Partial BCM Mode Of Operation Improves Efficiency For Primary-Side Regulated Flyback LED Driver

by Roy Mi, Fairchild Semiconductor, Chicago, Ill.; John Jing, Fairchild Semiconductor, Los Angeles, Calif.; and Jason Guo Sr., Fairchild Semiconductor, San Jose, Calif.

This article explains the operating principles of the primary-side regulated (PSR) flyback topology in the LED lighting application with high power factor (PF) and high operating efficiency. The basics of the PSR flyback LED driver are presented and its merits are addressed. A design procedure is proposed and described in detail for engineers to achieve the optimum design balance between the ac-dc LED driver’s efficiency, power factor (PF), THD and its flyback transformer design.

The unique PSR flyback design approach described here allows the driver to operate partially in boundary conduction mode (BCM) at low input voltage. Therefore the driver can achieve high efficiency, while maintaining high power factor and low THD. A quick design example is also given at the end of the article for verification.

LED Lighting Power Requirements

As a solid state light source, light-emitting diodes (LEDs) have been widely used due to their superior longevity, excellent efficacy and friendly environmental attributes. Moreover, LED technology offers excellent flexibility to control the color, illumination pattern as well as the light itself. Today, LEDs are replacing existing lighting sources such as incandescent, fluorescent and HID lamps.

Driving and lighting of LEDs mainly requires constant current, while the LED ballast must maintain high power factor. High PF is necessary for LED ballasts, since LEDs are inherently a reactive and resistive load that causes additional system-related line current harmonics. While the new Energy Star directive for solid-state lighting requires a power factor greater than 0.9 for power levels above 3 W, the ballast input line current harmonics also need to meet the requirements set by IEC61000-3-2 Class C regulations.

To achieve IEC61000-3-2 Class C regulation requirements, a single-stage flyback converter with PFC is typically used for low power (<25-W) LED lighting applications. Furthermore, among variations of the flyback topology, a primary-side-regulated (PSR) flyback can be the most cost-effective solution. By using a single-stage topology with primary-side regulation, an LED lighting board can be implemented with few external components and minimal cost. For example, a PSR flyback does not require an input bulk capacitor, nor the feedback circuitry that would be needed with secondary-side regulation. Fig. 1 shows a single-stage PSR flyback LED driver circuit.

Fig. 1. Single-stage PSR flyback LED driver with high power factor.
Primary-Side-Regulated Flyback

DCM mode

Generally, the discontinuous conduction mode (DCM) of operation is preferred for primary-side regulation because it offers very precise output regulation and unity power factor. The operating principles of a DCM flyback converter are presented in detail in reference 2.

To achieve high power factor and low THD, constant on-time control is usually adopted for DCM flyback converters with fixed switching frequency. Fig. 2 shows the typical theoretical waveforms of the primary-side switch current, the secondary-side diode current and MOSFET switch gate signal.

![Fig. 2. Timing and input current waveforms for a power-factor-corrected flyback converter operating in DCM.](image)

With the constant on-time control, the average input current can be represented by the following equation.

\[
I_{in}(\theta) = \frac{T_{on} \cdot D}{2L_m} \cdot V_{AC}^{peak} \cdot |\sin \theta| = \frac{I_{Lm}^{peak} \cdot D}{2} \cdot |\sin \theta| \quad (1)
\]

Here, \( D \) is the switching duty cycle of the converter and \( L_m \) is the primary magnetizing inductance of the flyback transformer. The above equation indicates that the input current waveform is always following the input voltage. Therefore, the converter achieves unity power factor.

The rms input current then can be given as:

\[
I_{in}^{rms} = \frac{T_{on} \cdot D}{2\sqrt{2}L_m} \cdot V_{AC}^{peak} = \frac{I_{Lm}^{peak}}{2\sqrt{2}} \cdot D \quad (2)
\]

where \( I_{Lm}^{peak} \) is the current at the MOSFET turn-off transient at the peak of the ac input voltage.
To stay in DCM, the maximum duty cycle $D$ must meet the following condition:[2]

$$D \leq \frac{V_R}{V_{AC}^{Peak} + V_R}$$

(3)

where

$$V_R = \eta_{ps} \times (V_{out} + V_d^{Fwd})$$

(4)

$V_R$ is the reflected voltage, which is the voltage that occurs across the primary side of the transformer when the secondary diode is conducting.

