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Design Guide for Off-line Fixed Frequency DCM Flyback Converter

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DCM Flyback

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I. Introduction

Flyback is the most widely used SMPS topology for low power application from 100W down to under 1W, whenever the output needs to be isolated from the input. Its best features are low system cost, simplicity, and relative ease of implementation. For low current output and power levels below 50W, DCM flyback is the usually the preferred operating mode, due to it's simpler control loop implementation and lower turn on loss. The objective of this paper is to develop a comprehensive, practical and easy to follow approach in designing an off line DCM Flyback power supply. This includes component selection guide, design knowledge and practical tips for a fast and well optimized design.

II. Fixes Frequency Flyback Modes of Operation: DCM vs CCM

Figure 1 shows the basic circuit diagram of a Flyback converter. Its main parts are the transformer, the primary switching MOSFET Q1, secondary rectifier D1, output capacitor C1 and the PWM controller IC. Depending on the design of T1, the Flyback can operate either in CCM (Continuous Conduction Mode) or DCM (Discontinuous Conduction Mode).

In DCM, all the energy stored in the core is delivered to the secondary during the turn off phase (Flyback period), and the primary current falls back to zero before the Q1 switch turns on again. For CCM, the energy stored in the transformer is not completely transferred to the secondary; that is, the Flyback current (ILPK and ISEC) does not reach zero before the next switching cycle. Figure 2 shows the difference between DCM and DCM mode in terms of Flyback primary and secondary current waveforms.

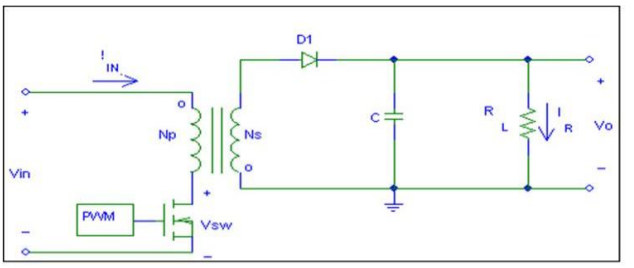

Figure 1: Flyback Schematic

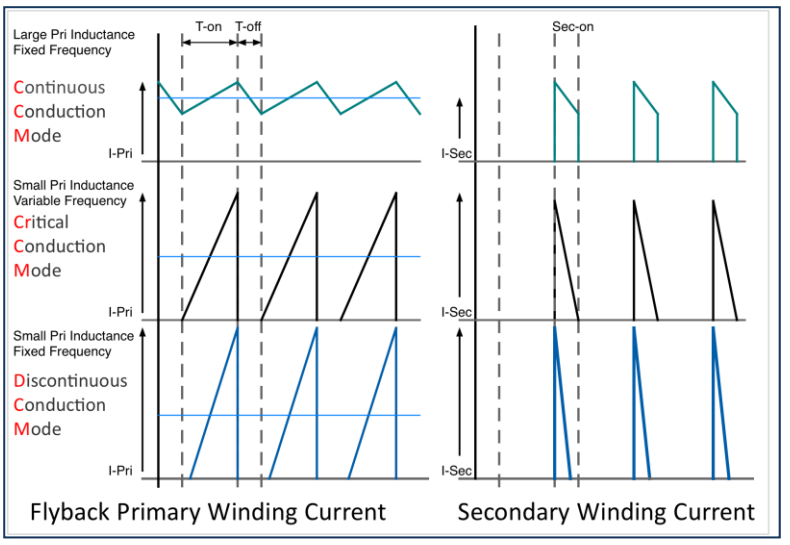

Figure 2: CCM and DCM Flyback Current Waveforms

Table 1: DCM vs CCM

Application Advantage	DCM	CCM
Smaller transformer * (not considering efficiency)	✔	
Faster Transient Response	✔	
Ease of Feedback Loop and Current Loop compensation	✔	
Zero Reverse Recovery Loss on Rectifier Diode and Low Turn-on Loss for Flyback Switch	✔	
Lower Primary and Sec RMS Current factor		✔
Smaller Output Capacitor, lower ripple current		✔
Cross regulation for multiple outputs		✔
Peak MOSFET and Diode current		✔
RMS Loss in transformer windings		✔
Flux ripple excursion in transformer core		✔

Table 1 highlights the main points of advantage for either DCM or CCM mode operation. DCM operation requires a higher peak currents to deliver the required output power compared to CCM operation. This translates to a higher RMS current rating on the primary MOSFET and ouput capacitor, and greater conduction losses in the transformer windings. When these higher peak and RMS current limits the fulfillment of design requirements, (e.g. larger output capacitor required or very high conduction loss on the MOSFET and transformer), switching to CCM mode is advised. This condition usually occurs for designs wherein the output voltage is low and output current is relatively high (> 6A), typically for output power over 50W, though for 5V or lower outputs this is a problem at power below 50W.

