either diodes or MOSFETs.

Overview of Flyback Topology

A basic understanding of the continuous flyback topology is necessary to derive the equations for the PoE power transformer design. A simplified schematic of the flyback topology is shown in **Fig. 1**. The primary switch S1 is a MOSFET, and the secondary switch S2 can either be a diode or a second MOSFET operated as a high-efficiency synchronous rectifier. Regardless of the exact implementation, the flyback topology's switching cycle has two distinct stages.

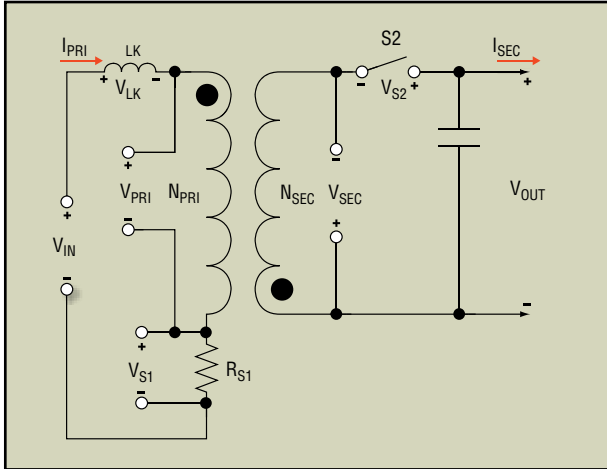


Fig. 2. During the on-time stage of the flyback's switching cycle, the transformer's primary-winding inductance appears as a separate element and switch S1 is represented by a resistor.

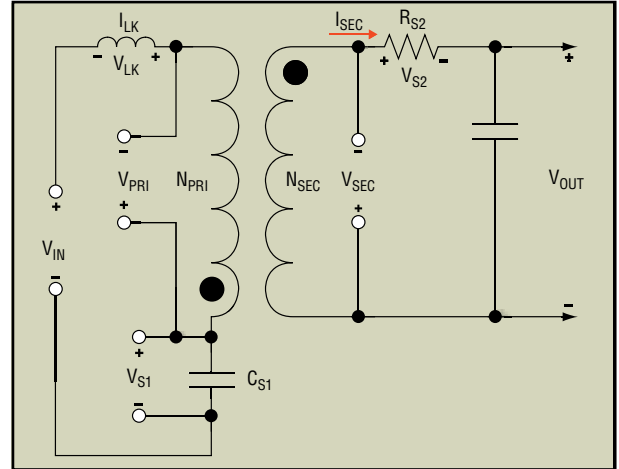


Fig. 3. During the off-time stage of the flyback's switching cycle, the open S1 is modeled by its output capacitance and the closed S2 is modeled as a resistor

The first stage is the on-time stage, in which S1 is closed and the input is connected to the primary winding (N_{PRI}) of the transformer for a period of time defined by t_{ON} .

$$(Eq. 1) \quad D = \frac{t_{ON}}{T}$$

During this stage, S2 is opened (representing either the turning off of the MOSFET or the reverse biasing of the diode), and the schematic reduces to the circuit in **Fig. 2** In this schematic, the transformer's primary-winding leakage inductance is shown as a separate inductor, and the resistance of S1 is shown as a resistive element, along with their respective voltage drops. When S1 is closed, current flows through the primary winding, but the change in current is limited by the primary winding's inductance according to Faraday's law:

$$(Eq. 2) \quad \Delta I_{PRI} = \frac{(V_{PRI} \times t_{ON})}{L_{PRI}},$$

$$(Eq. 3) \quad \text{or } \Delta I_{PRI} = (V_{in} - V_{S1}) \times D \times \frac{T}{L_{PRI}},$$

where ΔI_{PRI} is the change in primary-winding current (amps), V_{PRI} is the voltage across the primary winding (volts), t_{ON} is the duration of the S1 closure and L_{PRI} is the primary winding inductance (microhenrys).