Usually, to guarantee the flyback converter operation in DCM for unity power factor and low THD, a transformer with a relatively low turns ratio is used. Such a transformer will lead to small switching duty cycles that create high peak and RMS currents through the MOSFET switch and transformer, resulting in more power loss. A relatively large EMI filter also will be needed due to the high peak switching current.

**BCM mode**

Boundary conduction mode (BCM) and its zero-voltage turn-on switching can minimize switching losses for a single-stage power-factor-corrected flyback converter. The operating principles of a single-stage PFC flyback converter operating in BCM are presented in detail in reference 2. However, a key aspect of converter operation that needs to be noted here is that, unlike a flyback in DCM operation, a BCM-operated flyback is controlled with constant on-time and variable switching frequency. As a result, the BCM flyback is usable for many applications that need a relatively high PF but do not need a low THD (i.e. lower than 10%). Fig. 3 shows the theoretical primary-side switch current, the secondary-side diode current and MOSFET gating signal.

![Fig. 3. Timing and input current waveforms of a a power-factor-corrected flyback converter operating in BCM.](image-url)
As reference 2 explains in detail, the average input current can be expressed as:

\[
I_{in}(\theta) = \frac{T_{on}}{2L_m} V_{AC}^{Peak} \frac{|\sin \theta|}{\left[ 1 + \frac{V_{AC}^{Peak} \cdot |\sin \theta|}{V_R} \right]}
\]

\[
= \frac{i_{Lm}^{Peak}}{2} \frac{|\sin \theta|}{\left[ 1 + \frac{V_{AC}^{Peak} \cdot |\sin \theta|}{V_R} \right]}
\]  \hspace{1cm} (5)

The rms input current is given as:

\[
I_{in}^{rms} = \frac{1}{\pi} \int_{0}^{\pi} I_{in}(\theta) \cdot V_{in}(\theta) \cdot d\theta \quad \frac{V_{in}^{rms}}{V_{in}}
\]

\[
= \frac{1}{\sqrt{2} \cdot \pi} \int_{0}^{\pi} T_{on} \frac{i_{Lm}^{Peak}}{L_m} V_{AC}^{Peak} \frac{(\sin \theta)^2}{\left[ 1 + \frac{V_{AC}^{Peak} \cdot |\sin \theta|}{V_R} \right]} \cdot d\theta
\]

\[
= \frac{I_{Lm}^{Peak}}{\sqrt{2} \cdot \pi} \int_{0}^{\pi} \frac{(\sin \theta)^2}{\left[ 1 + \frac{V_{AC}^{Peak} \cdot |\sin \theta|}{V_R} \right]} d\theta
\]  \hspace{1cm} (6)

Unless the ratio \( R_{VR} = \frac{V_{AC}^{Peak}}{V_{VR}} \) is made very small, it is quite disappointing that the denominator of the above input current equation makes the current shape clearly non-sinusoidal. Fig. 4 shows the input current shape of a BCM flyback with \( R_{VR} \) as a parameter.\(^2\) Harmonic analysis of the input current shape shows that it is difficult to get a THD lower than 10% with \( R_{VR} = 2 \).
During the off-time of the converter switch, the maximum voltage across the MOSFET is the peak input voltage plus the reflected voltage $V_R$. Due to the voltage rating limit of the MOSFET, the feasible values for $R_{VR}$ are only in the range of 1 for U.S. and 2 to 3 for European input voltages. For lighting applications with universal input voltage (90 to 305 V ac), to achieve a relatively low THD, a MOSFET with $BV_{DSS} > 800$ V must be used to allow the ratio $R_{VR}$ to be small enough.