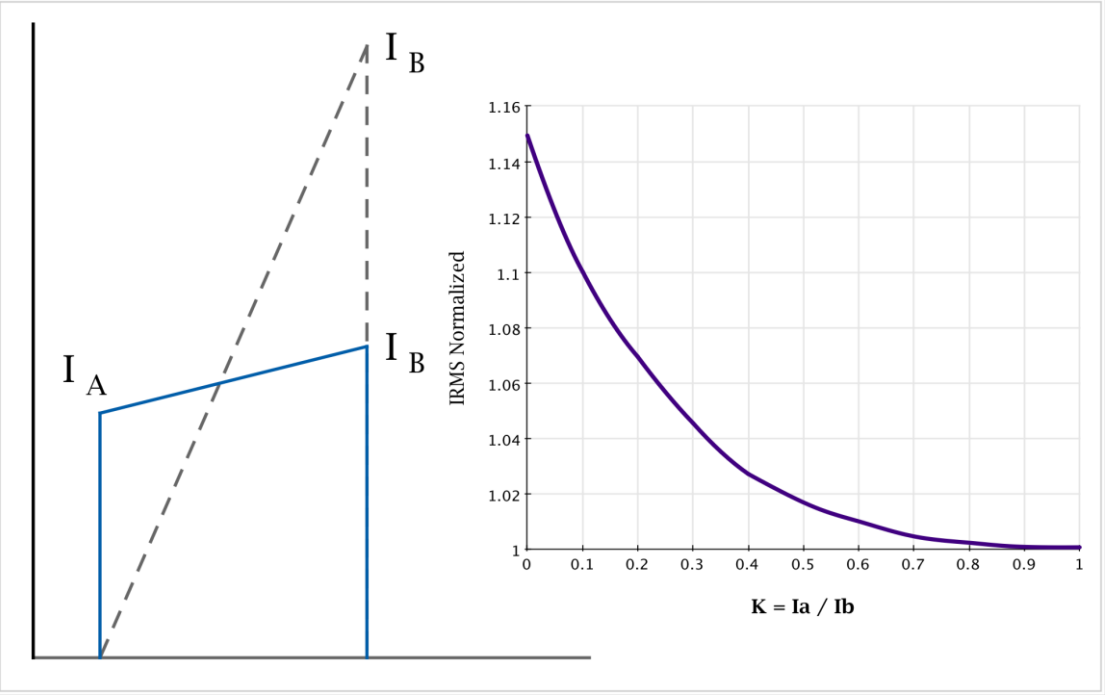

Figure 2B: Normalized RMS Current Ratio for Trapezoidal/Triangular Waveforms

Figure 2B shows the relationship between trapezoidal and triangular current waveforms on the primary side of the Flyback. I_A is the nominal starting point of the waveform, which will be zero for a triangular waveform and some higher value determined by the current still flowing in the primary winding during CCM operation when the switch turns back on. I_B is the end point for the current level during the Ton interval. The IRMS normalized current value as a function of the K factor (I_A/I_B) is shown on the Y axis; this is the multiplier that should be used for estimating resistive losses for different wave shapes in comparison to a flat top trapezoidal waveform, and highlights the additional DC conduction losses inherent to the transformer windings and semiconductors as a function of the current waveform. This can give an 8-12% conduction loss advantage to a well designed CCM converter; this is something to consider in applications where higher RMS currents are required, and also if optimizing efficiency is a key goal. The additional copper losses can be overcome, but that in itself may require a larger core to accommodate an increased winding window, compared with core requirements alone.

Because the DCM mode may allow a smaller transformer and provide fast transient response and lower turn-on losses, it is the usually the best choice for lower power or a Flyback with a high output voltage and low output current requirement. While a Flyback converter can be designed to operate in both modes, it is important to take note that when the system shifts from DCM to CCM operation, its transfer function is changed to a two pole system with low output impedance; thus, additional design rules have to be taken into account including different loop and slope compensation for the inner current loop. In practice, what this means is designing for CCM and allowing the converter to work in DCM at light loads.

Also, more advanced transformer techniques make it possible to extend CCM mode and clean light load regulation and high cross regulation over a wide load range by using a stepped gap transformer. In this type of design, a small portion of the core gap is very small or bypassed with solid material to provide high initial inductance, and CCM mode operation at light load. A discussion of the stepped gap transformer technique is outside the scope of this paper.

Given these characteristics of DCM mode, it is the preferred choice for a simple, easy to design for low power SMPS. The following is a step-by-step design guide on designing a DCM operation Flyback converter.

III. DCM Flyback Design Equations and Sequential Decision Requirements

STEP 1: Define and determine system requirements: Every SMPS design starts in determining the system requirements and specifications. The following parameters need to be defined and determined.

Parameters	Descriptions
VACmax	Maximum AC input voltage
VACmin	Minimum AC input Voltage
fsw	Switching Frequency
Eff	Efficiency
Pout	Output Power (maximum)
Vout	Main Output voltage
fline	AC Line frequency
ΔVout	Output ripple voltage
Standby Power, Light load efficiency, Various Protections, EMI	Other Requirements

Table 2: Input specifications and system requirements

Efficiency is required to calculate the maximum input power. If no previous requirement is set, use 75%-80% as a reasonable target for low cost Flyback. Choosing Fsw is usually a tradeoff between the transformer size and switching loss, as well as EMI concerns and keeping the first fundamental below 150 kHz. Common practice is between 50kHz-100kHz. Additionally, if there is more than one output voltage involved, Pout max should be the sum of each individual output.

Note that while traditionally the controller and power MOSFET switch have been separate components, the wide popularity of this converter type has led to the development of high performance integrated components combining the MOSEFET switch and controller in one package, such as the Infineon CoolSET™ products, which use custom CoolMOS™ transistors that include an integrated depletion mode start up FET for the controller power supply. The CoolSET™ controllers are optimized for DCM applications, and the standalone controller (ICE3BS03LJG) includes a startup cell.

