CHAPTER 4: Simple Flyback CONVERTERS

History:

Many high voltage converters use a flyback topology because it removes the need for a separate inductor in the output circuit. Buck regulators and feed-forward converters require such an output inductor, and the newer Cuk converter requires two. To circumvent this need, high voltage flyback converters rely on the built-in inductance of the step-up transformer secondary to store energy and hand it off to the output capacitor during rectification. This is an energy efficient way to process energy. At first, many power supply designers are struck by the term: "flyback" because it is not really obvious what is actually "flying back" in the circuit. In this chapter we will study at length this topology for both step-down and step-up use, taking a detailed look when stray capacitance is in the circuit.

Let's settle the name first. The flyback name originates from early days of television sets, back in the late 1940's when the high voltage to power the picture tube was generated by what is called a fyback transformer (See Figure 4.1). To conserve components and ensure proper timing, this transformer was powered by the same circuitry used to magnetically scan the electron beam across the face of the CRT. As the horizontal scan line progressed from left to right, it got to the edge and "flew back" in its spatial position to start a new trace, blanking the scan at the same time. By generating a high voltage pulse during this "phosphor off time" any noise generated by the pulse didn't matter because the tube was essentially dark when the pulse happened. To move the electron beam required a great deal of current and this was provided by the horizontal output tube, usually a 6BG6, which could almost carry an Ampere pulse. As a result of all of this inductance driving, the voltage on the plate of that tube "flew up" to such a high potential (several thousand volts) you could draw a high voltage RF arc to a screwdriver blade if held close to its anode cap on the top of the tube. This sudden pulse voltage was used in not only the deflection but channeling the spike current into the primary allowing the step-up "flyback" transformer to create the 10,000 volts needed to power the black and white picture tube. Designers at that time placed this dangerous transformer in its own metal cage preventing inquiring hands from touching the device and getting shocked or burned. Many an armchair technician was instantly surprised with a high voltage RF arc jumping to one's fingers just due to body capacitance. Located in the high voltage cage was also a high voltage diode tube, usually a 1B3, whose floating one volt filament was driven by a single loop of wire wrapped around one leg of the magnetic core of the flyback. The 1B3 really never glowed brightly but it was enough to rectify the HV output.

The old TV chassis in Figure 4.1 shows such a flyback transformer connected to the anode of the 1B3 tube. Somewhat visible underneath that tube was the high voltage ceramic filter capacitor (500 pF) with one lead being connected directly to the metal chassis (ground). The large tube to the right is the 6BG6 driver powering the primary of the transformer. The second smaller tube on the right was a damper diode that prevented excessive ringing on the primary. Auxiliary windings from the flyback went to control the fidelity of the horizontal yoke magnet that steered the electron beam in the CRT. Dust buildup was a vast problem in old TV sets. A large arc could cause a failure, or even worse, a fire.

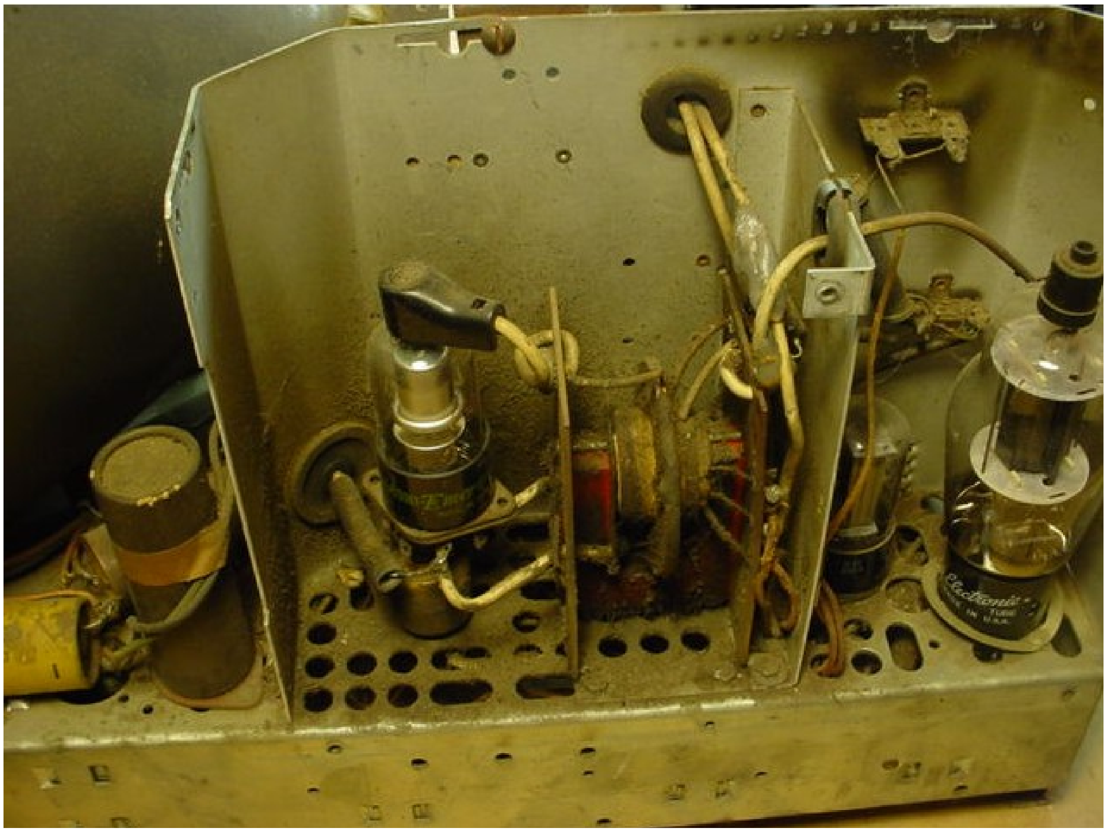


Figure 4.1: A dusty HV power supply for a GE TV. (electronixandmore.com)

Flyback technology:

Understanding the operation, equations and waveforms of flyback topology will allow us to understand more complicated systems as we move forward through this book. One thing we will find in all high voltage power supplies that utilize high frequency transformers – is that we are always fighting the capacitance "seen" by the secondary of the transformer. This includes both winding and stray capacitance which may add up substantially. This capacitance is a main factor that lowers the efficiency in high voltage power supplies by increasing the primary current draw because this capacitance has to be charged and discharged every cycle. As you will see, the nice linear ramps we saw in the last chapter on the Ringing Choke Converter now become plagued by resonant effects due to this stray capacitors. Any increase in flowing currents will be met with lower efficiency from i – squared R loss. We will find ways to get around this capacitance.

Figure 4.2 details a simple low voltage step-down flyback schematic drawn in LT SPICE. Neglecting stray capacitance for the moment, when the switch (M1) is on, energy is placed into the primary of the transformer, and the DRAIN *current* ramps up storing magnetic energy in the primary inductor. When the switch turns off the DRAIN *voltage* now rings up – exactly like in the ringing choke design and energy is transferred through the output diode to the output capacitor and load. This schematic shows the first cut at making a 150 to 3V step-down converter with a load of 10 Watts.

