Flyback Switchmode PSU for Powering Legacy Equipment

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Background

Back in the early 1990's I acquired a 10 watt Travelling Wave Tube Amplifier for 10GHz which I intended to use for /P operation. A Switch Mode Power Supply (SMPSU) for generating the high voltages for the TWT from a 12 Volt input had to be built. I rapidly taught myself the basics of SMPSU design and made something that worked. It broke many rules of proper SMPSU architectures; used old ferrite components and semiconductors and was all a bit of a bodge. But it did do the job it was supposed to with the design subsequently being duplicated by several others.

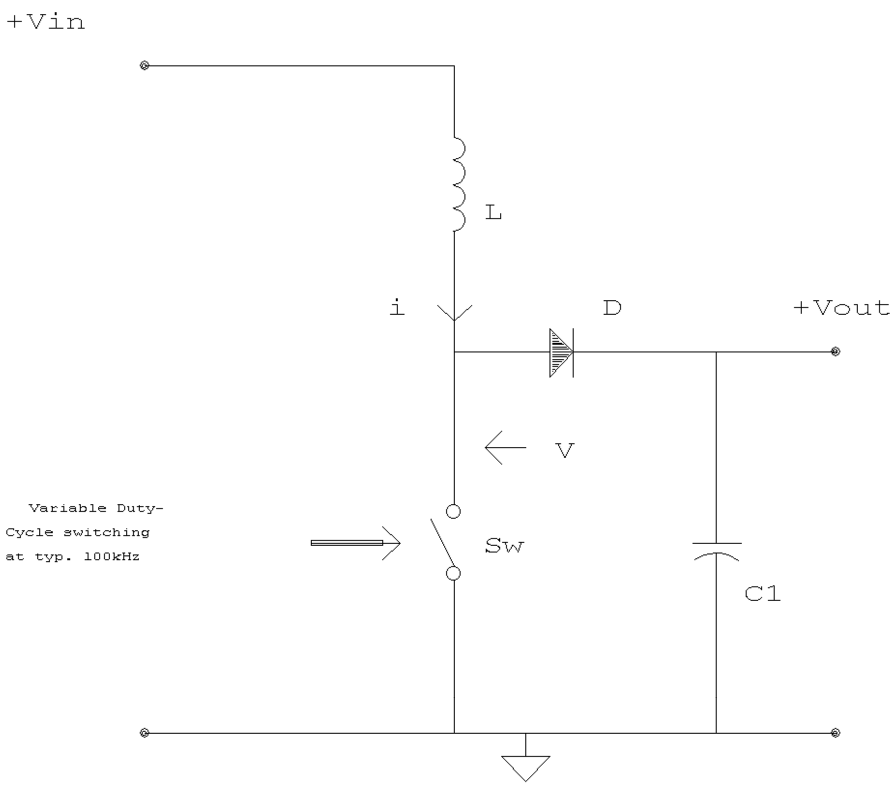
Step forward 20 years. Alan G8LSD approached me to say he wanted to build part of that old TWT PSU to power valve equipment and projects as part of his virtual-valve museum [1] and asked for advice about suitable parts. I immediately cringed at the thought of all the basic mistakes and bodges in that original design, with the now impossible-to-get ferrite cores, and promised him a modern design using readily available up-to-date components. An output voltage around 300V with about 30 Watt rating was considered to be suitable. Here is the result of that promise…