STEP 2: Choose the right controller considering the output power requirement:

Integrated Controller and MOSFET Option	Fsw	Package	MOSFET VDS	MOSFET RDSON	Pout (85V-265V)
ICE3BR0365	65kHz	DIP7, DIP8,DSO16	650V	6.4 Ohm	<10W
ICE3BR4765	65kHz	DIP7, DIP8,DSO16	650V	4.7 Ohm	5W-15W
ICE3BR1765	65kHz	DIP7,DIP8	650V	1.7 Ohm	15W-30W
ICE3BR1065J	65kHz	DIP8	650V	0.92 Ohm	20W-40W
ICE3BR0665	65kHz	DIP7,DIP8	650V	0.65 hm	25W-45W
ICE3AR0680	100kHz	DIP7	800V	0.65Ohm	30W-50W
Controller IC	**Fsw**		**Reccomended MOSFET**		
ICE3BS03LJG	65/100/130kHz	DIP8	IPA65R600E6 650V/0.6Ohm/TO220FP		30W to above 50W

Table 3: Infineon FF Flyback Controller

Choosing a Flyback controller usually depends on specific application and other design consideration such as cost, design form factor and ease of design. Other requirements such as standby power and protection features are easily be met by choosing the right controller. Table 3 shows a selection guide using Infineon's solution for fixed frequency Flyback with regards to its maximum output power. An integrated solution offers low parts count and easier implementation while a separate controller and MOSFET approach has more flexibility, especially on operating frequency and thermal design

Determining Input Capacitor Cin and the DC input voltage range: The capacitor Cin is also known as the DC link capacitor, depending on the input voltage and input power, the rule of the thumb for choosing Cin is shown below.

Input Voltage	Capacitance	Working Voltage
115Vac	2uF/W	~200V
230Vac	1uF/W	~400V
85Vac-265Vac	2-3uF/W	>400V

Table 4: Recommended Cin per Watt of Input

For wide range operation use a DC link capacitor more than 2uF per watt of input power so as to get a better quality of DC input voltage. With the input capacitor chosen the minimum DC input voltage (DC link capacitor voltage) is obtained by:

$$VDC_{min} = \sqrt{2 \times Vac^2 - \frac{Pin_{max} \times (1 - d_{charge})}{Cin \times fline}} \quad (1)$$

Where: dcharge is the DC link capacitor duty ratio, typically around 0.2.

Figure 3 shows the DC link capacitor voltage. The minimum DC input voltage occur at maximum output power and minim AC input voltage while the maximum DCinput voltage occurs at minimum input power (no load) and maximum AC input voltage. The maximum DC input voltage can be found during no load condition when the capacitor peak charge to the peak of the AC input voltage and is given by;

$$VDCmax = VACmax * \sqrt{2} \quad (2)$$

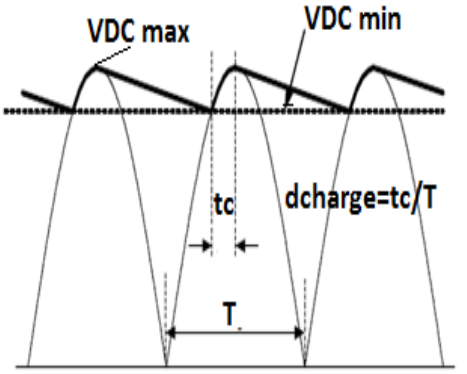

Figure 3: DC link capacitor max and min voltage

STEP 3: Decide on the Flyback reflected voltage (VR) and the maximum VDS MOSFET voltage stress: The reflected voltage VR is the volatage across the primary winding when the switch Q1 is turned off. This also affects the maximum VDS rating of Q1.The maximum drain to source voltage is given by:

$$VDSmax = VDCmax + VR + Vspike \quad (3)$$

Where: Vspike is the voltage spike caused by the leakage inductance of the transformer. For a starting point assume Vspike is 30% of VDSmax. The table below list a recommended reflected voltage given a 650V and 800V rated MOSFET. As a starting point limit VR below 100V for a wide range input voltage.

	VDS max	VR
Wide range AC Input: 85Vac-264Vac	650V	60V-100V
Input from High voltage DC : 400VDC	800V	80V-200V

Table 5: Recommended VR for 650V and 850V MOSFET

Choosing the VR is a compromise between the primary MOSFET and the secondary rectifier voltage stress. Setting it too high, by means of higher turns ratio, would mean higher VDSmax but lower voltage stress on the secondary diode. While setting it too low, by lower turn ratio, would lower VDSmax but would increase secondary diode stress, considering the required adjustments to the transformer turns ratio. Apart from cost, a higher primary VDSmax will mean lower diode stress and lower primary side current stress.

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Determine Dmax based on Vreflected and Vinmin: The maximum duty cycle will appear during VDCmin, at this condition we will design the transformer to be at the boundary of DCM and CCM. Duty cycle here is given by;

$$Dmax = \frac{VR}{VR+VDCmin} \quad \text{(Equation 4)}$$

STEP 4: Calculate primary inductance and primary peak current: The primary peak current can be found by using Equation 5 and 6.

$$Pin_{max} = \frac{Pout_{max}}{n} \quad \textit{(Equation 5)} \qquad Ip = \frac{2 \times Pin_{max}}{VDCmin \times Dmax} \quad \textbf{\textit{(Equation 6)}}$$

The primary inductance should then be design within the limit of maximum duty cycle;

$$Lpri_{max} = \frac{VDCmin \times Dmax}{Ipri \times fsw} \quad \text{(Equation 7)}$$

In order to ensure that the Flyback would not enter into CCM operation at any loading condition make sure to consider the maximum power in calculating Poutmax in Equation 5. Increasing inductance beyond the calculated Lprimax can also push the converter towards CCM mode.

DCM Flyback Transformer Design: Steps 7- Step 12

STEP 5: Choosing the proper core type and size: Choosing the core type and geometry for the first time is quite difficult and usually involve a lot of factors and variables to consider. Among these variables to consider are the core geometry (e.g. EE core/RM core/PQ core etc.) , core size (eg. EE19, RM8 PQ20 etc.) and core material (eg.3C96. TP4, 3F3 etc.).

If there is no previous reference on choosing the right core size, a good way to start is to refer to the manufacturer's core selection guide. Below are some commonly used core size for 65kHz DCM Flyback with respect to the output power.