**Hybrid (Partial BCM) Mode**

In order to avoid using a high-voltage MOSFET while still achieving relatively low THD and high efficiency, a hybrid mode that allows the converter to operate partially in BCM mode can be a good choice (Fig. 5.) This proposed hybrid mode reduces the peak and RMS current through the MOSFET switch and transformer. Its zero-voltage switching also helps to reduce the max power losses under worst-case operating conditions; i.e. min ac input voltage and max load.

In addition, when this control scheme is implemented with a transformer that has a higher primary-to-secondary winding turns ratio, the voltage rating required for the output rectifying diode is reduced. This helps to reduce the output diode’s conduction losses and improves the system’s overall efficiency.
(θ₁: The phase angle when the flyback’s operating mode changes from DCM to BCM.)

Over a half cycle of the ac line input, the flyback converter’s initial operating mode is DCM, which switches with constant on-time and fixed switching frequency. As the input voltage rises within the half cycle, the discharge time \( t_{\text{dis}} \) becomes longer and longer (Fig. 5). When the end of secondary diode conduction time is over a switching period set by the preliminary DCM frequency, the operating mode changes from DCM to BCM.

Fig. 5 illustrates this hybrid (partial BCM) mode principle. For the flyback converter to operate in hybrid mode, the controller needs to detect the discharge time \( t_{\text{dis}} \). Fortunately, detecting the inductor current discharge time is one of the many unique features of a PSR controller.\(^1\) For a flyback driver with universal ac input, it is not hard to understand that this hybrid mode will only happen at the low end of the ac input voltage range.

Fig. 6 shows the approximate shape of the input current waveform of a flyback converter in different operating modes. Obviously, in terms of THD performance, the hybrid mode is better than BCM.
The average input current can be represented by the following equation:

\[ I_{in}(\theta) = \frac{T_{on}}{2L_m} V_{AC} \left\{ \begin{array}{ll}
D \cdot |\sin \theta|, \quad \theta_1 < \theta < \theta_2 & \text{or} \quad \theta > \pi - \theta_1 \\
\frac{|\sin \theta|}{\left[ 1 + \frac{V_{AC}^{Peak} \cdot |\sin \theta_1|}{V_R} \right]}, & \theta_1 \leq \theta \leq \pi - \theta_1,
\end{array} \right. \]  

(7)

The rms input current is:

\[ I_{in}^{rms} = \frac{1}{\pi} \int_{0}^{\pi} I_{in}(\theta) \cdot V_{in}(\theta) \cdot d\theta \]

\[ V_{in}^{rms} \]

\[ = \frac{\sqrt{2} \cdot I_{Lm}^{Peak}}{\pi} \left\{ \int_{0}^{\theta_1} D \cdot (\sin \theta)^2 \, d\theta + \int_{\theta_1}^{0.5\pi} \frac{(\sin \theta)^2}{\left[ 1 + \frac{V_{AC}^{Peak} \cdot |\sin \theta_1|}{V_R} \right]} \, d\theta \right\} \]

(8)

\[ = \frac{\sqrt{2} \cdot I_{Lm}^{Peak}}{\pi} \left\{ \int_{0}^{\theta_1} D \cdot (\sin \theta)^2 \, d\theta + \int_{\theta_1}^{0.5\pi} \frac{(\sin \theta)^2}{\left[ 1 + \frac{V_{AC}^{Peak} \cdot |\sin \theta_1|}{V_R} \right]} \, d\theta \right\} \]
Here $\theta_1$ is the angle where the operating mode changes from DCM to BCM. Indeed, the above equation is applicable for all kinds of operating modes. For DCM operation, $\theta_1$ is equal to 0.5$\pi$, so equation 8 is exactly the same as equation 2. For BCM operation, $\theta_1$ is 0, so equation 8 is equivalent to equation 6.

**PSR Principle**

For the PSR flyback LED driver, control of the output current is based on the peak drain current ($I_{pk}$) of the MOSFET switch and the discharge time ($t_{dis}$) of the inductor current. Taking the Fairchild FL7733A controller as an example, Figs. 7 and 8 show the control block diagram of this PSR flyback controller.