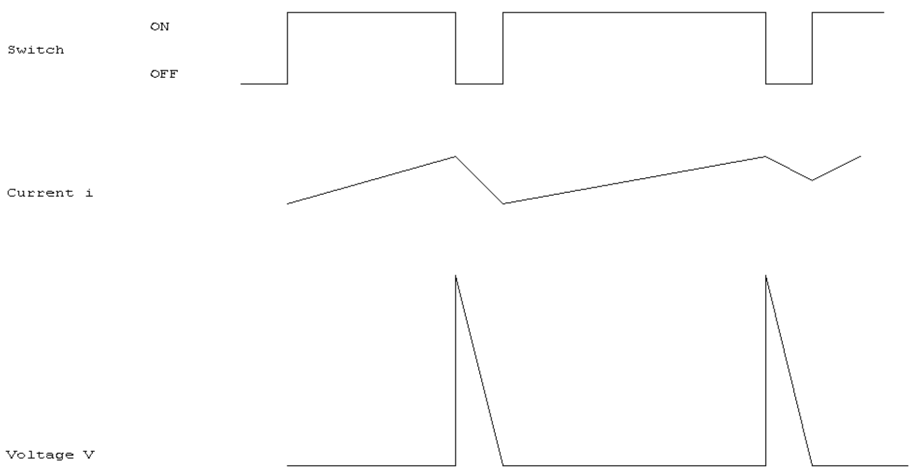
SMPSU Design

All SMPSUs work on the same simple basic principle, although their architectures can be very different depending upon the voltage input range, the wanted outputs and the power throughput. They all work by chopping a DC input voltage using a semiconductor switch, typically operating at around 100kHz, then using an inductor followed by capacitor to smooth out the chopped waveform and give a clean DC output. By varying the mark-to space ratio of the switching waveform the output voltage can be varied at will, and by measuring the value of this and feeding it back to a control loop, the M/S ratio is continuously and automatically tweaked to give a stable output voltage whatever load current is drawn. Dedicated SMPSU controller ICs have been around for decades – long before my original TWT PSU was built – and they do all the complicated stuff involving voltage references, voltage stabilisation / error amplifier, duty cycle control and switch driving.

In most SMPSU designs a transformer between the chopping stage and the output inductor provides voltage transformation. For mains input designs this transformer also gives the vital safety isolation barrier between raw mains input and the user's low voltage equipment. As the transformer is being run at 100kHz, it is considerably smaller and lighter than one would be that had to run at 50Hz. For the typical SMPSU just described, two wound components are needed; the transformer and the output inductor.

For low voltage input, high(er) voltage out, at modest power ratings of a few watts to tens of watts, there is one SMPSU topology that allows the functions of transformer and inductor to be combined into a single unit. This is commonly referred to as the 'flyback' design and has the lowest component count. The outline, with no transformer and just a single inductor is shown in Figure 1 and makes use of the back EMF generated when current though an inductor is suddenly interrupted. When the switch **Sw** is closed, current **i** builds up in the inductor **L**, storing energy in the magnetic field. When **Sw** opens, the magnetic field collapses, dumping the stored energy as a high voltage spike shown as **V** on the diagram. The diode conducts this away into the output capacitor. The term 'flyback' comes from the old Cathode Ray Tube television line output stages where the energy in the line deflection coils had to be suddenly dumped when the scan line retraced. This stored energy was used to generate the high voltage needed for the final anode of the CRT.





The Flyback PSU in Depth (and the maths)

There is no inherent mechanism for controlling the voltage generated by this method. The current though an inductor cannot change abruptly, so when S1 is opened, this current still continues to flow through whatever medium it can.   If there were no other path for the current to pass though, the voltage would just

rise to the point at which it sparks over or breaks down whatever get it its way. This is the principle behind spark ignition, and shows why you should always include a diode across relay coils when switching them with transistors. With D1 and C1 present, the current behaves in a more benign manner and dumps the energy into C1 with the voltage then rising to some more modest value.   But the voltage generated cannot be immediately defined in the way a simple transformer on AC can be, by knowing the turns ratio.   To use the flyback topology to generate a fixed and controllable supply we need to know several things in order to build a feedback stabilisation loop.

Firstly, the energy stored in the inductor during the on time of the switch S1 is given by $\frac{1}{2}L.I^2$ where L is the inductance and I is the current at the point the switch is opened. The instantaneous current though an inductor when a voltage is switched across it (S1 closed) is given by $I = V . t_{on} / L$ where t_{on} is the 'on' time of S1. We also know that the average voltage across any inductor for the combination of both the on and the off periods of S1 <u>must</u> be zero.