The changing primary current creates a magnetic field in the core of the transformer, which in turn induces a voltage across the secondary winding.

$$(Eq. 4) \quad \Delta B_{ON} = (V_{in} - V_{S1}) \times \left(\frac{t_{ON}}{(N_{PRI} \times A_{EFFECTIVE})} \right)$$

$$(Eq. 5) \quad V_{SEC} = \Delta B_{ON} \times N_{SEC} \times \left(\frac{A_{EFFECTIVE}}{t_{ON}} \right)$$

However, because the S2 is open, no current can flow and the energy created in the primary winding is stored in the magnetic field of the core. The energy stored is equal to:

$$(Eq. 6) \quad E_{IN} = \frac{1}{2} \times L_{PRI} \times (\Delta I_{PRI})^2,$$

And, by definition, power is simply energy divided by time:

$$(Eq. 7) \quad P_{IN} = \frac{1}{2} \times L_{PRI} \times \frac{(\Delta I_{PRI})^2}{T}.$$

It should be noted that, even though no current is flowing in the secondary, voltage is present (V_{SEC}) and does create stress across S2, which is an important factor in determining the turns ratio (N_{PRI}/N_{SEC}):

$$(Eq. 8) \quad V_{S2} = V_{OUT} + V_{in} \times \left(\frac{N_{PRI}}{N_{SEC}} \right),$$

where V_{S2} is the voltage across S2 (volts), V_{IN} is the input voltage (volts) and V_{OUT} is the output voltage (volts). Eq. 8 implies that the turns ratio must be large enough to prevent excessive stress across S2. It also should be noted that, during this time, the primary leakage inductance is storing energy, but that this energy represents a loss in the circuit, as there is no mechanism to transfer it to the secondary. During the on-time stage, the output current is supplied by the output capacitor, because no current is flowing through the transformer's secondary winding.

Off-Time Stage

During the second stage of the flyback-topology switching cycle, S1 is opened and S2 is closed. The schematic reduces to that shown in **Fig. 3**. The primary current is zero because S1 is closed. However, the magnetic field in the transformer cannot change instantaneously and, therefore, looks for a path to dump the stored energy. It does this by reversing the polarity on the secondary winding and sourcing current to the load. The current being delivered to the load ramps down linearly according to Faraday's law starting from its peak value:

$$(Eq. 9) \quad I_{SEC_{PK}} = I_{PRI_{PK}} \times V_{in} \times \left(\frac{N_{PRI}}{N_{SEC}} \right),$$

where $I_{SEC_{PK}}$ is the peak secondary-winding current (amps).

The slope of the ramp is:

$$(Eq. 10) \quad \Delta I_{SEC} = (V_{OUT} + V_{S2}) \times \left(\frac{t_{OFF}}{L_{SEC}} \right),$$

and knowing that $L_{SEC} = \frac{L_{PRI}}{N^2}$, where N equals N_{SEC} / N_{PRI} .

$$(Eq. 11) \quad \Delta I_{SEC} = (V_{OUT} + V_{S2}) \times t_{OFF} \times \left(\frac{N}{L_{PRI}} \right)^2.$$

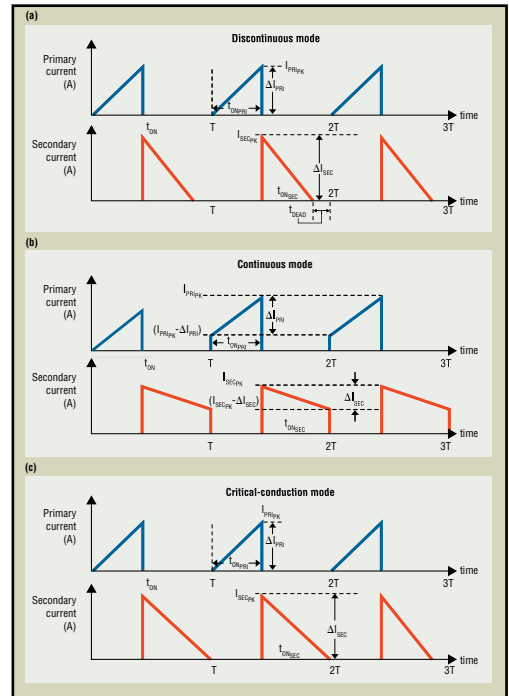


Fig. 4. These waveforms show the timing relationships between the primary and secondary winding currents for the continuous (a), discontinuous (b) and critical-conduction modes (c).