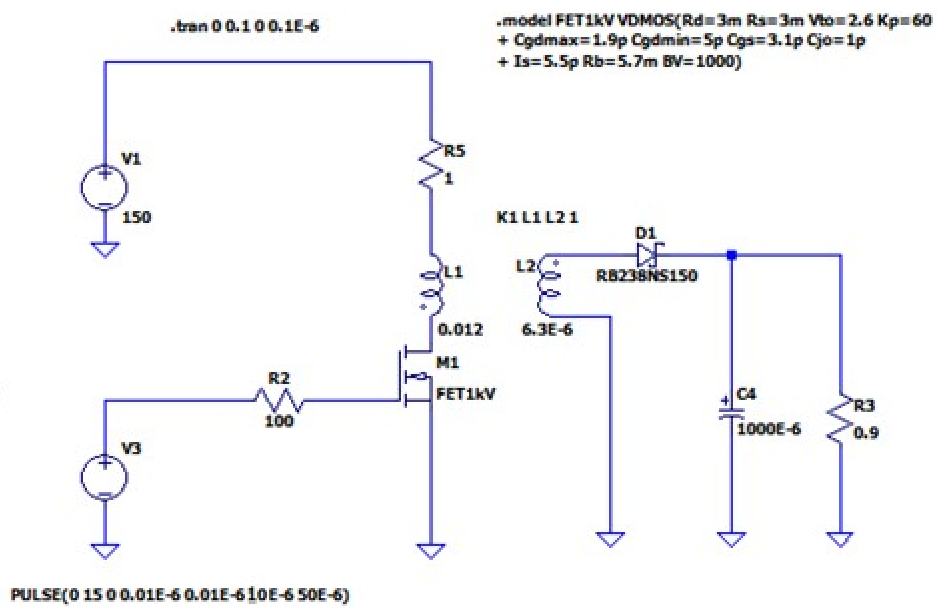


Figure 4.2: Basic LT SPICE 150V to 3V step-down flyback converter

To configure a transformer in LT SPICE, just link the two inductors together using a K statement as shown from a SPICE directive available at the top line in LT SPICE by: .ap. In the schematic you can see we are linking L1 with L2 with a coupling coefficient of 100% - something you want to start off with just to see the first order run. You can always lower this down to more realistic values later on. Typical high voltage transformers have a value of 0.95 or less and we will cover this more in Chapter 5.

Figure 4.2 shows this converter running at 20kHz (look at the gate drive pulse generator) with a duty cycle of only 20%, that is, the ON-time of the switch is set to 10 µS. Upon running this simulation we find that the output is only 1.04 volts DC – it is not making the required 3.0V that we need. Wanting to raise the output voltage and store more energy in the primary of the transformer we will increase the ON-time but how much is enough? In this example we have chosen the primary inductance to be 12 millihenries and the secondary inductance to be 6.3 µH for a reason we will see in example 1 coming up next. Figure 4.3 shows the simulated DRAIN waveforms using a 1kV FET as the switch.

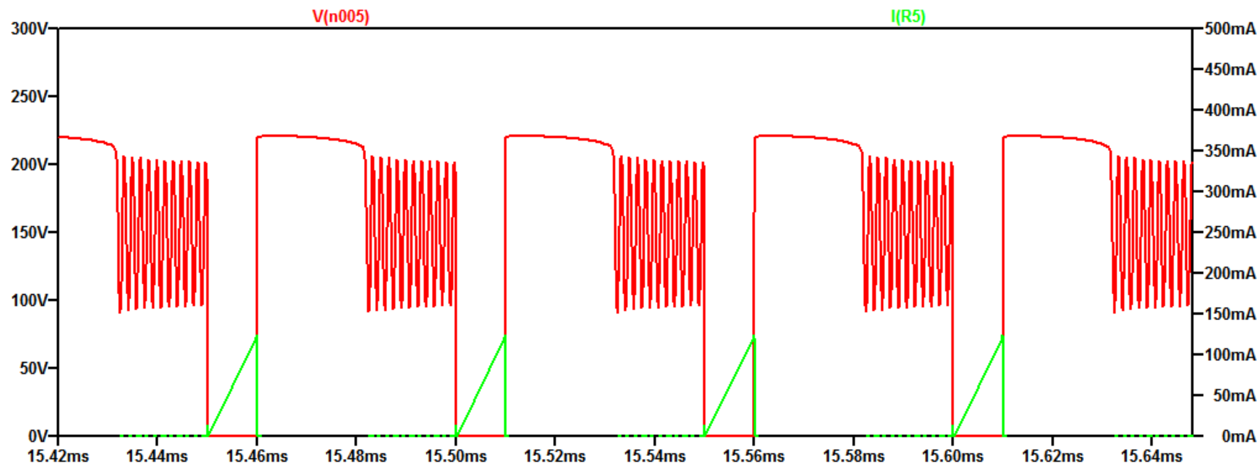


Figure 4.3: DRAIN voltage (red left), DRAIN current (green right)

Let's examine this waveform. Notice that the voltage on the transistor DRAIN (red) is nearly zero when the transistor is on for the 10 µS - causing the primary current (green) ramp up. At the end of the ON-time the switch shuts off and the DRAIN voltage rings highly upwards to a value limited by the energy we stored in the primary, much like in our ringing choke converter. Looking at the schematic we have inserted a one Ohm resistor in series with the primary of the transformer to take into account winding and connection resistance. This is also helpful to make the LT SPICE program converge easier, especially when the coupling coefficient is less than unity. The 1000 µF output capacitor means that our ripple will be low and the simulation will need at least 4.5 mS to run (five RC time constants). We will run to 50 mS just to be safe to be far away from any transient effects that always occur at the beginning of a simulation. We have not included any secondary winding capacitance in this simulation to keep it simple – the secondary turns at most are probably less than a dozen.

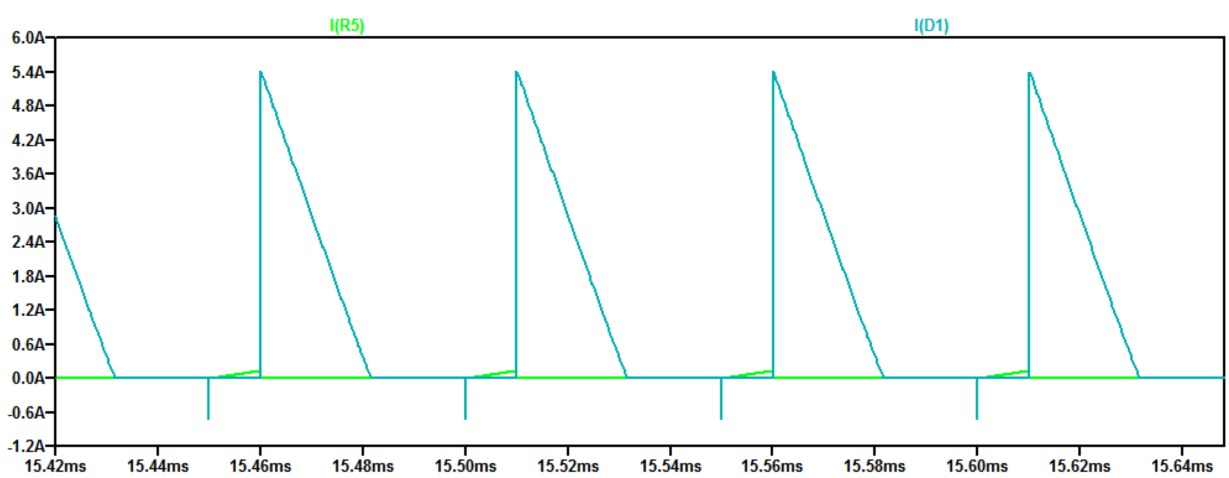


Figure 4.4: Drain current (green) and output diode current (light blue)

Figure 4.4 compares the DRAIN current (green) with the output diode current (light blue). Notice the diode current is much larger because tis is a step-down converter. The transformer turns ratio governs the current flow just as it does the voltages that appear at the secondary. Also notice that there is an off-time in Figure 4.4 where all currents go to zero. This appears as a burst of high frequency in Figure 4.3 where the DRAIN rings at 592 kHz. This is caused by the inductance of the transformer resonating with the transistor internal capacitance forming the high frequency region. It is not caused by the output capacitor.