Even if you choose to work with a transformer vendor such as Wurth/Midcom for the ultimate design and construction to applicable safety standards, working through a design example is a good basis for submittal to a transformer vendor as you will want to establish turns ratio and other parameters for your supplier.

Pout Range	Core Area, Ae (mm2)	Core Size
0-10W	5	EE8.8/4.1/2
	10.1	EE13/6/3
	12.4	EE13/7/4
	20.1	EE16/8/5
10-15W	22.6	EE19/8/5
15-30W	32	EE/20/10/6
30-50W	52	EE25/13/7

Table 6: Recommended core size for DCM Flyback (Fsw=65kHz)

After selecting core size, the right bobbin can also be chosen on the corresponding core's data sheet. Choose the bobbin considering the number of pins, through-hole or surface mount and horizontal or vertical orientation. Core materials are chosen considering the frequency of operation, magnetic flux density, and considering core losses. Core material name varies depending on the core manufacturer, a suitable core materials to start with are 3F3, 3C96 or TP4A.

Determining minimum primary turns: The minimum number of turns on the primary is a function of the magnetic core area and the allowed operating flux density for the chosen material.

$$Np = \frac{Lpri \times Ipri}{Bmax \times Ae} \quad \textbf{\textit{(Equation 8)}}$$

Where Bmax is the operating maximum flux density, Lp is the primary inductance, Ip Is the primary peak current and Ae is the cross sectional area of the chosen core type.

It is important that operating Bmax should not exceed the saturating flux density (Bsat) given on the core's data sheet. Bsat of ferrite core varies depending on the core material and temperature but most of them has a Bsat rating closed to 400mT. If there is no further reference data used Bmax= 300mT. Higher Bmax allows for lower number of primary turns for lower conduction loss but with higher core loss. For optimized design the sum of both the core loss and the copper loss should be mutually minimized. This usually happened near the point where core loss is equal to the copper loss.

STEP 6: Determine the number of turns for the secondary main output (Ns) and other auxiliary turns (Naux):

To get the secondary turns first determine the turns ratio, n

$$n = \frac{Np}{Ns} \quad \textit{(Equation 9)} \qquad n = \frac{VR}{Vout + VD} \quad \textbf{(Equation 10)}$$

Where: Np and Ns are the primary and secondary turns respectively, Vout is the output voltage and VD is the secondary diode voltage drop, typically 0.5V for schottky diode at low to moderate current.
For additional number turns such as auxiliary winding for VCC supply the number of turns can be calculated as follows; Where Vaux is the flyback auxiliary winding , VDaux is the diode voltage drop on this winding.

$$Naux = \frac{Vaux+VDaux}{Vout+VD} \quad \text{(Equation 11)}$$

Most Flyback controllers need an auxiliary winding to supply the IC; this is true of all CoolSET™ types, too. Use the start up VCC supply, as indicated on the data sheet, to decide the auxiliary number of turns. For non integer number of turn round off to the next highest integer

STEP 7: Determining the wire size for each output windings: In order to determine the required wire size the RMS current for each winding should be determined.

Primary winding RMS current: $Ip_{RMS} = Ip \times \sqrt{\frac{Dmax}{3}}$ ***(Equation 11)***

Secondary Winding RMS current:

(Equation 12) $Isec_{pk} = Ip \times \frac{Np}{Ns}$

(Equation 13) $Isec_{RMS} = Isec_{pk} \times \sqrt{\frac{1 - Dmax}{3}}$

: A current density between 150 - 400 circular mil per Ampere can be used as a starting point to calculate the required wire gauge. Below is the quick selection for choosing the appropriate wire gauge using 200CM/A, given the winding's RMS current. The wire diameter with basic insulation for different magnet wire gauges are also shown.

Gauge	Dia (mm) Basic	Dia (mm) Reinforced	RMS Current
34	0.262	0.465	0.2
32	0.305	0.508	0.3
30	0.356	0.559	0.5
29	0.389	0.592	0.65
26	0.584	0.709	1.3
24	0.716	0.815	1.9
22	0.744	0.947	3.1

Table 6: Recommended Wire Gauge by RMS current

STEP 8: Transformer Construction and Winding Design Iteration:
Once transformer parameters have been decided, determine whether the number of turns and the wire size chosen would fit in the given transformer core size. This step may require several iteration of between the chosen core, winding gauge and number of turns.

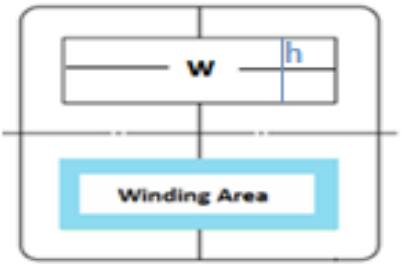

Figure 4: Ferrite core winding area

Figure 4 shows the winding area for an EE ferrite core, using the wire diameter and the number of turns for each winding, we can approximate if the desired winding will fit given its winding area (w and h). If winding will not fit, either the number of turns, wire gauge or core size (controlling window area) will need to be adjusted.
The winding scheme has a considerable influence on the performance and reliability of the transformer. To reduce leakage inductance, the use of a sandwich construction, as shown in Figure 5, is recommended. It also needs to meet international safety requirements. A transformer must have adequate insulation between primary and secondary windings. This can be achieved by using a margin-wound construction (Figure 5A) or by using triple insulated wire for the secondary winding (Figure

5B).

Figure 5: Example of transformer winding scheme (A) using margin tape (B) using triple insulated wire

Using triple insulated wire (reinforced insulation) on the secondary is easier and more preffered way of meeting theis safety requirement. Take note that in fitting triple insulated wire on the choosen core/bobbin, the outside diameter is thicker than the same gauge normal magnet wire.