The changing current (ΔI_{SEC}) reduces the flux density of the core:

$$(Eq. 12) \quad \Delta B_{OFF} = \frac{(L_{SEC} \times \Delta I_{SEC})}{(N_{SEC} \times A_{EFFECTIVE})}$$

Substituting in Eq. 11:

$$(Eq. 13) \quad \Delta B_{OFF} = (V_{OUT} + V_{S2}) \times \frac{t_{OFF}}{(N_{SEC} \times A_{EFFECTIVE})}$$

During steady-state operation, the magnitude of the decrease in the magnetic field (DB) during the off time must be equal to the magnitude of the increase during the on time. This leads to the relationship:

$$(Eq. 14) \quad \frac{((V_{IN} - V_{S1}) \times t_{ON})}{N_{PRI}} = \frac{((V_{OUT} + V_{S2}) \times t_{OFF})}{N_{SEC}}$$

$$(Eq. 15) \quad \text{Or, } \frac{N_{PRI}}{N_{SEC}} = \frac{(V_{IN} - V_{S1}) \times t_{ON}}{(V_{OUT} + V_{S2}) \times t_{OFF}}$$

It is important to note that although there is no current flowing in the primary winding during the off-time stage, the primary winding does see the reflected voltage of the secondary winding (V_{SEC}):

$$(Eq. 16) \quad V_{PRI} = V_{SEC} \times \left(\frac{N_{PRI}}{N_{SEC}} \right) = (V_{OUT} + V_{S2}) \times \left(\frac{N_{PRI}}{N_{SEC}} \right)$$

When S2 is turned off, the energy in the leakage inductance of the primary winding must be dumped. However, because S1 is open, there is no immediate path for current to flow, so the voltage across the leakage inductance rapidly increases until it can push current through the parasitic elements of the capacitance of the transformer primary and the output capacitance of the MOSFET. This leakage spike along with the induced voltage on the primary winding creates a voltage stress S1:

$$(Eq. 17) \quad V_{S1} = V_{IN} + V_{LK} + V_{SEC} \times (N_{PRI}/N_{SEC}),$$

$$(Eq. 18) \quad \text{And } V_{LK} = I_{PRI_{PK}} \times \left(\frac{L_{LK}}{(C_{PRI} + C_{MOSFET_{DS}})} \right)^{1/2},$$

where V_{LK} is the voltage across the leakage inductance (volts), C_{PRI} is the capacitance across the primary winding (Farads) and $C_{MOSFET_{DS}}$ is the drain-source capacitance of the MOSFET (Farads).

Eq. 17 implies that the turns ratio should be small enough to prevent excessive stress on S1. In addition to stressing S1, which often requires extra snubbing circuitry for protection, the leakage inductance delays the transfer of power from the primary to the secondary. Therefore, minimizing leakage inductance is a key element to flyback-transformer design.

Continuous, Discontinuous Flyback Modes

Fig. 4 shows three different sets of primary and secondary waveforms. In both continuous and discontinuous modes, the initial primary current starts at zero. In **Fig. 4a** during the off-time the secondary current ramps down to zero because all of the energy stored in the magnetic field during the on-time of the transformer has been delivered to the load before the start

of the next cycle. This creates a period of time (t_{DEAD}) in which no current is flowing in the transformer and the circuit is said to be in discontinuous mode. As can be seen in **Fig. 4** at the start of the next cycle, the primary current again ramps up from zero.