The big take away from this circuit is: a typical flyback circuit charges the output DC filter capacitor and load through a diode supplying energy from the transformer secondary after the switch has turned off. Replenishing the charge lost in a capacitor is accomplished at nearly 100% efficiency when this current is supplied by an inductor – no matter how deep the capacitor has been discharged. You may recall that elementary circuit theory shows when an *empty* capacitor is charged from a voltage source (no inductor) it loses half of the energy from the source in the form of resistance heating. Even if the resistance value is reduced practically towards zero, half of the energy is still dissipated in whatever resistance is left. That's why we are charging the capacitor from the inductance of the secondary of the transformer – it is very efficient!

Why is flyback topology useful? You may have noticed that some step-down low voltage power supplies use *feed-forward* topology. This requires a separate inductor be inserted before the output filter capacitor and the inclusion of an extra diode, a free-wheeling diode to keep the current flowing when needed. High voltage power supplies using feed-forward topologies are usually impractical because the inductance needed is typically very high. For example, a 5 volt output line operated computer switching power supply in feed-forward mode may only need tens of microhenries as an inductor but a high voltage feed-forward circuit, producing, say, 3 kV would require inductors as large as one Henry in the output circuit – a rare component to find and one with large series resistance. As you will see, low voltage switching power supplies have an inductor in their output circuitry. Most high voltage switching power supplies do not.

The operation of a flyback converter is easy to mathematically analyze. As mentioned earlier, when the transistor switch is turned on, energy is stored in the magnetic field of the transformer due to current flowing through the primary of the transformer. Because of the separation in circuits between primary and secondary, no energy feeds to the output during the primary ON-time by virtue of the orientation of the output diode - the phasing of the transformer insures this condition. The energy magnetically stored depends upon the primary inductance and the peak current flowing through the primary:

$$E = \tfrac{1}{2} L_{primary}\, i_{primary\ peak}^{\ 2} \qquad (4\text{-}1)$$

When $L_{primary}$ is in Henries and the current, $i_{primary\ peak}$, in Amperes, the energy stored in the magnetic field has the units of Joules. When the switch is turned off this magnetic field energy is released in the form of current flowing from the secondary winding through the output diode flowing into the load resistance and topping off the output filter capacitor. Unlike the RCC, no further energy is drawn from the battery after the switch shuts off.

Let's examine the equations that govern how a flyback converter operates. It will be easier if we start with a step-down converter because stray capacitance effects will not bother us as we develop the equations of operation. Later on, when we attempt step-up designs we will have to be careful to take this capacitance into consideration.

If you recall the ringing choke converter of Chapter 3, we found that the highest efficiency was obtained when the OFF-time, τ_{OFF}, is reduced to zero limiting the loafing of any power component in the circuit. Let's start off by determining the best ON-time of the switch. We will use the voltage across an inductor equation remembering we have to deal with two inductances this time, the primary L_P , and secondary L_S , as well as their peak currents, $i_{PRIMARY\ PEAK}$, and $i_{SECONDARY\ PEAK}$ respectively. It was easier in the RCC circuit where there was only one inductance to deal with and now because of the transformer, there are two. As usual we will stipulate that the input source is a battery, V_{BAT}, and we have enough output capacitance to keep the ripple low for our simulation. From our inductor equations we have:

$$V_{BAT} = L_P\, i_{PRIMARY\ PEAK} / \tau_{ON} \qquad (4.2)$$

$$V_{OUT} + V_D = L_S\, i_{SECONDARY\ PEAK} / \tau_{RING} \qquad (4.3)$$

allowing for zero OFF-time. Adding the times together, we get the total period:

$$1/f = \tau_{ON} + \tau_{RING} \qquad (4\text{-}4)$$

so:

$$1/f = L_P\, i_{PRIMARY\ PEAK} / V_{BAT} + L_S\, i_{SECONDARY\ PEAK} / (V_{OUT} + V_D) \qquad (4\text{-}5)$$

Since the primary and secondary are wound on the same core, their inductances depend on the square of the turns. Based on this we can write the following relationship:

$$(N_S / N_P)^2 = L_S / L_P \qquad (4\text{-}6)$$

where N_S is the number of secondary turns and N_P the number of primary turns. By virtue of Faraday's Law, the primary and secondary voltages are linked by the turns ratio but the primary and secondary peak currents are inversely related to the winding ratio:

$$i_{PRIMARY\ PEAK} \,/\, i_{SECONDARY\ PEAK} \quad = \quad N_S / N_P \qquad (4\text{-}7)$$

Combining the above relationships together we can obtain the ON-time for the switch that has a zero OFF-time:

$$\tau_{ON} \quad = \quad (V_{OUT} + V_D) / f\,((V_{OUT} + V_D) + (N_S/N_P)(V_{BAT})) \qquad (4\text{-}8)$$

Although we don't know the *exact* number of turns on the primary or secondary yet, we do know the winding ratio is really the voltage gain of the transformer from primary to secondary.

$$\textit{winding ratio} \quad = \quad N_S / N_P \qquad (4\text{-}9)$$

and, it is reasonable to assume that this ratio is related to the input and output voltages. Taking the diode drop into account we have:

$$\textit{winding ratio} \quad = \quad V_{OUT} + V_D / V_{BAT} \quad = \quad N_S / N_P \qquad (4\text{-}10)$$

This does not include systems that utilize voltage multiplying circuits. Combining the three above equations together we get:

$$\tau_{ON} \quad = \quad 1 / 2f \qquad (4\text{-}11)$$

This tells us that the best ON-time for zero OFF-time is obtained when we operate the switch for exactly half the period time. This of course was to be expected since we forced the winding ratio to be linked with the input and output voltages. To find the value of primary inductance we will need, all we have to do is consider the storage of energy into the system. As we said before, the only input of energy to the entire system occurs when the switch is closed and a current ramp is formed:

$$E_{IN} \quad = \quad \tfrac{1}{2}\, L_P\, i_{PRIMARY\ PEAK}^{\,2} \qquad (4\text{-}1)$$

because this happens at a frequency f, the power input is:

$$P_{IN} \quad = \quad \tfrac{1}{2}\, f\, L_P\, i_{PRIMARY\ PEAK}^{\,2} \qquad (4\text{-}12)$$

We can find L_P, the primary inductance by adding the effect of assumed efficiency, η, in our calculation,:

$$P_{OUT} = \eta P_{IN} = \tfrac{1}{2}\,\eta\, f\, L_P\, i_{PRIMARY\ PEAK} \qquad (4\text{-}13)$$

combining this equation with (4-1) we get:

$$L_P = \eta\, V_{BAT}{}^2 f\, \tau_{ON}{}^2 / 2P_{OUT} \qquad (4\text{-}14)$$

using 4-11:

$$L_P = \eta\, V_{BAT}{}^2 / 8 f P_{OUT} \qquad (4\text{-}15)$$

This value of inductance, when running with an ON-time equal to half the period of operation will give us just enough energy in to provide the power out. Let's try an example.