In steady state, over the half line-cycle, the driver output current is the same as the average current of the output diode ($I_o$). Hence, the output current can be presented by the equation below.

$$I_{out} = \frac{1}{2 \pi} \int_0^\pi \frac{V_{cs}(\theta)}{R_s} \cdot n_{ps} \cdot t_{dis}(\theta) \cdot f_s(\theta) \cdot d\theta$$

(9)

Here, $R_s$ is the current sensing resistor ($R_{sense}$ as shown in Fig. 1), $n_{ps}$ is the primary-to-secondary turns ratio of the flyback transformer, $V_{cs}(\theta)$ is the sensed peak drain current signal of the MOSFET during the turn-off transient, $t_{dis}(\theta)$ is the discharge time of the inductor current, and $f_s(\theta)$ is the MOSFET switching frequency at the phase angle $\theta$ of the input line voltage. In DCM mode, $f_s(\theta)$ is a fixed switching frequency and can be simply expressed as $f_s$. In BCM operating mode, $f_s(\theta)$ is a function of the phase angle ($\theta$).
In Fig. 7, the functional block called "TrueCurrent Calculation" fulfills the calculation of \( V_{CS}(\theta) * t_{dis}(\theta) * f_s(\theta) \). The signal \( V_{PSR} \), which is proportional to the calculation result, is sent to the error amplifier, which compares it with the voltage reference \( V_{REF} \). The error-amplifier output signal \( V_{COM} \) is compared with a saw-wave signal to control the on-time of the MOSFET switch.

The \( V_{PSR} \) signal can be expressed as

\[
V_{PSR}(\theta) = Gain_{IC} * V_{CS}(\theta) * t_{dis}(\theta)
\]  

(10)

Depending on how the PSR control IC processes the signals \( V_{CS}(\theta), t_{dis}(\theta) \) and \( f_s(\theta) \), the parameter \( Gain_{IC} \) can be a constant or a linear function of \( \{ V_{CS}(\theta) * t_{dis}(\theta) * f_s(\theta) \} \). For a different PSR control IC from a different company, the math model of \( Gain_{IC} \) can be different. In most cases, \( Gain_{IC} \) can be simply modeled with a constant in steady-state analysis.

In steady state, the average of signal \( V_{PSR}(\theta) \) over a half ac line-cycle must be equal to \( V_{ref} \).

\[
\frac{1}{\pi} \int_0^\pi Gain_{IC} * V_{CS}(\theta) * t_{dis}(\theta) * f_s(\theta)d\theta = V_{ref}
\]  

(11)

Based on equations 9 and 11, the output current \( I_{out} \) can be expressed as:

\[
I_{out} = \frac{V_{ref} * \eta_{ps}}{2Gain_{IC} * R_s}.
\]  

(12)

For the Fairchild FL7733A PSR controller, \( V_{ref} = 2.5 \) V and \( Gain_{IC} = 10 \). Therefore, as presented in reference 1, the output current \( I_{out} \) is given as

\[
I_{out} = \frac{1}{9} * \frac{\eta_{ps}}{R_s}.
\]  

(13)

The operating principle of the PSR flyback LED driver now can be simply modeled with the above equation 12. As long as the transformer primary-to-secondary turn ratio \( \eta_{ps} \) and the current sensing resistor \( R_s \) are determined, the driver output current \( I_{out} \) can be precisely controlled. So a practical approach to designing a PSR flyback with PFC starts with the determination of the transformer turn ratio.

However, the determination of \( \eta_{ps} \) can be very complicated. The selection of \( \eta_{ps} \) has to be based on consideration of the following factors:

1. MOSFET voltage rating
2. MOSFET maximum switching duty cycle
3. Flyback converter operating mode
4. Flyback transformer primary-side magnetizing inductance.
5. Power factor and THD requirement.
Overall, the turns ratio $n_{ps}$ is the key factor for the optimum design of a high-power factor and high-efficiency PSR flyback LED driver. The determination of $n_{ps}$ can dramatically affect all other system parameters, and hence the overall electrical performance and cost of the LED driver.