In continuous mode, the initial primary waveform starts from zero, but during the off time, the secondary current does not deliver all of the stored energy to the load before the start of the next cycle. As a result, energy is left in the transformer's magnetic field. At all times, current is flowing in the transformer ($t_{DEAD} = 0$) and the circuit is said to be continuous. As can be seen in **Fig. 4b** the primary current begins its ramp at a nonzero point, defined as:

$$(Eq. 19) \quad I_{PRI_MIN} = I_{SEC_MIN} \times (N_{S_EC}/N_{P_RI}),$$

where I_{PRI_MIN} is the minimum primary-winding current (A) and I_{SEC_MIN} is the minimum secondary-winding current. It should be apparent from these waveforms that the two main factors, other than the turns ratio, that influence whether a circuit is continuous or discontinuous are the transformer primary inductance (which controls the slope of the current ramp) and the required output current (which controls how much energy is taken from the transformer). As will be seen later in the article, although it is possible to design a transformer that always operates in discontinuous mode, any continuous-mode transformer will eventually go discontinuous as the load current is decreased below a minimum threshold I_{OUT_MIN} .

Flyback Topology Waveforms

Before starting any flyback-transformer design it is necessary to understand how the previously discussed current waveforms relate to the output current I_{OUT} , I_{RMS} (which affects heating in the transformer) and I_{PRIPK} (which affects transformer saturation). Once this is understood, it is easy to derive the required design parameters. Figs. 5 and 6 show a close-up of the primary and secondary waveforms for a discontinuous- and continuous- mode flyback transformer. Note that the dc value of any waveform is simply the average value of the waveform over the fundamental period of the switching cycle (T) and that, by definition, the rootmean-square (rms) current is:

$$(Eq. 20) \quad I_{RMS} = \sqrt{\frac{1}{T_2 - T_1} \int_{T_2}^{T_1} (i(t))^2 dt},$$

where T_1 and T_2 define the start and stop times of the period, respectively. Evaluating the integral and calculating the average (dc) value leads to the following equations:

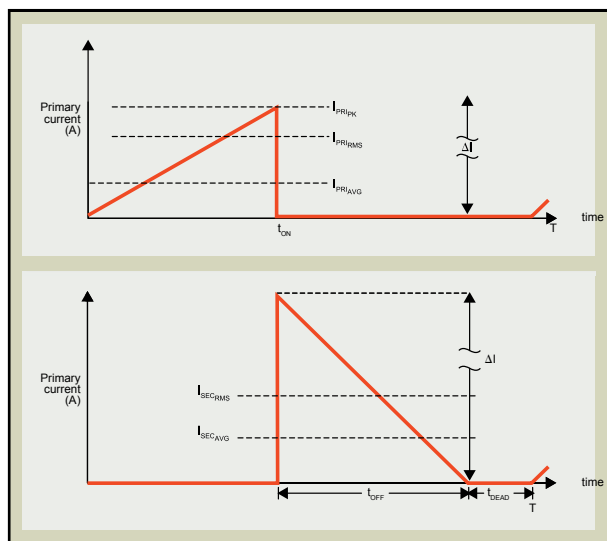


Fig. 5. The average and rms waveforms of the primary (upper) and secondary (lower) winding currents of the flyback transformer in the discontinuous mode are functions of duty cycle, dead time, and ΔI .

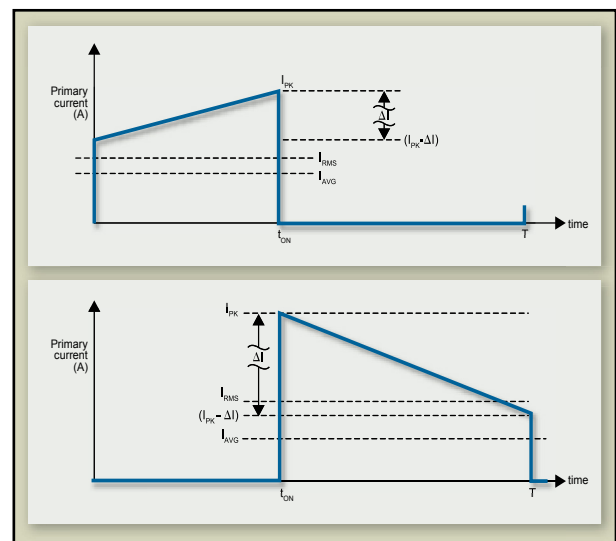


Fig. 6. The average and rms waveforms of the primary (upper) and secondary (lower) winding currents of the flyback transformer in the discontinuous mode are functions of peak currents, ΔI , and duty cycle, but not deadtime.