If we switch at a constant frequency F, then t_{on} is given by the period of the switching frequency 1/F, multiplied by the duty cycle D. So $t_{on} = D / F$ and $t_{off} = (1\text{-}D) / F$. Since the average voltage across the inductor during both on and off periods together must be zero, we know that the volt-seconds for the on time must equal the volt-seconds for the off time. So $V_{IN} * t_{on}$ must be equal to $V_{OUT} * t_{off}$ and $V_{OUT} = V_{IN} * t_{on} / t_{off}$

If duty cycle, D, is used instead of t_{off} and t_{on}, the equation relating input and output voltage becomes : $V_{OUT} = V_{IN} * D / (1\text{-}D)$. We see that output voltage is determined only from the input supply and D. Not the switching frequency, or the value of the inductor . But that is not the whole story by any means

We now have to consider the energy throughput, or the power rating of the overall SMPSU, and the characteristics of practical ferrite components. During t_{on}, the current in L builds to a maximum defined by the input voltage and the value of L. After a time t_{on} the energy stored is $\frac{1}{2}L.I^2$ joules. A joule is a watt per second, so at a switching frequency F, the total energy transferred is $\frac{1}{2}L.I_{peak}^2.F$ Watts So now we know our maximum power throughput is set by the switching frequency, the inductance and the maximum current we can allow to flow though that inductor. I_{peak} and so L cannot be defined arbitrarily; I and L are related by time as current rises linearly during the t_{on} period. So for a higher peak current, we have to have a correspondingly lower inductance. But we can see that power increases with I^2, so in general terms, a higher power rating will need a lower value of inductor, or a higher switching frequency, or higher V_{IN}. or any combination of these.

A secondary winding on the inductor, with a different number of turns to the primary allows us another degree of freedom in keeping voltages within practical limits. It may be that the wanted V_{OUT} is a lot higher than our switch is happy to have across it. In this case, a secondary winding is added, closely linked to the primary so it shares all the magnetic flux that therefore develops a voltage across it. The secondary voltage is related to that across the primary by the turns ratio. So now if we wanted, say, 1000V out but our switch can only withstand 100V, then a secondary of ten or more times the number of primary turns allows this safely. The inductance calculation applies to the primary winding, and the secondary is then calculated after the primary is defined. However, note that although the resulting two-winding assembly looks like a transformer, it most certainly does not behave like one. It is fundamentally very different as the next section will show. Also note that the winding polarity is important. The positive voltage spike at the switch only becomes a spike of the same polarity on the secondary if the winding is in the same sense – start and stop are in the same direction. If these are reversed, a negative spike results. This is one way to make a negative voltage generator.

Using Ferrite Cores

Now we look at the characteristics of the ferrite materials we have to use to make our typical SMPSU inductor. An air cored coil has no theoretical limit to the current, or volt-seconds it can endure. Practical issues of size, flux leakage, efficiency and just about everything else make air cored coils totally impractical in SMPSUs.

Ferrite cores saturate if too much current is passed though a winding around them in an attempt to get too high a magnetic flux density. Saturation in ferrite materials as a function of current is particularly difficult to define. For one thing inductance enters into the equation, and inductance itself depends on the ferrite material characteristics. But, basic electromagnetic theory assists us, as it also defines flux density in terms of volt-seconds (which we've used already) and the cross sectional area of the core, making no reference to the core material. This leads to the most important and fundamental design equation applicable to any cored inductor :

$V . t_{on} = N.A.B.$

N is the number of turns, A is the cross sectional area of the core in metres^2 and B is the flux density in Teslas. For nearly all ferrite materials, saturation happens at around B = 0.3T and most suppliers advise keeping B to a value of 0.15T or lower to minimise core losses and heating. (Incidentally, B_{MAX} for iron cored transformers used at 50Hz is much higher, being around 1.4Tesla. This is very fortunate for the power generation industry with their 250MW supergrid transformers!) For any given core size, voltage input and operating frequency this defines the minimum number of turns we are allowed to have to avoid saturation. It is independent of the core material characteristics or its inductance.

A pure ferrite core will always end up with quite a high, and somewhat unpredictable value of inductance. Few ferrites are stable over temperature or manufacture from batch to batch. Also, with the minimum number of turns defined above, the resulting inductance will be very much higher, by typically several tens or hundreds of times, than that needed for SMPSU operation. The solution is to use a core with a gap machined in it. The gap, which is typically less than 1mm, reduces the inductance to a value that is determined almost completely by the gap length and cross-sectional area, and not the ferrite material. The ferrite now serves only to channel all the magnetic flux into this tiny and controlled gap. To avoid complex equations, manufacturers of gapped ferrite cores supply an A_L value for each particular core. This is the inductance, usually in nH, of one turn passed around the core assembly. As inductance always increases with the square of the number of turns we know that for any desired value of inductance, $L = A_L .N^2$.

We now have all the tools needed for our complete flyback PSU and can safely leave the maths and theory behind.