Example 1: Take the flyback circuit of Figure 4.2 determine the inductances and ON-time to provide a 3.0 Volt output at 10 Watts running at 20kHz.

Listing the parameters for our converter we have:

V_{BAT}	=	150 V
V_{OUT}	=	3.0 V
f	=	20kHz
P	=	10W
η	=	0.85 (assume lower than 90% due to low output V)
V_D	=	0.45 V

Using the parameters for our converter we find:

winding ratio = $V_{OUT} + V_D / V_{BAT} = N_S / N_P = 0.023$

τ_{ON} = 25 μS

and L_P is calculated using equation (4-15) to be:

$$L_P = \eta\, V_{BAT}{}^2 / 8 f P_{OUT} \qquad (4\text{-}15)$$

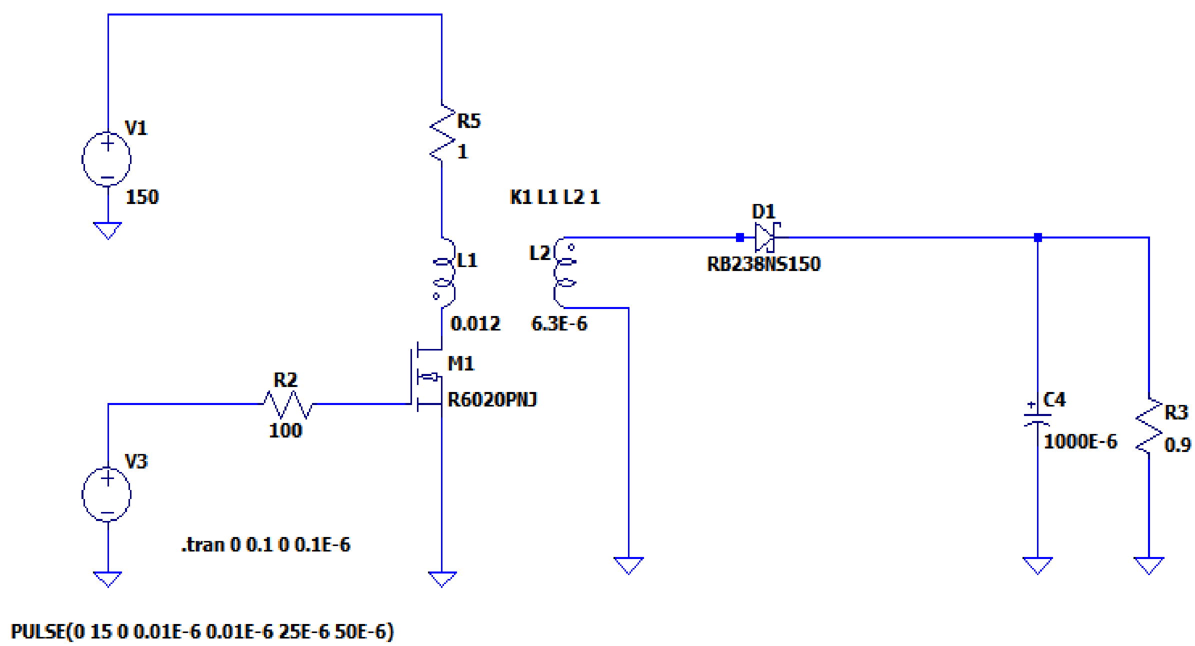
$$L_P = 0.012\ \text{H}$$

using equations 4-6 and 4-10 we get;

$$Ls = Lp\,((Vout + VD)/VBatt)^2 \qquad (4\text{-}16)$$

resulting in: $Ls = 6.30\ \mu H$

Now, using the above equations we can form a schematic and run a simulation.



. Figure 4.5: 150V to 3V step-down 10W converter

Figure 4.5 shows such a circuit and Figures 4.6 and 4.7 shows the waveforms of this circuit when running steady state.

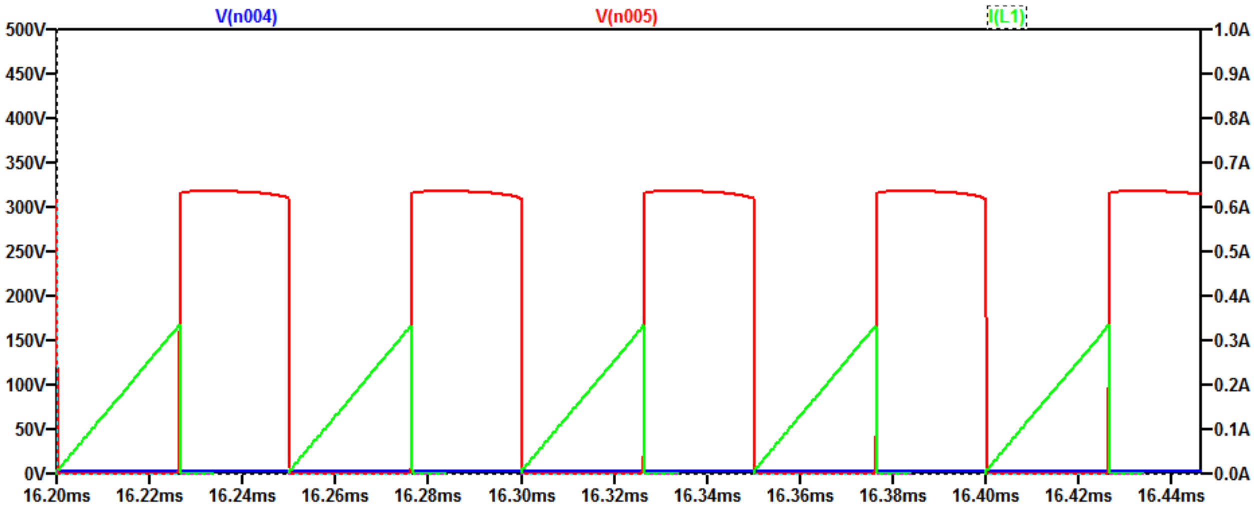


Figure 4.6: Drain voltage (red) and Drain current (green) for 150V to 3V flyback converter

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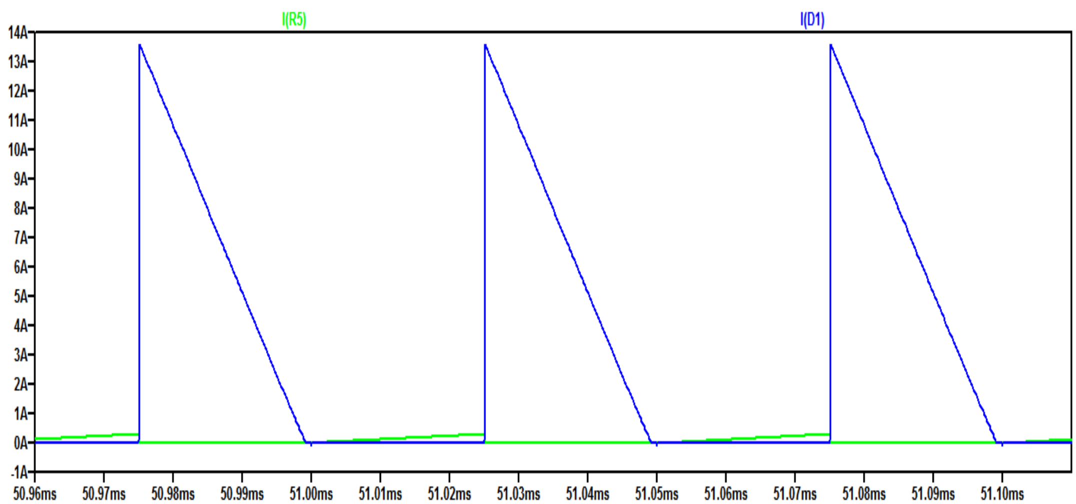

Figure 4.7: Secondary current (blue) and Primary current (green) for flyback converter

The LT SPICE simulation yields the following important data for this 150 yto 3V converer::

IPRIMARY PEAK:	=	0.327 A
IIN AVG.:	=	0.091 A
VOUT:	=	3.14 v

The efficiency is: η = POUT / PIN = 10.955 / (150)(0.091) = 80.3%

While this efficiency can be considered low, some blame is to be put on the 1.0 Ohm resistor in series with the primary.

