# 1200 WATT FLYBACK SWITCHING POWER SUPPLY WITH SILICON CARBIDE SEMICONDUCTORS

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**Abstract**: This paper shows how a recent advancement in power semiconductor technology and some improvements can allow the flyback switching power supply to operate effectively at high output powers. The switching frequency can be increased, which allows for a more effective output filter design and a smaller transformer. A basic design example of a 1200 watt power supply power part is presented.

**Keywords**: flyback converter, high frequency, silicon carbide semiconductor, output filter, power supply design

## 1. INTRODUCTION

The flyback converter is a simple switching power supply topology, which, unlike the forward converter, does not require an output inductor. In the flyback converter, the transformer itself works as an inductor at the same time. This makes it possible to construct a power supply with smaller dimensions and even higher efficiency than a forward converter. However, because of some practical limitations (as will be discussed later), the flyback topology is rarely used with output powers over a few hundred Watts – a forward converter is used instead.

With recent advancements in semiconductor technology, it becomes practical to construct a high power flyback converter switching power supply. The focus of this paper is to show how these new semiconductors, especially Silicon Carbide (SiC) MOSFETs and diodes, allow us to realize a simple, high power and efficient flyback power supply.

## 2. SILICON CARBIDE SEMICONDUCTORS IN A FLYBACK CONVERTER

In the simplest variant, the flyback converter uses a single-switch topology. The problem of this topology is that the transistor switch needs to have a high breakdown voltage, because in addition to the input DC-link voltage, it also "sees" the reflected voltage from the secondary winding. This means that for a 230 V~ mains, at least 800 V rated transistor must be used. IGBT transistors are commonly available with a voltage rating of 1200 V and they can switch high currents. However they are too slow for the construction of an efficient flyback converter - so MOSFETs must be used instead. However, MOSFETs with such voltage ratings have high ON-state resistance, which means they would dissipate too much power in a high-power application.

In recent years, a new type of MOSFETs emerged – the Silicon Carbide (SiC) MOSFETs. These transistors combine a high breakdown voltage (usually 1200 V) with low ON-state resistance (much below 1 Ω, usually around 0.1 Ω) and very high switching speeds at the same time. These new transistors are perfectly suited for a construction of a simple and high power flyback converter.

These SiC MOSFETs are made by different manufacturers, but Cree is the biggest one. Take for example the recently released part C2M0025120D in the standard TO-247 package. This part combines a breakdown voltage of 1200 V with ON-state resistance of 0.025 Ω (!) and a continuous drain current of 60 A @ 100 °C. The turn-on and turn-off times are around 50 ns.

## 3. OTHER LIMITATIONS OF A FLYBACK CONVERTER AND HOW TO SOLVE THEM

One of the biggest problems of a flyback power supply is the pulsed output current after the secondary rectifier – see Figure 1.

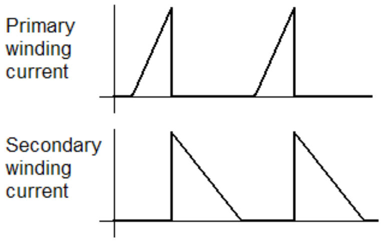


**Figure 1:** Basic flyback converter current waveforms.

This pulsed current puts large stress on the output filter capacitor (large rms current). In comparison, in the forward converter, the output current is continuous with just a small ripple, so a much smaller output capacitor can be used.

However, with the help of fast SiC semiconductors, we can increase the switching frequency greatly above 100 kHz even for high output powers. The increased frequency results in a lower required capacitance of the output filter capacitor (for the same voltage ripple). This then allows us to use film capacitors, which offer only relatively small capacitances, but they withstand much larger currents than their electrolytic counterparts. To further smooth the output voltage, an additional LC filter after the main output capacitor can be used. This filter requires only a very small inductor, which takes up much less space and dissipates much less power than the output inductor for a forward converter. This additional filter also allows us to choose a larger voltage ripple (up to 5-10%) on the main (first) filter capacitor, which further decreases the required capacitance. See Figure 2.