STEP 9: Design the primary clamp circuit: During turn off, a high voltage spike due to the transormer's leakage inductance appears on MOSFET. This excessive voltage spike on the MOSFET may lead to an avalanche breakdown and eventually failure of the MOSFET. A clamping circuit placed across the primary winding helps to limit the voltage spike caused by this leakage inductance to a safe value.

There are two types of clamping circuit that can be used as shown in Figure 4. These are the RCD clamp and Diode-Zener clamp. The easiest way is to used a Zener clamp ciruit which consist of a diode and high voltage zener or TVS (transient voltage suppressor) diode. The Zener diode effectively clips the voltage spike until the leakage energy is totally dissipated in the Zener diode. The advantage of using this circuit is that it will only clamps whenever the combined VR and Vspike is greater than it's breakdown voltage. At low line and lighter loads where the spike is relatively low, the Zener may not clamp at all, therefore there is no power dissipated in the clamp.

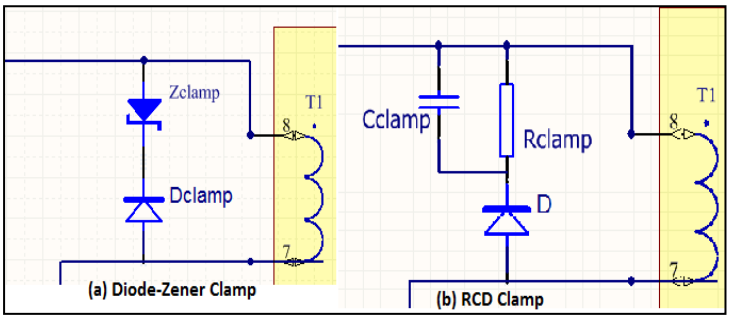

Figure 4: Flyback Primary Clamp Circuit

Choose the Zener/TVS diode rating to be twice the reflected voltage VR. The diode should be ultra fast type with voltage rating greater than the maximum DC link voltage.

The RCD type not only clamps the voltage level but slows down the MOSFET dv/dt. We can used the RCD clamp if passing EMI compliance is an issue without it. The resistor element is crucial in limiting the maximum voltage spike. A lower Rclamp will helps lower Vspike but increases the power dissipation. On the other hand, a higher Rclamp value lower the power dissipation but allows higher Vspike.

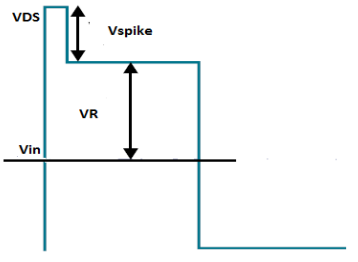

Figure 5: Clamp VDS Voltage

Setting VR= Vspike Rclamp can be determined by:

$$Rclamp = \frac{4 \times VR^2}{Lleak \times Ip^2 \times fsw}$$

Where Lleak is the leakage inductance of the transformer, which can be determined through measurement by shorting the secondary windings. If this is not known, assume Lleak around 2-4% of the primary inductance.

The capacitor Cclamp needs to be large enough to limit the voltage rise while absorbing the leakage energy. Cclamp value may range 100pF- 4.7nF. Rclamp will discharge the capacitor back to the initial value of the switching cycle.

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STEP 10: Output Rectifier Diode Selection:

$$VRVdiode = Vout + VDCmax \times \frac{Ns}{Np} \quad \textit{(Equation 14)} \qquad Isec_{RMS} = Isec_{pk} \times \sqrt{\frac{1 - Dmax}{3}} \quad \textit{(Equation 15)}$$

Chose the output diode so that its VRRM (maximum reverse voltage) is at least 30% higher than VRVdiode and IF (ave forward current) is at least 50% higher than the IsecRMS. Use schottky diode on the main secondary output for lower conduction losses. For DCM mode Flyback peak current is high, and keeping the forward voltage low may require using what seems like a relatively high current diode, depending on the efficiency target.

STEP 11: Output Capacitor Selection: For Flyback converter the proper choice of output capacitor is extremely important . This is because the Flyback mode converters have no inductive energy storage between the rectifier and ouput capacitor. The output capacitor needs to be selected to meet these 3 important parameters: capacitance, ESR (equivalent series resistance) and RMS current rating.

Determining the minimum output capacitance is a function of the allowable maximum peak to peak ouput ripple voltage:

$$Cout_{min} = \frac{Iout_{max} \times Ncp}{fsw \times Vout\_ripple} \quad \textit{(Equation 16)}$$

Where: Ncp is the number of internal clock cycles needed by the control loop to reduce the duty cycle from maximum to minimum value. This usually takes around 10-20 switching periods. Iout is the maximum output current ($\mathrm{Iout} = \mathrm{Poutmax/Vout}$).

The minimum capacitor RMS current rating of the chosen capacitor is:

$$Icap_{RMS} = \sqrt{Isec_{RMS}{}^2 - Iout^2} \quad \textit{(Equation 17)}$$

Given the high switching frequency, the high secondary peak current for a Flyback will produce a corresponding ripple voltage across the output capacitor's equivalent series resistance (ESR). The capacitor must be chosen not to exceed ESRmax or the permissible ripple current capability of the capacitor, which is a thermal limitation for the capacitor. The final selection may more reflect the required voltage rating and ripple current capability, depending on the actual ratio of output voltage and current.

$$ESRmax < \frac{\Delta \mathrm{Vout}}{\mathrm{Isec_{pk}}} \quad \textit{(Equation 18)}$$

Make sure to use the ESR specification from the data sheet at a frequency greater than 1kHz; this will usually be 10 kHz or 100 kHz.

Note: Using a single capacitor with lower ESR value could meet the desired output ripple voltage. A small LC filter is also advisable for higher peak currents specially when operating in DCM mode in order to hit a consistent output ripple voltage performance.