Example 2: Using the topology of Figurer 4.2, design a 135 volt to 30 volt flyback converter for 100 Watts running at 50 kHz to power a string of white LEDs. Here are the parameters:

V_{BAT}	=	135 V
V_{OUT}	=	30 V
f	=	50kHz
P	=	100W
η	=	90% (assume)
V_D	=	0.7

Using the parameters for our converter we find:

$$\text{winding ratio} = V_{OUT} + V_D / V_{BAT} = N_S / N_P = 0.23$$

$$\tau_{ON} = 10\ \mu S$$

and L_P is calculated using equation (4-15) to be:

$$L_P = \eta\, V_{BAT}^{\,2} / 8\, f\, P_{OUT} \tag{4-15}$$

$$L_P = 410\ \mu\ H$$

for the secondary inductance we get:

$$L_S = L_p\, ((V_{out} + V_D)/V_{Batt})^2 \tag{4-16}$$

yielding: $L_S = 21.4\ \mu H$

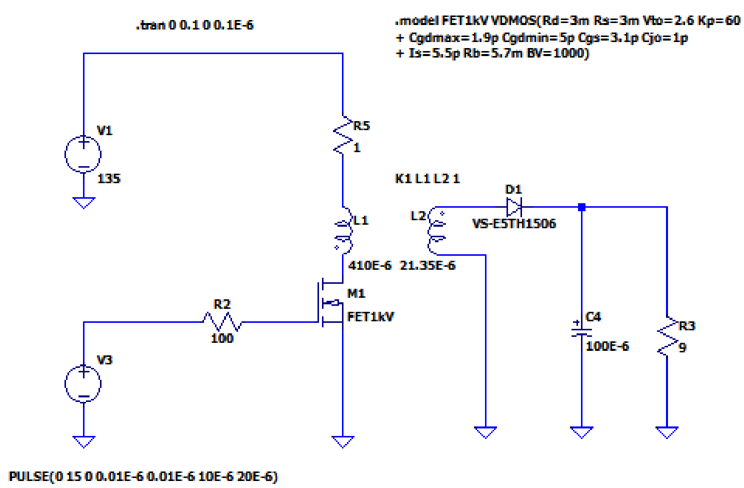

Figure 4.8: 150V to 30V step-down flyback converter running at 50kHz

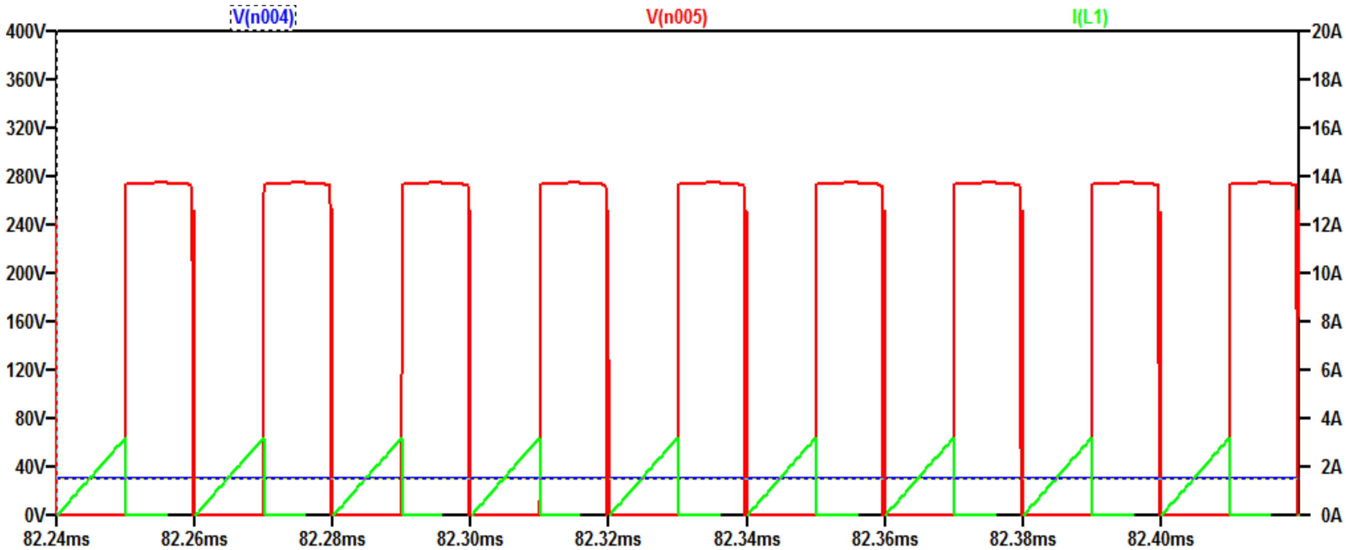

Figure 4.9: 150V to 30V 100W step-down flyback converter with limited off-time
DRAIN voltage (red - left axis), DRAIN current (green – right)
Vout (blue – left axis)

The LT SPICE simulation yields the following important data:

IPRIMARY PEAK	=	3.17 A
IIN AVG	=	0.832 A
VOUT	=	30.67 v

The efficiency is: $\eta = P_{OUT} / P_{IN} = 104.5 / (135)(0.832) = 93.1\%$

a value very close to what we assumed it would be and higher than Example 1 because we do not have the diode loss in the 30V converter as we did with the Schottky diode in the 3V converter. Notice in Figure 4.9 that off time is almost zero indicating that we are using all of our components to the fullest extent. Also notice that the DRAIN waveform shoots upward in voltage, to 270 volts, well above the applied B+ voltage of 150 VDC one of the reasons we had to use a 1kV FET in our simulation. As before with the Ringing Choke Converter, if the load is suddenly removed there had better be a way to cut back the ON-time of the switch to keep the output voltage constant. If not, the DRAIN voltage may rise up to and beyond the breakdown voltage of the switch. To offset this, a pre-loads may be incorporated at the output but this will also waste power and lower the efficiency of the converter. It sometime comes down to a trade-off between having a preload and efficiency. Also, for high input voltage you have to be careful and take steps to utilize a semiconductor switch with sufficient breakdown capabilities in your design. In this case with a 135 VDC input, a device with a BV_{DS} of 400 volts is a minimum requirement, and that is just squeeking by - marginally allowing for 10% high line specification.

As mentioned, our simulation shows there is a very small off-time τ_{OFF} component. This occurs once the energy stored in the magnetic field of the transformer secondary has been fully transferred to the output capacitor and load. At this point the DRAIN voltage drops to the voltage of the incoming line, in this case 135 VDC, as shown. Here τ_{OFF} lasts less than a microsecond. Unlike the RCC when current is always being sourced by the battery at ON-time and RING-time both, current here only flows from the battery input when the switch is conducting. All energy to power the load has to be placed into the magnetic field during this ON-time by virtue of the primary current ramp (green trace).