Discontinuous mode:

$$(Eq. 21) \quad I_{PRI_{RMS}} = \Delta I_{PRI} \times \left(\frac{D}{3}\right)^{1/2}$$

$$(Eq. 22) \quad I_{SEC_{RMS}} = \Delta I_{SEC} \times \left(\frac{\left(1 - D - \left(\frac{t_{DEAD}}{T}\right)\right)}{3}\right)^{1/2},$$

and continuous mode:

$$(Eq. 23) \quad I_{PRI_{RMS}} = \left(D \times ((I_{PRI_{PK}})^2 - I_{PRI_{PK}} \times \Delta I_{PRI} + \frac{(\Delta I_{PRI})^2}{3})\right)^{1/2}$$

$$(Eq. 24) \quad I_{SEC_{RMS}} = \left(\left(1 - D \times ((I_{SEC_{PK}})^2 - I_{SEC_{PK}} \times \Delta I_{SEC} + \frac{(\Delta I_{SEC})^2}{3})\right)\right)^{1/2}.$$

Discontinuous mode:

$$(Eq. 25) \quad I_{PRI_{AVG}} = \frac{1}{2} \times I_{PRI_{PK}} \times D$$

$$(Eq. 26) \quad I_{SEC_{AVG}} = \frac{1}{2} \times I_{SEC_{PK}} \times \left(1 - D - \left(\frac{t_{DEAD}}{T}\right)\right),$$

and continuous mode:

$$(Eq. 27) \quad I_{PRI_{AVG}} = \left(I_{PRI_{PK}} - \left(\frac{\Delta I_{PRI}}{2}\right)\right) \times D$$

$$(Eq. 28) \quad I_{SEC_{AVG}} = \left(I_{SEC_{PK}} \times \left(\frac{\Delta I_{SEC}}{2}\right)\right) \times (1 - D).$$

Critical-Conduction Mode

It should be clear that the boundary between continuous mode and discontinuous mode, also known as the critical-conduction mode (CCM), occurs when the secondary of the transformer delivers all of the stored energy at the end of the switching period T (as shown at the bottom of **Fig. 4c**). In this case, t_{DEAD} is equal to zero:

$$T_{ON} + T_{OFF} = T. \text{ Or, using Eq. 1, } T_{OFF} = (1 - D) \times T.$$

If all of the stored energy is delivered to the output (minus system losses or efficiency), then:

$$E_{IN} = \frac{E_{OUT}}{\eta} \text{ and } P_{IN} = \frac{P_{OUT}}{\eta}, \text{ where } \eta \text{ equals efficiency.}$$

$$P_{OUT} = I_{OUT} \times (V_{OUT} + V_{S2}).$$

$$\text{And from Eq. 7, } P_{IN} = \frac{1}{2} \times L_{PRI} \times \frac{(\Delta I_{PRI})^2}{T}$$

$$\text{From Eq. 3, } \Delta I_{PRI} = (V_{IN} - V_{S1}) \times D \times \left(\frac{T}{L_{CCM}} \right),$$

$$\text{so } \frac{1}{2} \times L_{CCM} \times \frac{\left((V_{IN} - V_{S1}) \times D \times \left(\frac{T}{L_{PRI}} \right) \right)^2}{T} = I_{OUT} \times (V_{OUT} - V_{S2}).$$

Simplifying gives:

$$\text{(Eq. 31) } L_{CCM} = \eta \times (V_{IN} - V_{S1})^2 \times \left(\frac{D^2}{(2 \times I_{OUT} \times (V_{OUT} + V_{S2}) \times f)} \right)$$

$$\text{(Eq. 32) } I_{OUT} = \eta \times (V_{IN} - V_{S1})^2 \times \left(\frac{D^2}{(2 \times I_{CCM} \times (V_{OUT} + V_{S2}) \times f)} \right).$$

For any given application, there is a primary-winding inductance that will dictate if the transformer operation is discontinuous across the full output-current range ($L_{PRI} < L_{CCM}$), or if it is in discontinuous mode below some minimum output current, and in continuous mode ($L_{PRI} > L_{CCM}$) above this current.