**Design Procedure**

**Transformer Turns Ratio Of The Flyback LED Driver**

The reflected $V_R$, together with the maximum input voltage and the overshoot voltage $V_{OS}$ due to leakage inductance, determines the maximum drain voltage (Fig. 9.) Therefore, a common approach is to start with the reflected $V_R$ to determine the maximum allowable turns ratio based on the MOSFET switch voltage rating. In this way, a relatively large switching duty cycle can be achieved for the best converter efficiency.

![Fig. 9 MOSFET drain-to-source voltage waveform.](image)

With a properly designed RCD snubber circuit, the overshoot voltage $V_{OS}$ can be about half the reflected voltage. Considering the voltage derating factor for MOSFET drain-to-source breakdown voltage (usually 80% of $BV_{DSS}$), we can have,

$$80\% \times BV_{DSS} = 1.5 \times V_R + V_{AC\text{Peak}}^{\text{Peak}}$$  \hspace{1cm} (14)

The max allowed turns ratio $n_{ps}^{\text{max}}$ then can be figured as follows.

$$n_{ps}^{\text{max}} = \frac{80\% \times BV_{DSS} - V_{AC\text{Peak}}^{\text{Peak}}}{1.5 \times (V_{out_{\text{max}}} + V_{d\text{Fwd}}^{\text{Fwd}})}$$  \hspace{1cm} (15)

Equation 15 calculates the maximum allowed transformer turns ratio regardless of DCM or BCM operation.

Next, we need to determine the turns ratio limit for DCM mode. The relation between the rms input current and average output current of an LED driver can be expressed as in equation 16.

$$I_{in}^{rms} = \frac{V_{out} \times I_{out}}{\eta \times V_{AC}^{rms}}$$  \hspace{1cm} (16)
To stay in DCM, the max duty cycle has to meet the condition given by equation 3. So based on equations 2, 3 and 4, we have

\[
I_{Lm}^{\text{Peak}} \geq 2\sqrt{2} \cdot I_{\text{in}}^{\text{rms}} \cdot \frac{V_{AC}^{\text{Peak}} + \eta_{ps} \cdot (V_{out} + V_d^{\text{Fwd}})}{\eta_{ps} \cdot (V_{out} + V_d^{\text{Fwd}})}.
\]  

(17)

The current sensing resistor \( R_S \) can be presented as

\[
R_S = \frac{V_{CS}^{\text{Peak}}}{I_{Lm}^{\text{Peak}}}
\]

(18)

where \( V_{CS}^{\text{Peak}} \) is the peak value of \( V_{cs}(\theta) \) at the peak of the min. ac input voltage.

From equations 12 and 18, we have

\[
I_{Lm}^{\text{Peak}} = \frac{2 \cdot \text{Gain}_{IC} \cdot V_{CS}^{\text{Peak}} \cdot I_{out}}{V_{\text{ref}} \cdot \eta_{ps}}.
\]

(19)

Combining equations 17 and 19 results in

\[
\eta_{ps} \leq \frac{\text{Gain}_{IC} \cdot V_{CS}^{\text{Peak}} \cdot I_{out}}{\sqrt{2} \cdot V_{\text{ref}} \cdot I_{\text{in}}^{\text{rms}}} \cdot \frac{V_{AC}^{\text{Peak}}}{V_{out} + V_d^{\text{Fwd}}}. 
\]

(20)

Inserting equation 16 into equation 20, we have

\[
\eta_{ps} \leq \frac{\text{Gain}_{IC} \cdot V_{CS}^{\text{Peak}} \cdot \eta \cdot V_{AC}^{\text{rms}}}{\sqrt{2} \cdot V_{\text{ref}} \cdot V_{out}} \cdot \frac{V_{AC}^{\text{Peak}}}{V_{out} + V_d^{\text{Fwd}}}. 
\]