This simulation of waveform for a step down converter shows the transistor switching with very fast rise and fall times. This is usually seen in practice with very little effort because the capacitance seen by the secondary of the transformer reflected to the primary is next to zero. Secondary capacitance is reflected back to the primary increases by the *square of the winding ratio allowing* step-down converters to often have textbook waveforms with minimal ringing.

$$C_{\text{SEEN ACROSS PRIMARY}} = C_{\text{SEEN SCROSS SECONDARY}} (N_S / N_P)^2 \qquad (4\text{-}17)$$

but we have adjusted both N_S and N_P in our converters to track the input and output:

$$V_{OUT} + V_D / V_{BAT} = N_S / N_P$$

$$C_{\text{SEEN ACROSS PRIMARY}} = C_{\text{SEEN SCROSS SECONDARY}} (V_{OUT} + V_D / V_{BAT})^2 \qquad (4\text{-}18)$$

For example, if the total secondary capacitance (winding and stray) is 20 pF, the capacitance as seen across the primary is:

$$C_{\text{SEEN ACROSS PRIMARY}} = (20\text{pF})(30.8/135)^2 = 1.04\text{ pF}$$

Almost zero. Once we start to build step-up converters the situation changes completely.

Example 3: 12V to 100V Flyback Converter

The previous circuit was designed to run a LED array. Let us design a converter to make 100 volts from 12 volts at a power level of 20 Watts to run a piezoelectric element. We will stick with the frequency of 50kHz because that makes our filtering a little easier. This is our first step-up design.

V_{BAT}	=	12 V
V_{OUT}	=	100 V
f	=	50kHz
P	=	20W
η	=	90% (hopefully)
V_D	=	0.80 V

We will use a standard silicon PN junction diode with breakdown above 100 VDC.

$$L_P = \eta V_{BAT}^2 / 8 f P_{OUT} \tag{4-15}$$

We find from equation (4-15) the primary inductance to be (using our constraints):

$$L_P = 16\ \mu H$$

and the secondary inductance is calculated from equation as:

$$L_S = L_p ((V_{out} + V_D)/V_{Bat})^2 \tag{4-16}$$

$$L_S = 1.1\ mH$$

We will be diligent and add some winding capacitance in the secondary of approximately 5pF. We will also lower the series resistor to 0.1 Ohm and hope this can be realized in practice. From LT SPICE we obtain the following schematic operating with an ON-time of 25 µS as is our practice:

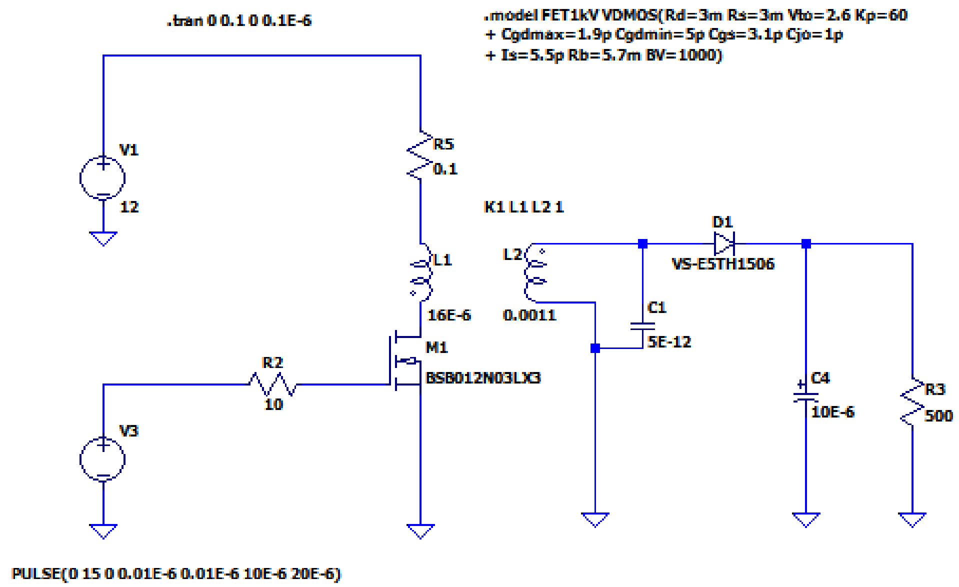

Figure 4.10: 12 volt to 100 volt 20 Watt step-up flyback converter

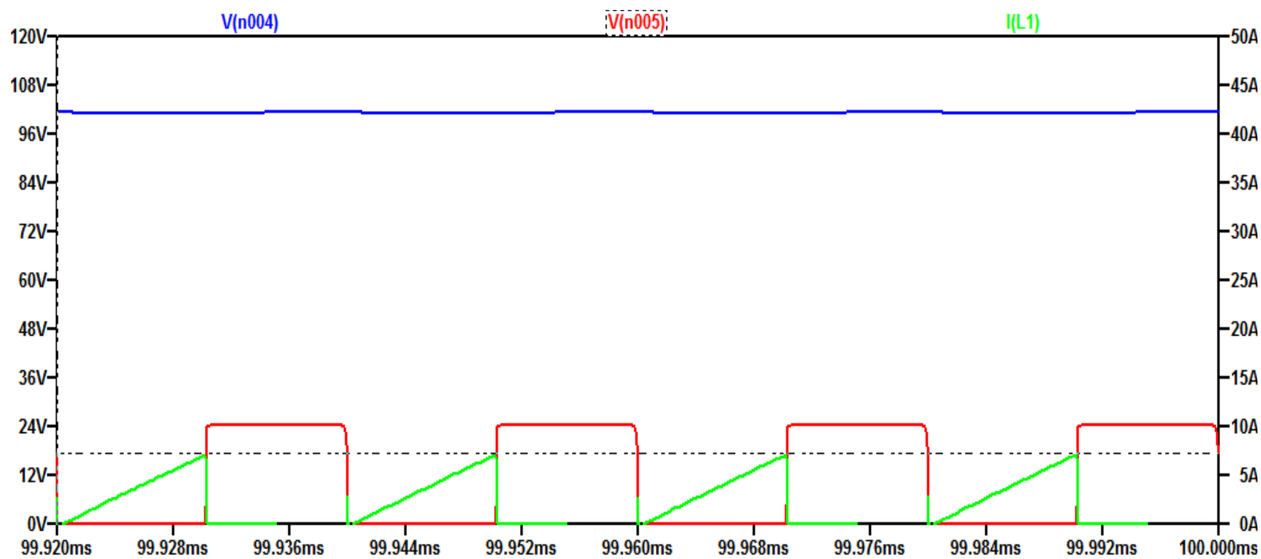

Figure 4.11: DRAIN voltage (red, left axis), DRAIN current (green, right axis), Vout (blue, left axis).

The LT SPICE simulation yields the following important data:

IPRIMARY PEAK	=	7.04 A
IIN AVG	=	1.80 A
VOUT	=	101.98 Volts

The efficiency is: $\eta = P_{OUT} / P_{IN} = 20.79 / (12.0)(1.80) = 96.3\%$

The efficiency is overly high because we have used a 0.1 Ohm series resistor. This is hard to achieve in real life especially for a 1 milliHenry inductance. A more realistic value of 0.5 Ohms is probably better to use but this will drop our efficiency to 89% based on the average input current. Sometimes in simulations it is better to err on the conservative side. We are on a roll!