STEP 12: Other Design Considerations

A. Input Diode Bridge Voltage and Current Rating:

$$Iac_{RMS} = \frac{Pin_{max}}{PF * Vacmin}; \qquad VRV > \mathrm{VACmax} * 1.414 \quad \textit{(Equation 19)}$$

Where PF is the power factor of the power supply, use 0.5 if there is no better reference data available. Select the bridge rectifier rating such that the forward current twice than that of IACRMS. A 600V part is commonly use for maximum input voltage of ~400V

B. Current Sense Resistor, Rsense: The sense resistor Rsense is used to defined the maximum output power. Vcsth can be found on the controller's data sheet while Ipmax is the primary peak current while also considering the short term peaks in output power.

$$Rsense = \frac{Vcsth}{Ip(\max)} \quad \textit{(Equation 20)}$$

C. VCC Capacitor: The capacitance value is important for proper startup time, a value of 22uF -47uF is usually in the right range for most applications. If the capacitance is too low, the under voltage lockout of the IC may trigger

before the VCC voltage develops through the converter, while a larger value will slow down the startup time. This capacitor should not be the cheapest type, but must have an adequate ESR and ripple current capability, just like the output capacitor, or it will deteriorate in time. In parallel with the VCC capacitor it is always recommended to use a 100nF ceramic capacitor placed very near to VCC pin and IC ground.

D. Feedback Loop Compensation: Feedback loop compensation is needed to prevent oscillation. For the DCM Flyback, loop compensation is less complicated compared to CCM, as there is no right half plane zero in the power stage to compensate for. A simple RC (Rcomp, Ccomp) as shown in Figure 6 is usually sufficient to make a stable loop. Typical Rcomp values can range from 1k -20k while Ccomp would usually range from 100nF-470nF. A detailed analysis about feedback loop can be found on reference [2].

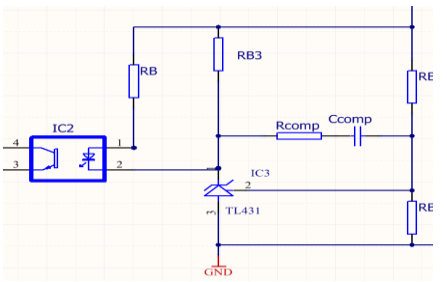


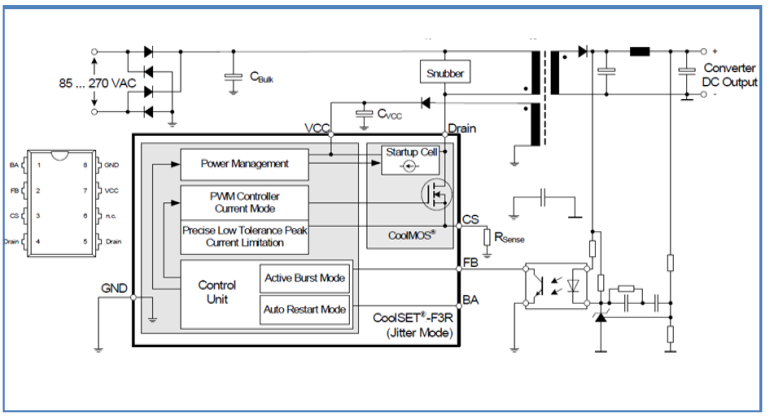
Figure 6: Feedback loop compensation

IV. DCM Flyback Design Example

STEP 1: Sytem Specifications and Requirements:

Parameters	Values	
VACmax	265V	Maximum AC input voltage
VACmin	85V	Minimum AC input Voltage
fsw	65kHz	Switiching Frequency
Eff	80%	Efficiency
Pout	25Wmax	Output Power (maximum)
Vout	12V +/- 2%	Main Output voltage
fline	60Hz	AC Line frequecny
Vout ripple	120mV	Maximum ripple voltage
No load Pin	<50mW	Standby input power

STEP 2: Choosing the right controller considering the Pout: Referring to Table 3, we chooses an integrated controller and MOSFET solution using ICE3BR1065J. Other features include built in startup cell, less than 50mW no load power and frequency jitter and soft driving for lower EMI. Below is the typical Flyback application using ICE3BR1765J



STEP 3: Determining Input Capacitor Cin and the DC input voltage range:

Maximum input power:

$$Pinmax = \frac{Pout}{n} = \frac{25W}{0.8} = 31W$$

Using 2uF per watt of input power, the required DC capacitor, Cin, is:

$$Cin = \frac{2uF}{W} \times 31W = 62uF \sim 68uF$$

> Use the standard capacitance value of **68uF/400V**

With the input capacitor chosen the minimum DC input voltage (DC link capacitor voltage) is obtained by:

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$$VDC_{min} = \sqrt{2 \times 85V^2 - \frac{31W \times (1-0.2)}{68uF \times 60Hz}} = \mathbf{92V} \qquad VDCmax = 265V\sqrt{2} = \mathbf{375V}$$

STEP 4: Flyback reflected voltage (VR) and the Max VDS MOSFET voltage stress: For a 650V MOSFET on ICE3BR0665 CoolSET, VR is chosen at 75V Assuming 30% leakage spike the expected maximum VDS is equal to:

$$VDSmax = VDCmax + VR + 30\%Vspike$$

$$\text{VDSmax} = 375\text{V} + 75\text{V} + 30\% \times 375V = 563V$$

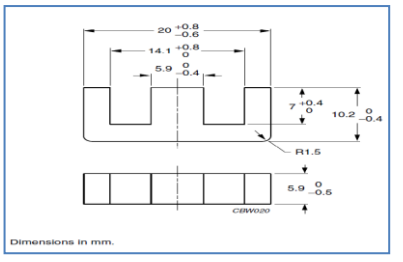
STEP 5: Determining Dmax based on Vreflected and Vinmin:

$$Dmax = \frac{VR}{VR + VDCmin} = \frac{75V}{75V + 92V} = \mathbf{44\%}$$

STEP 6: Calculating primary inductance and primary peak current:

$$Pin_{max} = 31W \quad ; \quad Ip = \frac{2 \times 31W}{92V \times 0.44} = \mathbf{1.53A} \quad ; \quad Lpri_{max} = \frac{92V \times 0.44}{1.53A \times 65kHz} = \mathbf{407uH}$$