A Practical Design

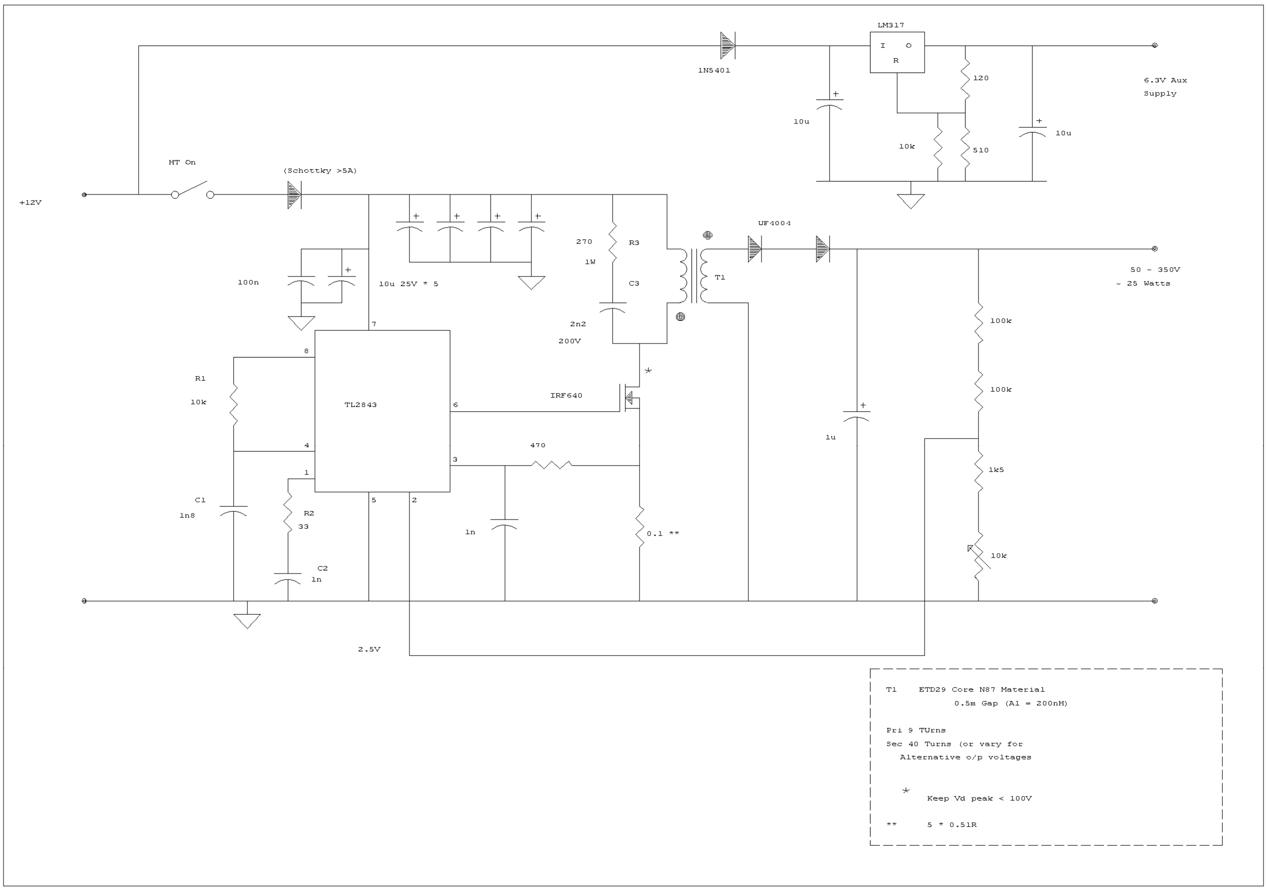


Figure 2 shows the complete circuit diagram of the high voltage power supply.

The core is a modern gapped type using N87 material. This is a low loss ferrite material designed for SMPSU operation at 100 to 300kHz. All the catalogues list a huge range of core shapes and sizes, I selected a popular shape, in its smallest size, referred to as ETD29. This is available from Farnell [2] as a gapped core with different gap sizes for different A_L values. The 0.5mm gap has a quoted A_L of 200nH. Cross sectional area of that core type is 76mm^2

Plugging these into the equations above dictate a minimum of nine turns on the primary for 100kHz operating frequency at a worst case duty cycle that I arbitrarily chose as 0.75, with 14V maximum V_{IN} while keeping B below 0.15T. Since maximum duty cycle should never coincide with maximum V_{IN} there is plenty of safety margin built in for most practical operating conditions. The resulting primary inductance for 9 turns then becomes 16uH.