Eq. 31 defines inductance at the CCM for a given duty cycle. Typically, a duty cycle is selected and the turns ratio is then calculated to produce the desired duty cycle. The turns ratio will affect the stresses on the primary and secondary switches. As shown in Eqs. 8 and 17, the turns ratio must be large enough to minimize stress on the S2 and small enough to minimize stress on the S1. This bounds the absolute maximum and minimum turns ratio.

Once the turns ratio is selected, the value of the primary inductance (L_{PRI}) is chosen to ensure that the transformer always runs in discontinuous mode up to the maximum output (I_{OUTMAX}) or that it runs in continuous mode down to some minimum output level (I_{OUTMIN}). As a result, Eq. 31 can be redefined as follows:

For discontinuous mode:

$$L_{PRI} \leq \eta \times (V_{IN} - V_{S1})^2 \times \left(\frac{D^2}{(2 \times I_{OUTMAX} \times (V_{OUT} + V_{S2}) \times f)} \right).$$

For continuous mode:

$$\text{(Eq. 34) } L_{PRI} \geq \eta \times (V_{IN} - V_{S1})^2 \times \left(\frac{D^2}{(2 \times I_{OUTMIN} \times (V_{OUT} + V_{S2}) \times f)} \right).$$

It should be noted that in continuous mode, because there is no dead time, the turns ratio alone for a given input and output voltage sets the duty cycle. However, for discontinuous mode operation, the existence of the dead time means that duty cycle is affected by the turns ratio and the primary inductance.

The easiest way to illustrate the use of the previous equation is to run through a quick design example that will also illustrate the differences between the continuous mode and discontinuous mode flybacks.

Design Example

For simplicity, the equations will be designed for a PoE input (33 V to 57 V) with a single output (5 V/12 W) switching at 200 kHz. It will be assumed that the primary switch voltage drop (VS1) is 0.4 V, the secondary switch voltage drop (VS2) is 0.3 V and that the overall efficiency is 90%. Multiple output applications can use the same design procedure by selecting one of the outputs as the master and combining all of the output power onto this one output. This will allow the selection of the duty cycle, primary inductance, turns ratio, primary peak and root-mean-square (rms) currents. The only difference is in calculating the rms currents on each of the secondary windings. This requires careful attention to the secondary waveforms of each output.

Step One

Select a maximum duty cycle, calculate the resultant turns ratio and analyze the voltage stresses on the primary and secondary switches. If the stresses are too high, the duty cycle can be adjusted and/or the switch ratings can be increased.

Select a duty cycle maximum equal to 45%.

Calculate the turns ratio using Eq. 15 and assuming that $t_{\text{DEAD}} = 0$:

$$\frac{N_{\text{PRI}}}{N_{\text{SEC}}} = \frac{(33-0.4)}{(5+0.3)} \times \left(\frac{0.45}{(1-0.45)} \right) = 5.03.$$

Select the actual turns ratio of $N_{\text{PRI}}/N_{\text{SEC}} = 5$.

Recalculate the actual duty cycles based on actual turns using Eq. 15:

$$D_{\text{MAX}} = \frac{1}{\left(1 + \frac{(33-0.4)}{(5+0.3)} \times \left(\frac{1}{5} \right) \right)} = 0.448 = 44.8\%.$$

$$D_{\text{MIN}} = \frac{1}{\left(1 + \frac{(57-0.4)}{(5+0.3)} \times \left(\frac{1}{5} \right) \right)} = 0.319 = 31.9\%.$$

Analyze the voltage stresses using Eqs. 8 and 17:

$$V_{\text{S2}} = (5 + 0.3) + \left(\frac{57}{5} \right) = 16.7 \text{ V.}$$

$$\text{Assuming } V_{\text{LK}} = 0.3 \times V_{\text{IN}} : V_{\text{S1}} = 1.3 \times 57 + 5.3 \times 5 = 100 \text{ V.}$$

Step Two

Calculate the primary inductance at the boundary condition and select inductance for application. If discontinuous, recalculate the actual duty cycles.