STEP 7: Choosing the proper core type and size: Using the table outline on Step 7, we can use EE20/10/6 ferriite core for this 25W power level
Core: EE20/10/6 Ferroxcube/TDK
Cross Sectionl Are, Ae=32mm2
Core Material: 3C96/Ferroxcube, TP4A/TDK
Bobbin:E20/10/6 coil former, 8 pins



STEP 8: Determining minimum primary turns:

$$Np = \frac{407uH \times 1.53A}{300mT \times 32mm^2} = 65Turns$$

STEP 9: Determine the number of turns for the secondary main output (Ns) and other auxiliary turns (Naux):

$$n = \frac{VR}{Vout + VD} = \frac{75V}{12V + 0.5V} = \mathbf{6}$$

$$Ns = \frac{Np}{n} = \frac{65}{6} = 10.83 \sim \mathbf{11Turns}; \qquad \mathbf{Np = 66Turns}$$

Note: Round off non integer secondary value to the next integer value, in this case Ns =11. Using this setup (Np/Ns=65/11) VR is decreased a bit, we can used this value or increase the primary turns to get the same VR as assumed. In this case we will adjust Np to 66 turns to maintain the same VR. Even turns is also desirable on the primary for low leakage split primary winding construction, with the same turns in each section.

An auxiliary winding, Naux, on the primary is needed for the VCC supply. For ICE3BR4765 an internal HV startup is used to supply the initial bias before Vaux kicks in. We set Vaux at 15V to be above 11.2V max turnoff voltage.

$$Naux = \frac{Vaux + VDaux}{Vout + VD} = \frac{15V + 0.5V}{12V + 0.5V} = \mathbf{14Turns}$$

STEP 10: Determining the wire size for each output windings: The RMS current on each winding is calculated using Equation 11-13:

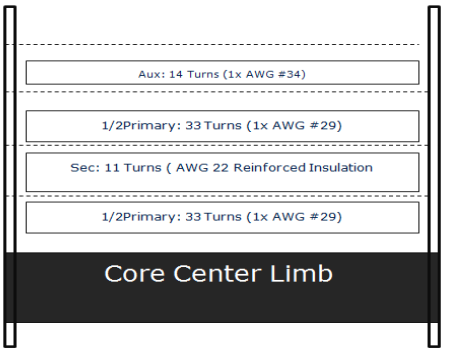
$$Ipri_{RMS} = Ip \times \sqrt{\frac{Dmax}{3}} = 1.53A \times \sqrt{\frac{0.44}{3}} = \mathbf{0.58A}; \quad Isec_{pk} = Ip \times \frac{Ns}{Np} = 1.53A \times \frac{66}{11} = 9A$$

$$Isec_{RMS} = Isec_{pk} \times \sqrt{\frac{1 - Dmax}{3}} = 9 \times \sqrt{\frac{1 - 0.44}{3}} = \mathbf{3.9A}$$

From Table 6 we can use: AWG 29-Primary, AWG22-Secondary, AWG 34-Auxiliary winding*

*Note; The load on the auxiliary winding is usually just the IC bias and MOSFET gate drive (~0.1A)

STEP 11: Transformer Construction and Winding Design Iteration



Using table AWG 29 diameter is 0.389mm, EE 20/10/6 winding area width and height is 14mm and 4mm respectively. This will give us a TPL (turns per layer) on the primary of 35Turns (14/0.389). For our required 66 turn we will need 2 layer on the primary winding.

For the secondary turns we will use triple insulated wire AWG 22, outside diameter is 0.947. TPL fusing this wire is 14 turns (14/0.947) which means a single layer 11 turns would fit in. For the auxiliary basic AWG 34 magnet wire will be used (d=0.262mm).

Checking the winding stacked (0.389mm+0.389mm+0.947mm+0.262) is 2mm which is less than the winding height. This should give us enough margin for the actual build. We can then used a split primary transformer construction as shown in Figure 11.

STEP 12: Design the primary clamp circuit: For the primary clamp we will used a Diode-Zener clamp TVS Diode is P6KE160 and fast recovery diode is BYB27C. An RCD clamp value can also be used as suggested on reference design on [2]

STEP 13: Output Rectifier Diode Selection: Voltage and current rating for the schotky rectifier diodes are:

$$VRVdiode = Vout + VDCmax \times \frac{Ns}{Np} = 12V + 375V * \frac{11}{66} = 74.5V$$

Choose 100V rated diode

$$Isec_{RMS} = 3.9A$$

Choose IF>5.9A

STEP 14: Output Capacitor Selection

$$Cout_{min} = \frac{Iout_{max} \times Ncp}{fsw \times Vout\_ripple} = \frac{2.1 \times 20}{65kHz \times 0.12V} = 270uF$$

Capacitance must be at least equal or higher than this value

For Flyback output capacitance most of the time the limiting factor is the ripple current rating and ESR .

$$Icap_{RMS} = \sqrt{Isec_{RMS}{}^2 - Iout^2} = \sqrt{3.9^2 - 2.08^2} = \mathbf{3.3}\boldsymbol{A}$$

The combined RMS current rating of the output cap(s) used should be higher than this value

$$ESRmax < \frac{\Delta \mathrm{Vout}}{\mathrm{Isec_{pk}}} = \frac{120\mathrm{mV}}{9\mathrm{A}} = 13\mathrm{mohm}$$