(21)

Under the operating condition with minimum ac input voltage and maximum output voltage and current, the switching current reaches the max peak level at the peak of ac input voltage, as does the \( V_{cs} \) signal. Therefore, the turns ratio limit for DCM mode, named here as \( n_{ps \_DCM}^{\text{max}} \), can be determined by the following condition.

\[
n_{ps \_DCM}^{\text{max}} \leq \frac{\text{Gain}_{IC} \cdot V_{CS}^{\text{Peak}} \cdot \eta \cdot V_{AC\text{min}}^{\text{rms}}}{\sqrt{2} \cdot V_{\text{ref}} \cdot V_{out\_max}} \cdot \frac{V_{AC\text{min}}^{\text{Peak}}}{V_{out\_max} + V_d^{\text{Fwd}}}
\]

(22)
Equation 22 gives the maximum transformer turns ratio to design a flyback driver with DCM operation. Meanwhile, equation 15 determines the maximum allowed transformer turns ratio based on the MOSFET voltage rating, regardless of operating mode.

In the case where \( n_{ps}^{max} > n_{ps, DCM}^{max} \), the driver can operate in hybrid mode if a turns ratio higher than \( n_{ps, DCM}^{max} \) is selected. A higher transformer turns ratio \( (> n_{ps, DCM}^{max}) \) with greater transformer magnetizing inductance allows use of a MOSFET switch with a relatively higher duty cycle; hence the peak and RMS current through the MOSFET switch can be minimized, thereby reducing the conduction losses of the MOSFET.

Inductance Calculation

Inductance For DCM Operation

For DCM, the calculation of flyback transformer magnetizing inductance can be quite straightforward. Since the max peak switching current \( I_{Lm, peak} \) of the MOSFET occurs at the peak of the min ac line voltage, the inductance can be determined with the minimum line input voltage and full-load condition:

\[
I_{Lm, peak} = \frac{t_{on} \cdot V_{AC, peak}}{L_m} \quad (23)
\]

To stay in DCM, the max duty cycle has to be\(^1\)

\[
D \leq \frac{V_R}{V_{AC, peak} + V_R} \quad (24)
\]

where

\[
V_R = \eta_{ps, DCM}^{max} \cdot (V_{out, max} + V_d^{Fwd}) \quad (25)
\]

Therefore, we have

\[
I_{Lm, peak} \leq \frac{V_{AC, peak} \cdot \eta_{ps, DCM}^{max} \cdot (V_{out} + V_d^{Fwd})}{L_m \cdot f_s \cdot [V_{AC, peak} + \eta_{ps, DCM}^{max} \cdot (V_{out} + V_d^{Fwd})]} \quad (26)
\]

With equations 16, 17 and 26, the max allowed transformer primary magnetizing inductance for DCM can be determined as shown in equation 27.
Min Inductance Requirement

The min inductance is determined by the total output power of the flyback LED driver with the min ac input voltage. It is not hard to understand that the flyback driver will operate under DCM with the min inductance.

Once the turns ratio is chosen (equation 15), the current sensing resistor $R_s$ is then determined (equation 13.)

Since the maximum allowed $V_{cs}$ voltage is limited by the controller IC specification[3], the maximum allowed peak current through the transformer primary side is limited. Therefore, to fulfill the LED load requirement, the minimum transformer primary magnetizing inductance needs to be big enough to transfer the required energy to the output.

For the flyback driver operating in DCM, we have

$$L_m \leq \frac{\eta}{4V_{out} * I_{out} * f_s} \left[ \frac{V_{Peak}^{ACmin} * \eta_{PS \_DCM}^{max} * (V_{out} + V_{d \_Fwd})}{V_{Peak}^{ACmin} + \eta_{PS \_DCM}^{max} * (V_{out} + V_{d \_Fwd})} \right]^{2} \quad (27)$$

Max Inductance

Usually, a flyback transformer with a large primary magnetizing inductance is preferred from an efficiency point of view. A large inductance can set the flyback driver to operate in hybrid mode at low ac input voltage; its operating duty cycle can be maximized and therefore the peak and rms switching current can be reduced with only slight sacrifice of THD performance. Therefore, the overall system efficiency can be improved.