Calculate the boundary inductance for discontinuous mode (DCM) using Eq. 33:

$$L_{\text{PRIBOUNDARY}} \leq \frac{0.9 \times (33 - 0.4)^2 \times 0.448^2 \times 1000}{\left(2 \times \left(\frac{12}{5}\right) \times (5+0.3) \times 200\right)} = 37.8 \mu\text{H}.$$

Calculate the boundary inductance for continuous mode (CM) using Eq. 34. Assuming that $I_{\text{OUT_MIN}} = 0.5 \times I_{\text{OUT_MAX}}$:

$$L_{\text{PRI}} \geq \frac{0.9 \times (33 - 0.4)^2 \times 0.448^2 \times 1000}{\left(2 \times \left(\frac{6}{5}\right) \times (5+0.3) \times 200\right)} = 75.6 \mu\text{H}.$$

Select the inductance for application with a 5% margin:

$$L_{\text{PRI_DCM}} = 37.8 \times 0.95 \cong 36 \mu\text{H}$$

$$L_{\text{PRI_CM}} = 7.56 \times 1.05 \cong 80 \mu\text{H}.$$

For DCM operation, the inductance chosen will affect the dead time (t_{DEAD}) and, therefore, the actual duty cycles.

Recalculate the DCM duty cycle using Eq. 31:

$$D_{\text{DCM_MAX}} = \left(\frac{\left(2 \times 36 \times \left(\frac{12}{5}\right) \times 5.3 \times 200\right)}{(0.9 \times (33 - 0.4)^2 \times 1000)} \right)^{1/2} = 43.7\%$$

$$D_{\text{DCM_MIN}} = \left(\frac{\left(2 \times 36 \times \left(\frac{12}{5}\right) \times 5.3 \times 200\right)}{(0.9 \times (57 - 0.4)^2 \times 1000)} \right)^{1/2} = 25.2\%$$

Calculate the off time ($D_{\text{OFF}} = t_{\text{OFF}}/T$) from Eq. 15:

$$D_{\text{DCM_OFF}} = \frac{1}{5} \times \frac{(33 - 0.4)}{(5 + 0.3)} \times 0.437 = 53.8\%$$

Step Three

Calculate the worst-case ΔI , I_{PK} and I_{RMS} for each winding.

Calculate the secondary ripple current from Eq. 11.

For DCM:

$$\Delta I_{\text{SEC}} = (5 + 0.3) \times (0.538) \times \left(\frac{1000}{200}\right) \times \left(\frac{5^2}{36}\right) = 9.9 \text{ A}.$$

For CM:

$$\Delta I_{\text{SEC_VIN_MIN}} = (5 + 0.3) \times (1 - 0.448) \times \left(\frac{1000}{200}\right) \times \left(\frac{5^2}{80}\right) = 4.58 \text{ A}.$$

Parameter	Discontinuous	Continuous
$N = N_p/N_s$	5	5
L_{pri}	36 μH	80 μH
$I_{\text{pk pri}}$	2 A	1.3 A
$E = 1/2 \times L_{\text{PRI}} \times I_{\text{PK}}^2$	70.6 μJ	70.6 μJ
I_{RMSPRI}	0.76 A	0.61 A
I_{RMSSEC}	4.2 A	3.4 A

Table. Design example parameters for PoE flyback, where V_{IN} equals 33 V to 57 V, 200 kHz, and V_{OUT} equals 5 V/12 W, and efficiency equals 90%.

For CM:

$$\Delta I_{\text{SECVINMAX}} = (5 + 0.3) \times (1 - 0.319) \times \left(\frac{1000}{200}\right) \times \left(\frac{5^2}{80}\right) = 5.64 \text{ A.}$$

Calculate the secondary peak currents (I_{PKsec}) using Eq. 28.