The IRF640 switching FET chosen is rated for 200V maximum and 18A, with an on resistance of 0.15 ohms, so should be more than adequate and not need much heatsinking. The 200V rating is an absolute maximum so, for safety, the voltage peak (the off voltage spike) should not really go much above half this, or 100V. A maximum value for the output voltage of 450V was assumed. With 100V maximum allowed on the FET switch, that means a turns ratio of 1:4.5, so with 9 primary turns the secondary ends up at a nice round 40 turns.

Now for the SMPSU controller. In the past (although several years after the TWT PSU) I had used the UC3843 device and knew it well. Somewhat annoyingly this proved to be obsolete and unobtainable from the regular suppliers. Fortunately one supplier suggested the TL2843 as a direct replacement for the UC2843, so this device was used instead. It is an 8 pin chip and probably the simplest SMPSU controller to get going. A single RC network (R1 / C1) sets the operating frequency. The chip contains an accurate 2.5V reference so a potential divider taps the output down to this value for feeding-back. A potentiometer in the bottom of the divider allows the output voltage to be varied from around 50 to 350V.

A current feedback pin is provided to limit the maximum current in the coil winding. This is sensed by the voltage dropped across a 0.1 ohm resistor in the source of the switching FET. In normal operation this current sense input does nothing, but as soon as the voltage drop exceeds 1V (10A flowing) the switch is turned off. If the current limit were not there, then excessive load demand cold lead to excessive switching current draw leading to ferrite saturation and FET overheating. With high switching frequency and fast edges, stray inductance in these resistors must be avoided, so the 0.1 ohms is made up from five 0.51 ohms units in parallel. A low pass filter removes sharp spikes from the sense voltage before it enters the controller chip to prevent extraneous triggering of the overcurrent protection. The only other components around the chip are the series pair (R2/ C2). These are part of the control loop and affect the loop dynamics. They don't appear to be critical here, so I just used values from an earlier design.

High Voltage Side

The output rectifier posed a few problems. At 100kHz operating frequency, high voltage diodes are not straightforward. They have to be fast switching, with a low storage time, but this is incompatible with high voltage operation. I had a substantial quantity of UF4004 diodes which are rated at 400V and 50ns storage time. This voltage rating is not really sufficient for a 300V PSU, especially if I eventually run at a higher output. The obvious solution was to use a UF4007 diode rated at 1000V. However, after some discussion on the RSGBTech group [3] the consensus was that the UF4007 with its 75ns storage time might be worse than using two 50ns UF4004 diodes in series. As I had so many of the latter, and would have had to buy UF4007 devices, the decision was clear! A 1uF 450V capacitor smoothes the output. The single 450V rated capacitor shown is adequate for operation up to 350V as designed. If the PSU is eventually used at 450V, this will have to be replaced – probably by just placing two capacitors in series, and the addition of voltage sharing resistors across each.

The only components worthy of further mention are the snubber network across the inductor primary (R3/ C3). These are what I commonly refer to as a bodge. They are not a fundamental part of the SMPSU circuit operation, but are there for practical reasons. In any SMPSU, and especially where two coupled windings are used, as they are here, there will always be some ringing immediately after the FET has switched off. This partly due to unwanted stray leakage inductance, but also a damped resonance of the primary inductor with the off-capacitance of the FET .This ringing can be at several MHz and can cause problems in the rectifier diode and other places, as the high frequency oscillatory components cause additional losses. In fact, without the snubber components in place (I forgot them originally) several rectifier diodes destroyed themselves from overheating. The snubber values are chosen to absorb the bulk of the ringing energy pulses, but not absorb too much from the main switching waveform. Values are not too critical, and many designers use a suck-it-and-see approach while monitoring the waveforms. I calculated values based on suggestions in a paper on SMPSU design [4] that appeared to work very well indeed, with all ringing pulses suppressed.