When the flyback driver operates in hybrid mode, the maximum turns ratio of the transformer is limited by the voltage stress on the switching MOSFET (equation 15), and the minimum is limited by the voltage stress on the output diode. From equations 12 and 18, we have

$$I_{L \_m \_Peak} = \frac{2 \text{Gain}_{IC} * I_{out} * V_{ CS \_Peak}^{Peak}}{V_{ref} * \eta_{PS}^{max}}$$
Combining with equation 23, we have

\[
L_m = \frac{V_{ref}}{2Gain_{IC}} \cdot \frac{t_{on} \cdot V_{ACmin} \cdot \eta_{ps}^{max}}{I_{out} \cdot V_{CS}^{Peak}}
\]  \hspace{1cm} (29)

\[
t_{on} \leq \frac{D_{max}}{f_s} .
\]  \hspace{1cm} (30)

For a PSR flyback driver running in hybrid mode, its max operating duty cycle \(D_{max}\) can be determined based on when it enters into BCM mode (Fig. 5.)

\[
D_{max} \leq \frac{V_R}{V_{ACmin}^{Peak} \cdot \sin \theta_1 + V_R}
\]  \hspace{1cm} (31)

Therefore, the max inductance finally can be determined as

\[
L_m \leq \frac{V_{ref}}{2Gain_{IC}} \cdot \frac{V_{ACmin}^{Peak} \cdot \eta_{ps}^{max}}{I_{out} \cdot V_{CS}^{Peak} \cdot f_s} \cdot \frac{V_R}{V_{ACmin}^{Peak} \cdot \sin \theta_1 + V_R}
\]  \hspace{1cm} (32)

Targeting \(\theta_1 = \pi/4\) is a good choice to design a flyback driver that operates in hybrid mode at min ac input voltage. This will improve the flyback efficiency at low ac input voltage; while still maintaining the PF and THD within a desirable performance range.

**Summarized Design Procedure**

A careful review of the above equations leads to the following key design guidance for the PSR flyback with PFC for LED lighting applications.

1. Based on the ac input specification, choose the MOSFET with the most suitable voltage rating, i.e., a 600-V or 800-V MOSFET.
2. For pure DCM operation, based on the MOSFET voltage rating, use equation 22 to determine the proper transformer primary-to-secondary winding turns ratio. If the delivered turns ratio is too small (empirically determined to be < 1.5), then a MOSFET with higher voltage rating is usually preferred from the efficiency viewpoint.
3. For hybrid-mode operation, use equation 15 to determine the proper transformer primary-to-secondary winding turns ratio.
4. Once the transformer turns ratio is determined. Use equations 27 and 28 to choose the transformer primary magnetizing inductance if DCM operation is preferred. Use equation 32 to determine the inductance hybrid mode operation is desired for higher efficiency.
Design Example

To verify the proposed design procedure, as well as the higher efficiency of the proposed hybrid-mode operation versus DCM operation, a quick design example for a single-stage flyback LED driver with power factor correction is presented below.

A. Design Specification
   Mains voltage range: V\text{ac} = 90 V to 305 V rms
   Mains frequency: 60 Hz
   Max dc output voltage V_{\text{out}} = 25 V
   Output constant current: I_{\text{out}} = 0.7 A

B. Pre-Design Choices
   Controller: FL7733A
   Switching frequency: 70 kHz
   Selected peak voltage of current sensor V_{\text{CS}}^{\text{Peak}}: 0.85 V
   MOSFET voltage rating: FQP6N80C (800 V/5.5 A)
   Voltage derating rule: 80%

With the above choices, from equations 15, 22, 27, 28 and 32, we can plot a graph as shown below in Fig. 10. The graph can be used for guidance in choosing the transformer primary magnetizing inductance versus the turns ratio, either for DCM or hybrid-mode operation.