For DCM:

$$(I_{\text{PKsec}}) = I_{\text{SECPK}} = \Delta I_{\text{SEC}} = 9.9 \text{ A}$$

For CM:

$$I_{\text{SECPKVINMIN}} = \frac{\left(\frac{12}{5}\right)}{(1 - 0.448) + \left(\frac{4.58}{2}\right)} = 6.63 \text{ A}$$

For CM:

$$I_{\text{SECPKVINMAX}} = \frac{\left(\frac{12}{5}\right)}{(1 - 0.319) + \left(\frac{5.64}{2}\right)} = 6.34 \text{ A}$$

Calculate the secondary rms currents (I_{RMSSEC}) using Eqs. 22 and 24.

For DCM:

$$\Delta I_{\text{SECRMS}} = 9.9 \times \left(\frac{0.538}{3}\right)^2 = 4.2 \text{ A.}$$

For CM:

$$I_{\text{SECRMSVINMIN}} = \left((1 - 0.448) \times \left(6.63^2 - (4.58 \times 6.63) + \left(\frac{4.58^2}{3} \right) \right) \right)^{1/2} = 3.2 \text{ A}$$

For CM:

$$I_{\text{SECRMSVINMAX}} = \left((1 - 0.319) \times \left(6.34^2 - (5.64 \times 6.34) + \left(\frac{5.64^2}{3} \right) \right) \right)^{1/2} = 3.4 \text{ A}$$

Calculate the primary peak current (I_{PRIpk}) using Eq. 9.

For DCM:

$$I_{\text{PRIpk}} = \frac{I_{\text{SECPK}}}{N} = \frac{9.9}{5} = 1.98 \text{ A.}$$

For CM:

$$I_{\text{PRIpkVINMIN}} = \frac{6.63}{5} = 1.33 \text{ A.}$$

For CM:

$$I_{PRI PK VIN MAX} = \frac{6.34}{5} = 1.28 \text{ A.}$$

Calculate the primary rms current ($I_{RMS PRI}$) using Eqs. 21 and 23.

For DCM:

$$I_{PRI RMS VIN MIN} = 1.98 \times \left(\frac{0.437}{3} \right)^{1/2} = 0.76 \text{ A.}$$

For DCM:

$$I_{PRI RMS VIN MAX} = 1.98 \times \left(\frac{0.252}{3} \right)^{1/2} = 0.58 \text{ A.}$$

Note: For CM, $\Delta I_{PRI} = \frac{\Delta I_{SEC}}{N}$.

For CM:

$$I_{PRI RMS VIN MIN} = \left(0.448 \times \left(1.33^2 - 1.33 \times \left(\frac{4.58}{5} \right) + \left(\frac{\left(\frac{4.58}{5} \right)^2}{3} \right) \right) \right)^{1/2} = 0.61 \text{ A.}$$

For CM:

$$I_{PRI RMS VIN MAX} = \left(0.448 \times \left(1.33^2 - 1.33 \times \left(\frac{4.58}{5} \right) + \left(\frac{\left(\frac{4.58}{5} \right)^2}{3} \right) \right) \right)^{1/2} = 0.44 \text{ A.}$$

Typically, it is not the design of the transformer that leads to a nonoptimized flyback design, but rather a failure to calculate the design inputs correctly. Once the primary inductance, primary peak current, turns ratio and primary and secondary rms currents are known, the transformer can be designed following any standard procedure. Typically, a core size is selected and the number of primary turns (to achieve the inductance and peak current without saturating) is calculated.

Knowing the primary turns and the turns ratio sets the secondary turns. To minimize leakage inductance, it is necessary to interleave the primary and secondary windings and keep the windings to one layer per section by choosing the appropriate wire size and number of strands. Once the wire size and strands are selected, the rms currents, along with the core losses, can be used to calculate the total losses in the transformer. If the total losses are too high (due to temperature rise of overall efficiency), it will be necessary to select a larger core size. If the total losses are too low, a smaller core size may be evaluated.

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