Example 4: 12V to 3,000 V Flyback Converter

Let us raise the step-up even higher and design a converter that makes 3,000 volts from 12 volts at a power level of 10 Watts. We will run at 20kHz this time.

$$V_{BAT} = 12\text{ V}$$
$$V_{OUT} = 3{,}000\text{ V}$$
$$f = 20\text{kHz}$$
$$P = 10\text{W}$$
$$\eta = 85\% \text{ (hopefully)}$$
$$V_D = 3.5\text{ V}$$

We will use a standard silicon HV diode with a 5kV breakdown voltage made from at least five junctions in series. From equation (4-15):

$$L_P = \eta V_{BAT}^2 / 8 f P_{OUT} \tag{4-15}$$

Giving us the primary inductance value:

$$L_P = 76\ \mu H$$

and the secondary inductance is calculated from equation (4-16) as:

$$Ls = Lp\ ((Vout + VD)/VBat)^2 \tag{4-16}$$
$$Ls = 4.8\ H$$

a rather high value. Again, we should add some winding capacitance in the secondary especially for such a large inductance. We will use 20 pF. We will also keep the series resistor at 0.1 Ohms because every drop there becomes an efficiency problem. Lowering the output filter capacitance to a reasonable value of 0.1 µF is also warranted to allow for a faster steady state time in our simulation. For our simulation we placed several 1kV diodes in series just to be safe.

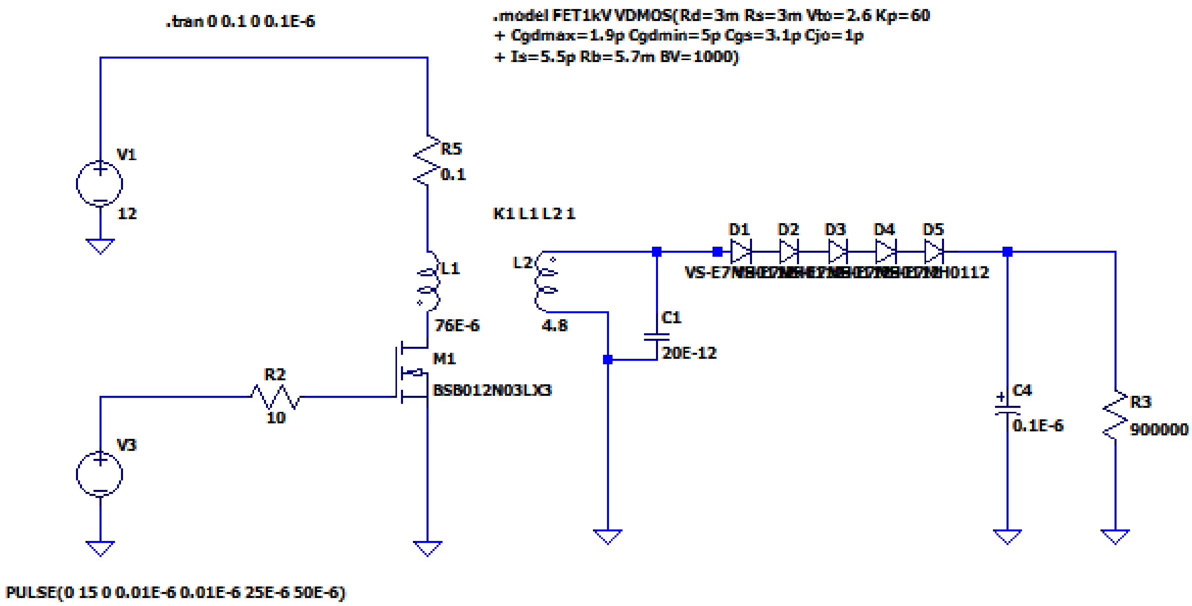

Figure 4.12: 12 to 3kV flyback converter with added secondary capacitance

LT SPICE simulation:

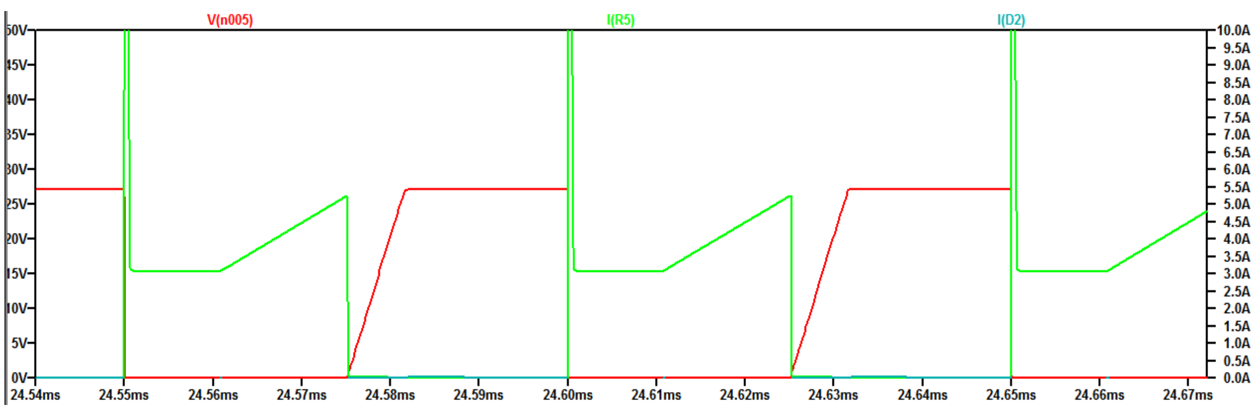

Figure 4.13: Waveforms showing DRAIN voltage (red left) and DRAIN current (green right)

Something seems to be really wrong with our waveforms because Figure 4.13 doesn't look at all like any of the earlier ones. For one thing, the DRAIN voltage on the switching transistor seems to slowly rise upwards, not abruptly and the current ramp of the switching FET already has a value of 3 Amperes when the switch turns on – that never happened before. In addition, there is a huge current spike too at switch turn-on probably wasting energy somewhere. What is causing this strange behavior? Let's see what the efficiency is as we did before using the following data:

IPRIMARY PEAK	=	4.75 Amperes (discounting the spike)
IIN AVG	=	2.84 Amperes
VOUT	=	3,404 Volts

This gives an efficiency of:

$$\eta = \text{POUT} / \text{PIN} = 12.87 / (12.0)(2.84) = 37.7\%$$

pretty dismal and not in line with our earlier simulations. Could it be the 20 pF capacitor we added to the transformer secondary caused this problem? How could such a small capacitor cause such a large problem? Let's run the LT SPICE simulation without this capacitor to see.