Fig. 10. Primary magnetizing inductance selection vs. turns ratio.

To compare the performance of DCM and hybrid-mode operation, two transformers with different parameters were designed for the flyback LED driver. One transformer design is strictly for DCM operation across the whole input voltage range, while the other transformer design sets the driver to operate in hybrid mode when it operates at the min input voltage (Fig. 11.)
Transformer one: (for DCM operation only)

- Primary winding: 51 turns, 28 AWG wire
- Secondary winding: 19 turns, 4 X 28 AWG wire
- Core: PC40RM8Z from TDK
- Primary winding inductance: 460 µH
- With the selected turns ratio: A 300-V/3-A fast recovery diode is used for output rectifier (EGP30F).

Transformer two: (for hybrid operation)

- Primary winding: 65 turns, 28 AWG wire
- Secondary winding: 19 turns, 4 X 28 AWG wire
- Core: PC40RM8Z from TDK
- Primary winding inductance: 720 µH
- With the selected turns ratio: A 200-V/3-A schottky diode is used for output rectifier (S320).
The table below compares the losses and the efficiency of the LED driver with these two different transformers. Compared with DCM operation, the hybrid mode of operation clearly achieves much higher efficiency.

Table. Comparison of flyback losses and efficiency under DCM and hybrid-mode operation.

<table>
<thead>
<tr>
<th>Loss analysis (at Vin = 90 V)</th>
<th>Transformer design 1 (DCM)</th>
<th>Transformer design 2 (Hybrid)</th>
</tr>
</thead>
<tbody>
<tr>
<td>MOSFET</td>
<td>1.058 W</td>
<td>0.83 W</td>
</tr>
<tr>
<td>Output diode</td>
<td>0.798 W</td>
<td>0.588 W</td>
</tr>
<tr>
<td>Transformer</td>
<td>0.674 W</td>
<td>0.642 W</td>
</tr>
<tr>
<td>Bridge rectifier</td>
<td>0.409 W</td>
<td>0.409 W</td>
</tr>
<tr>
<td>Snubber</td>
<td>0.535 W</td>
<td>0.483 W</td>
</tr>
<tr>
<td>Others</td>
<td>0.151 W</td>
<td>0.151 W</td>
</tr>
<tr>
<td>Total loss</td>
<td>3.625 W</td>
<td>3.103 W</td>
</tr>
<tr>
<td>Driver Efficiency</td>
<td>82.61%</td>
<td>84.73%</td>
</tr>
</tbody>
</table>

**Conclusion**

The principle of a primary-side-regulated flyback converter with power factor correction for the LED lighting application has been explained in detail. A straightforward design procedure has been presented as well. A simulation of a design example also verified that the proposed hybrid mode of operation (partial BCM) achieves superior driver efficiency in the LED application compared to the traditional DCM operation of the flyback. At the same time, the hybrid mode keeps PF and THD within the desirable range.

**References**


**About The Authors**

Roy Mi is an applications engineer for Fairchild Semiconductor’s technical marketing team based in Chicago, Ill. He joined Fairchild after 20+ years working in various R&D positions in the power electronics industry. Roy graduated from National University of Singapore with an MSEE in power electronics.
John Jing is an applications engineer for Fairchild Semiconductor, focusing on web-based online design tools for various topologies. He is based in Los Angeles, Calif. Prior to Fairchild, he worked as an applications engineer at Semtech. Before joining Semtech, he worked at Rockwell Science Center as a research scientist in the power management group for dual-purpose projects. John holds engineering degrees from Huazhong University of Science and Technology and the Virginia Polytechnic Institute and State University.

Jason Guo is a Fairchild Semiconductor senior applications engineer focusing on high-performance power conversion, web-based design tools, and tier-one customer support. He is based in San Jose, Calif. Prior to joining Fairchild, Jason worked at Semtech and Artesyn Technologies. Jason received an MSEE in power electronics from Virginia Tech.

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