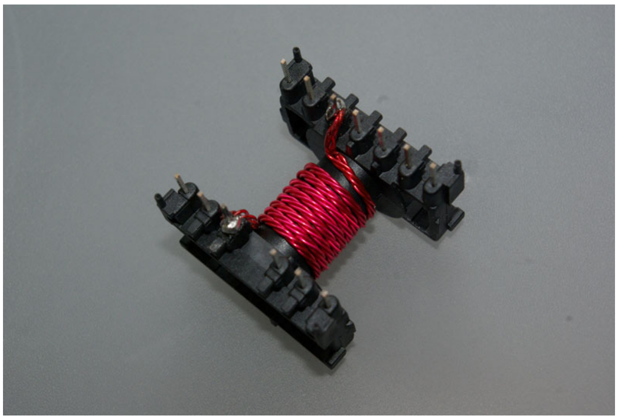
Making the Inductor

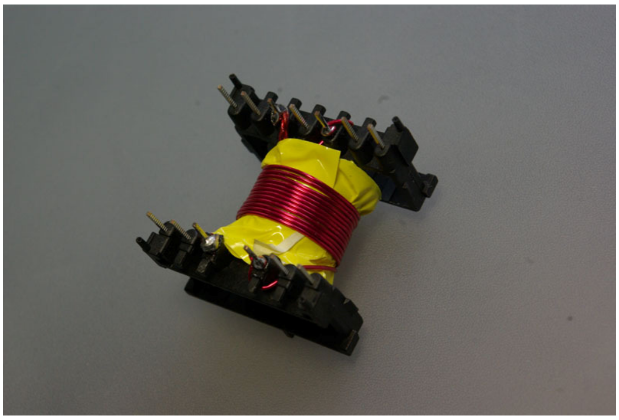
The ETD cores come in various parts, consisting of the ferrite core itself, coil former and two clips. All need to be ordered separately. To add confusion, cores actually come as core halves. A particular A_L value is defined for a gapped half-core in conjunction with an upgapped half core. So beware, you need to order two different half-core types. For example, all the different parts to make up the complete inductor assembly, with the manufacturer's part numbers, is shown in Table 1.

Table 1

EPCOS B66358G0500X187	FERRITE CORE, ETD29, N87	0.5mm GAP	
EPCOS B66358G0000X187	FERRITE CORE, ETD29, N87	0.0mm GAP	
EPCOS B66359W1013T1	ETD29, 13PIN Coil Former		
FERROXCUBE CLI-ETD29	ETD29; Clip Style:Mounting Clip; Length		Quantity 2 required

The primary is the most critical winding as it has to have a low loss resistance and therefore wound with thick wire. At 100kHz the skin depth in copper is only of the order of 0.2mm, so there is little point in using a single conductor diameter much more than 0.4mm at this frequency. The primary was wound by twisting together four strands of 0.5mm diameter enamelled wire to make a crude Litz type of construction. More strands of even thinner wire selected to give a similar cross sectional area for the copper would no-doubt work slightly better, but the advantage would probably go unnoticed. It is not that critical. Wind the primary turns as shown in Photo 1. The 40 turns for the secondary are wound with thinner wire. I had a substantial quantity of 0.5mm enamelled conductor and used this, in four layers of ten turns each with insulation tape between each layer. The enamel itself is adequate insulation for the voltages here, but layers of sticky tape between the windings does allow a slightly neater construction. The final layer of the secondary can be seen in Photo 2.