Choose a combined paralleled ESR lower that this value

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STEP 15: Other Considerations

A. Input Diode Bridge Voltage and Current Rating

$$Iac_{RMS} = \frac{Pin_{max}}{PF * Vacmin} = \frac{31W}{0.5 * 85V} = 0.73A$$

Chose forward current rating >1.4A

$$VRV > 375V$$

Commonly used is 600V rated bridge rectifier

B. Current Sense Resistor, Rsense

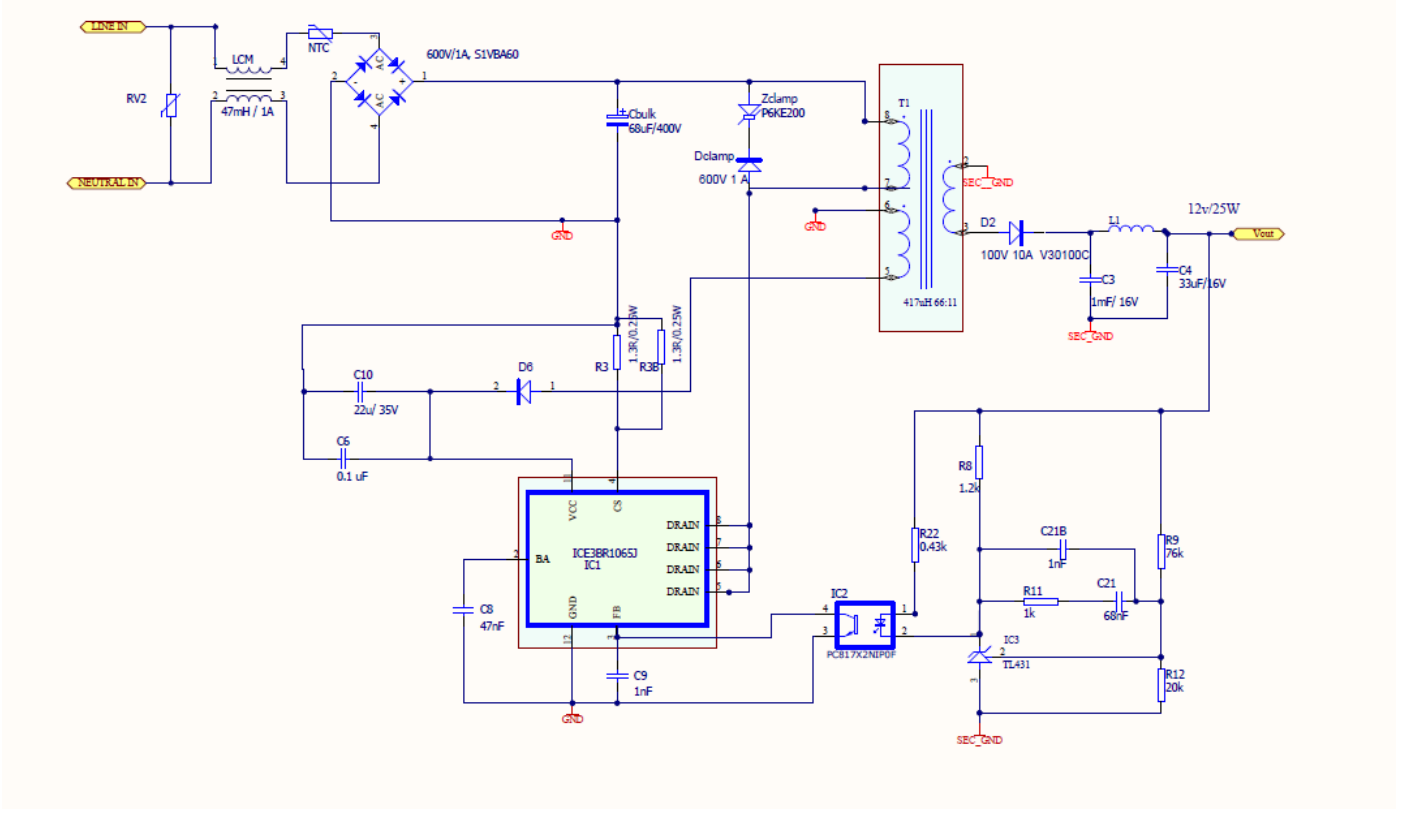
$$Rsense = \frac{Vcsth}{Ip(\max)} = \frac{1V}{1.53A} = 0.65\ \Omega$$

Use 1.3 ohm 0.25W x 2 in parallel

C. VCC Capacitor:

$$Vcc\ capacitor = 22uF/35V$$

D. **Feedback Loop Compensation:** Below shows the complete Flyback schematic including values and implementation of the feedback circuit.



V. References

[1] Switching Power Supply Design by Abraham Pressman

[2] 25W 12V SMPS Evaluation Board with CoolSET ICE3BR1065J

http://www.infineon.com/dgdl/AN_SMPS_ICE2xXXX_V12.pdf?folderId=db3a304412b407950112b418cef926b2&fileId=db3a304412b407950112b418cf5226b3

[3] Infineon Fixed Frequency CoolSET product list.

http://www.infineon.com/cms/en/product/power-management-ics/ac/dc/integrated-power-ics/coolset-tm-f3/channel.html?channel=ff80808112ab681d0112ab6a8b78055a

[4] Infineon Technologies Application Note: “ICE2xXXX for OFF – Line Switch Mode Power Supply (SMPS)”. Feb 2002.

http://www.infineon.com/dgdl/AN_SMPS_ICE2xXXX_V12.pdf?folderId=db3a304412b407950112b418cef926b2&fileId=db3a304412b407950112b418cf5226b3