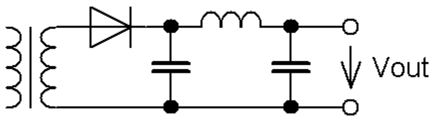


**Figure 2:** A CLC output filter.

A common belief is that the transformer for a flyback converter comes out much larger than for a forward converter for the same output power. However, as will be shown further, with proper design and with the help of a high switching frequency, the transformer can be made relatively small even for high output powers (*ETD4917* core, with top dimensions of 49 x 17 mm, for a 1200-Watt power supply).

## 4. DESIGN EXAMPLE OF A 1.2 KW POWER SUPPLY

For the purpose of verifying the concept, an experimental 1200-Watt power supply has been designed and built.

Selected parameters of the power supply are following:

- DC-link voltage $V_{bus}$ = 300 Vdc (for rectified 230 Vac single-phase mains)
- Output voltage: $V_{out}$ = 60 Vdc
- Output current: $I_{out}$ = 20 A
- Switching frequency: $f_{sw}$ = 160 kHz

## 4.1. POWER TRANSFORMER DESIGN

For the power transformer, the ETD 4917 core was selected. A two-part core must be used, because an air gap in the core is required. This core has the following parameters [1]:

- Core area: $A_e$ = 211 mm$^2$
- Area available for winding: approximately 264 mm$^2$

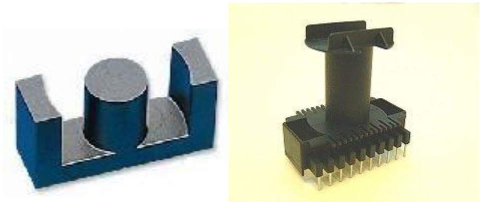

**Figure 3:** ETD core and bobbin.

For such high switching frequency (160 kHz), the flux density ripple $\Delta B$ in a ferrite core must be only around 0.15 T, however a maximum operational flux density $B_{max}$ = 0.3 T still can be used, which means the core will be "biased" into continuous-flux operation. The advantage of this operation is that the rms currents through windings are decreased.

The maximum primary duty cycle is 0.5 (see further for the converter topology), so a duty cycle of 0.33 will be selected to allow some headroom for regulation. Then we can move onto the calculations of the transformer windings.

The number of primary turns will be [2]:

$$N_1 = \frac{V_{bus} \cdot duty}{f_{sw} \cdot \Delta B \cdot A_e} = \frac{300 \cdot 0.33}{150000 \cdot 0.15 \cdot 211 \cdot 10^{-6}} = 20.85 \sim 21 \tag{1}$$

The primary winding rms current will be approximately:

$$I_{1,rms} = \frac{P}{V_{bus}} \cdot \frac{1}{\sqrt{duty}} = \frac{1200}{300} \cdot \frac{1}{\sqrt{0.33}} \cong 7\text{ A} \tag{2}$$

The primary winding wire cross-section, considering 4 A/mm$^2$ current density:

$$S_{Cu1} = \frac{I_{1,rms}}{4} \cdot \frac{7}{4} = 1.75\,\text{mm}^2 \tag{3}$$

The number of secondary turns will be:

$$N_2 = N_1 \cdot \frac{V_{out}}{V_{bus}} \cdot \frac{1-duty}{duty} = 21 \cdot \frac{60}{300} \cdot \frac{1-0.33}{0.33} = 8.53 \sim 9 \tag{4}$$

The secondary winding rms current will be approximately:

$$I_{2,rms} = \frac{I_{out}}{\sqrt{1-duty}} = \frac{20}{\sqrt{1-0.33}} = 24.4\text{ A} \tag{5}$$

The secondary winding wire cross-section, considering 4 A/mm$^2$ current density:

$$S_{Cu2} = \frac{I_{2,rms}}{4} \cdot \frac{24.4}{4} = 6.1\,\text{mm}^2 \tag{6}$$

The total transformer winding copper area will be:

$$A_{Cu,tot} = N_1 \cdot S_{Cu1} + N_2 \cdot S_{Cu2} = 21 \cdot 1.75 + 9 \cdot 6.1 = 91.6\,\text{mm}^2 \tag{7}$$

The area available for winding is approximately 261 mm$^2$, so the winding will fit on the selected core. A HF-Litz wire must be used for the windings because of the high-frequency currents.