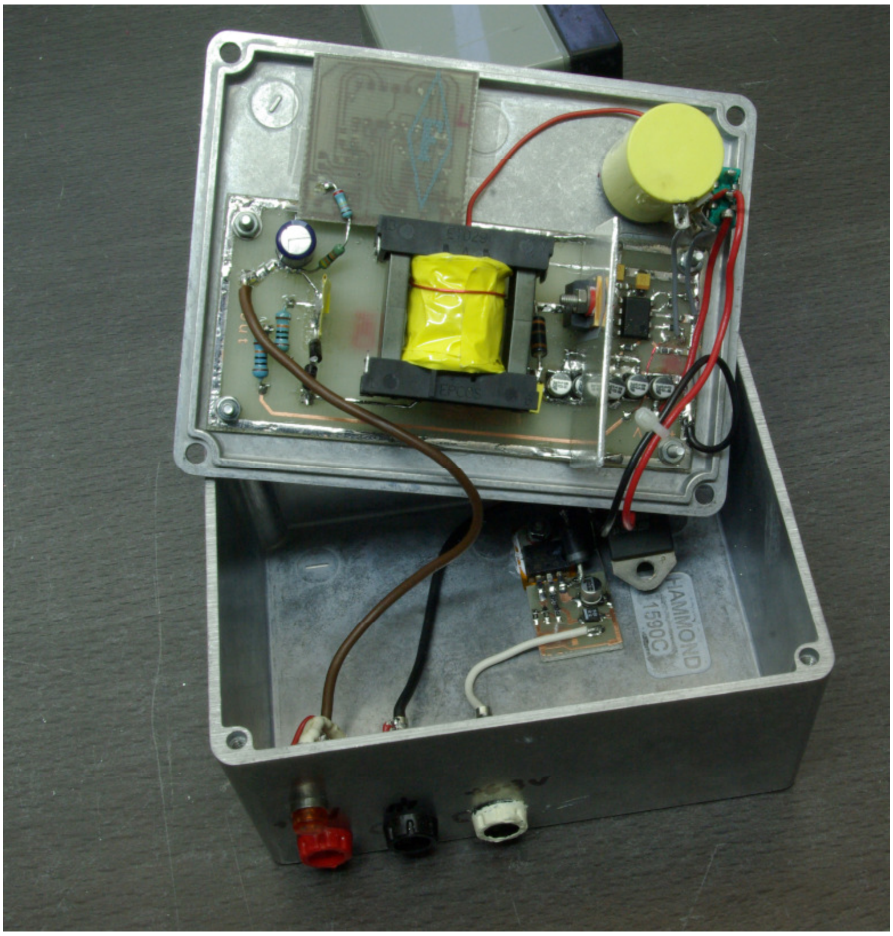




Construction and Layout

Low input voltage flyback PSUs are quite hostile environments as far as EMC and construction is concerned. High currents are being switched abruptly and voltage drop leads to extra loss. Also, voltage drop is not only due to resistive losses. Even the inductance of PCB tracks can contribute loss with the fast switching edges in use here. Thick low inductance connections are essential around the high current switching loop consisting of the FET, inductor primary, sense resistors and input decoupling capacitors. For the latter, several 10uF capacitors are used to minimise stray inductance.

The current sensing resistor must have extremely low inductance, so is made up from several small SMT units in parallel. Although five 0.51 ohm resistors were used here, ten 1 ohms resistors in parallel would have been a better choice.
The rest of the circuitry is non critical and any reasonable construction medium can be employed. Ensure you keep decent spacing around the high voltage areas.



I designed a single sided PCB using mostly surface mount components. To assist track placement, a DIL packaged version of the controller was used, mounted SMT style so it was easier to pass tracks between the pins than if a smaller package type designed for SMT construction were used.

As this PSU is intended for powering valve equipment, it was convenient to include an additional 6.3v regulator for the heater supply. This is formed from an LM317 regulator with feedback resistors suitable for a 6.3Vouput. The circuit of this part appears in the top right of Figure 2.

Photo 3 shows the innards of my version of the finished unit, mounted in a diecast box. The 6.3V regulator sits on the bottom of the box, using the case as its heatsink. The small PCB mounted upside down contains a small Digital Voltmeter module used to display the output voltage.

Conclusions

In this description I have attempted to describe the basic concepts needed to design a flyback type of Switch Mode PSU from first principles. Using the outline described here, it should be possible to extrapolate the design to other voltages and power levels using different core types and windings. For example, to generate 2000V at the same power, the primary can stay the same, but the secondary would now need something like 180 turns and the feedback resistors changed to suit the increased voltage.

Revisiting SMPSU construction after 20 years one factor that has come to light: just how much cheaper the components have become. These days SMPSUs are everywhere, (even in light bulbs!) so the price of things like ferrites, FETs and fast diodes has plummeted. While not being able to recall exactly how much I paid for an ETD29 core set in 1990, I do remember it was quite expensive, costing substantially more in actual pounds than it does now. So after inflation is taken into account the price has reduced by, perhaps, five times or more.

References

[1] http://www.r-type.org/ National Valve Museum
[2] http://uk.farnell.com/ Farnell Components. (Search for ETD29)
[3] http://groups.yahoo.com/group/rsgbtech/
[4] *Unitrode Power Supply Design Seminar*. Many references can be found on line by searching for this term, and similar ones