### 4.2. PRIMARY TRANSISTOR INVERTER

Because of the leakage inductance between primary and secondary winding, some form of snubber circuit must be used to prevent the inductive spikes from damaging the transistor.

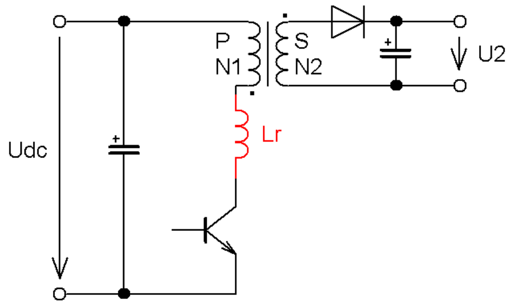


**Figure 4:** Transformer leakage inductance (separated on the primary side).

For a high power inverter, the energy stored in the leakage inductance (see *Figure 4*) and the resulting power dissipation in a standard lossy snubber would be unacceptable. For this reason, we can use a lossless clamping snubber as shown in *Figure 5*. This kind of snubber limits the voltage on the collector of the transistor to $2V_{bus}$, which also means the maximum duty cycle is 0.5 otherwise the transformer would saturate. An additional "reset winding" on the transformer is required, however the current through this winding is small (corresponding only to the energy stored in the leakage inductance).

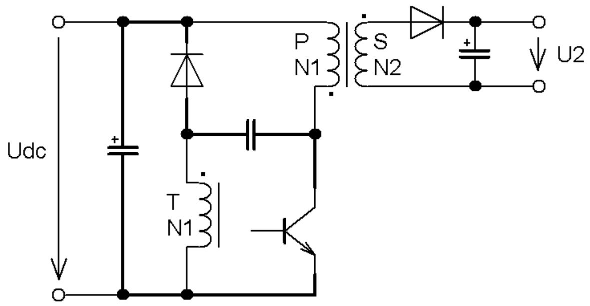


**Figure 5:** Lossless clamping snubber.

The rms current through the transistor switch is the same as the primary winding rms current – 7 A. The *C2M0160120D* Silicon Carbide MOSFET from Cree will suffice. This transistor has an ON-state resistance of 0.16 Ω and a continuous drain current of 12.5 A @ 100 °C.

The free-wheeling diode also "sees" a voltage of 2x $V_{bus}$. A SiC Schottky diode can be used.

### 4.3. SECONDARY FILTER

When the transistor is turned on, no current flows through the secondary winding and all of the output current is taken from the filter capacitor. From this and from the allowable voltage ripple, we can calculate the capacitance value. If we use a CLC filter, we can choose a relatively large voltage ripple on the first capacitor – we choose 5 % of the output voltage. From the basic capacitor differential equation, we can write the following:

$$C_1 = I_{out} \cdot \frac{\frac{1}{f_{sw}} \cdot duty}{V_{ripple}} = 20 \cdot \frac{\frac{1}{160000} \cdot 0.33}{60 \cdot 0.05} = 13.75\ \mu\text{F} \tag{8}$$

We can use for example 3x 4.7 μF film capacitors in parallel.

If the resonant frequency of the following LC filter is 1/10 of the switching frequency, it will attenuate the voltage ripple by 100 times (second order filter). This will result in 0.05 % voltage ripple, which is perfectly acceptable. We choose 2x 4.7 μF film capacitors in parallel for the second capacitor. From the Thompson equation, the series inductance value then must be:

$$L = \frac{1}{4\pi^2 f_r^2 C_2} = \frac{1}{4\pi^2 \cdot 16000^2 \cdot 9.6 \cdot 10^{-6}} = 10.3\ \mu\text{H} \tag{9}$$

This is a small inductance value and can be realized on a small iron-powder core (for the required current).

## 5. CONCLUSION

Using the new Silicon Carbide semiconductor technology, it is possible to realize a high power, simple and physically small power supply using the flyback topology. The price of the SiC parts is still somewhat high, but it is going down rapidly. We can expect a widespread use of the SiC parts in the future.


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