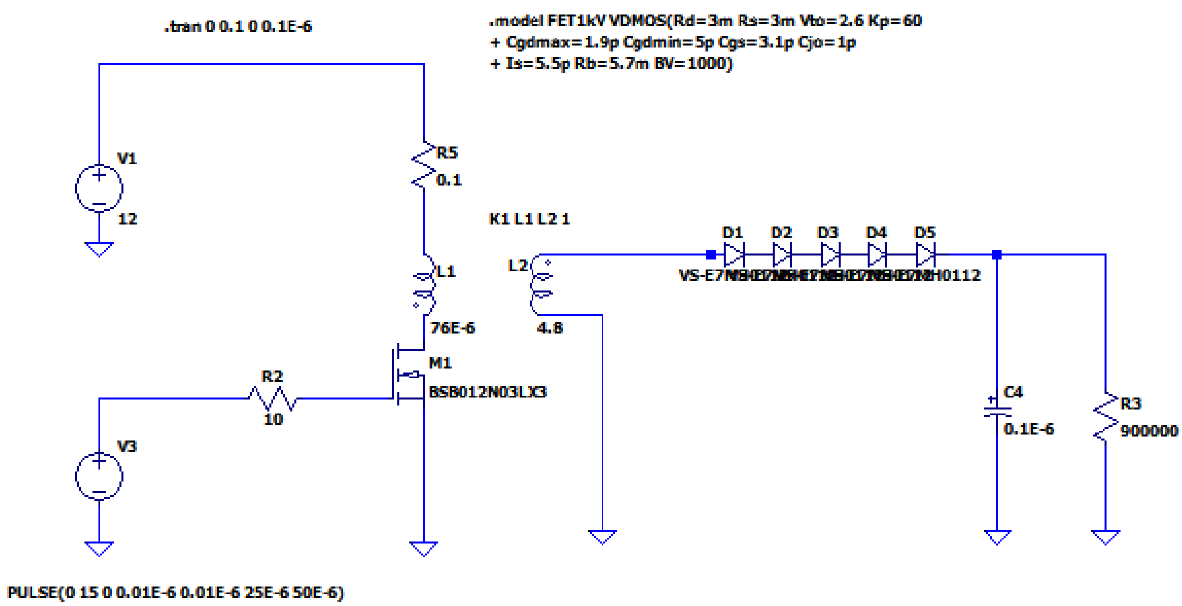

Figure 4.14: 12 to 3kV converter minus the secondary stray capacitance

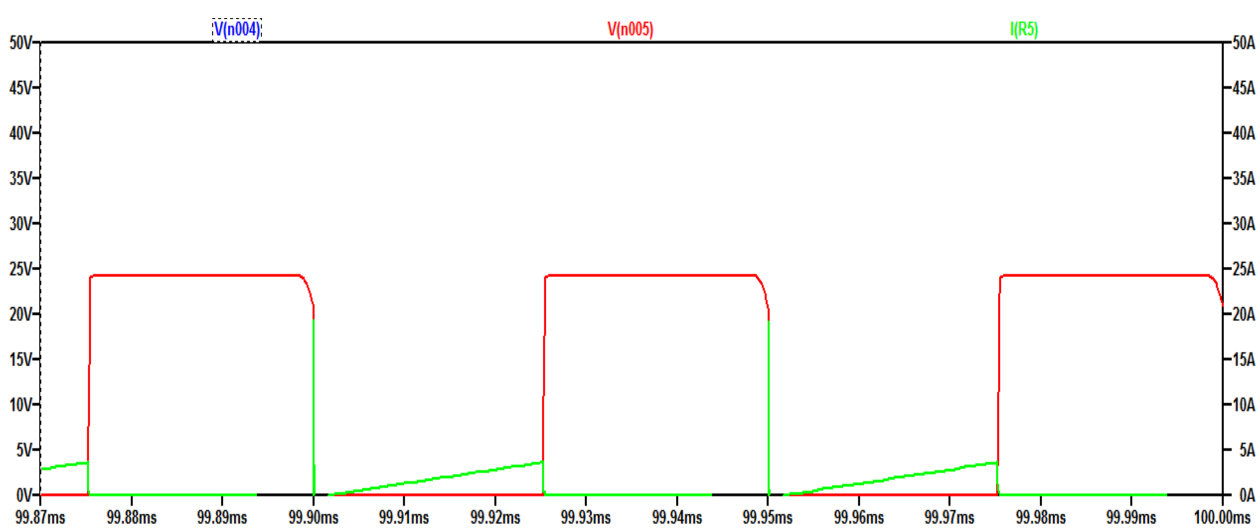

Figure 4.15: 12 to 3kV converter no secondary capacitance

Looking at Figure 4.15 we see the textbook waveforms again and no horrendous current spike. The current ramp looks very linear and starts from zero. Let's take a look at the efficiency of this simulation without the 20 pF secondary capacitance:

$I_{PRIMARY\ PEAK}$	=	3.66 (discounting the small spike at turn-on)
$I_{IN\ AVG}$	=	0.881 Amperes
V_{OUT}	=	3,063 Volts

This gives an efficiency of:

$$\eta \quad = \quad P_{OUT} / P_{IN} \quad = \quad 10.42 / (12.0)(0.881) \quad = \quad 98.6\%$$

Now that's more like it! Can't we just present this simulation as our design and be done with it? The answer is sure, but it would be wrong and a waste of time. Building up the circuit with a transformer having a secondary with 4.8 Henry's of inductance will certainly have some winding capacitance that we must take into account because, remember, it gets reflected back to the primary by the square of the turns ratio. You can prove to yourself that even if the secondary capacitance was reduced down the efficiency for our circuit (12V to 3kV converter) would be:

Secondary capacitance:	20 pF	Efficiency:	37.7%
Secondary capacitance:	10 pF	Efficiency:	73.2%
Secondary capacitance:	5 pF	Efficiency:	81.8%
Secondary capacitance:	1 pF	Efficiency:	92.2%
Secondary capacitance:	0 pF	Efficiency:	98.6%

That's why our previous example, the 12 to 100 volt converter worked quite well even with our equations even though it had a 5 pF secondary winding capacitance and besides the gain was rather low.

So what should we do? Looking at the poor waveforms of Figure (4.13) it seems to reason that the low resonant frequency of the secondary winding inductance acting with the 20 pF capacitor is causing the transistor DRAIN to ring up very slowly. The large secondary inductance and capacitor have a resonant frequency as given by the simple frequency equation:

$$f \quad = \quad 1/2\pi\ \mathrm{SQRT}\ (LC) \qquad (4\text{-}19)$$

Using our values of L and C we have:

$$f \quad = \quad 1/2\pi\ \mathrm{SQRT}((4.8)(20E\text{-}12))$$

$$f = 16.24 \text{ kHz}$$

Here is the problem! We are trying to drive this circuit above it's resonant frequency. No wonder why our waveforms are poor and efficiency is terrible. The only thing we can do in this situation is to lower the secondary inductance to a value where the resonant frequency is much higher than the driving frequency. Because of the stray capacitance and high step-up our derived model equations really don't help us much.

As a test, let's reduce the secondary inductance by a factor of ten to 0.48 H. That smaller inductance can also show a smaller winding capacitance, an extra gain for us but let's just keep it the same for now at 20pF. Quickly calculating the resonant frequency:

$$f = 1/2\pi \text{ SQRT}((0.48)(20\text{E-}12))$$
$$f = 51.36 \text{ kHz}$$

Much higher and we can operate at 20kHz and simulate our circuit out. Because our analysis has taken a turn here, we will start with an ON-time of just 10 µS or 20% of the period. We can calculate a primary inductance if we hold fast to the voltage, inductance and turns ratio rule:

$$(N_S / N_P)^2 = L_S / L_P \tag{4-6}$$

and:

$$\textit{winding ratio} = V_{OUT} + V_D / V_{BAT} = N_S / N_P \tag{4-10}$$

Since we are defining our secondary inductance to be 0.48 H and our step-up is 250, the primary inductance is calculated to be:

$$\text{Lpri} = \text{Lsec} / (\text{Vout} / \text{Vin})^2$$
$$\text{Lpri} = 7.6 \ \mu\text{H}$$

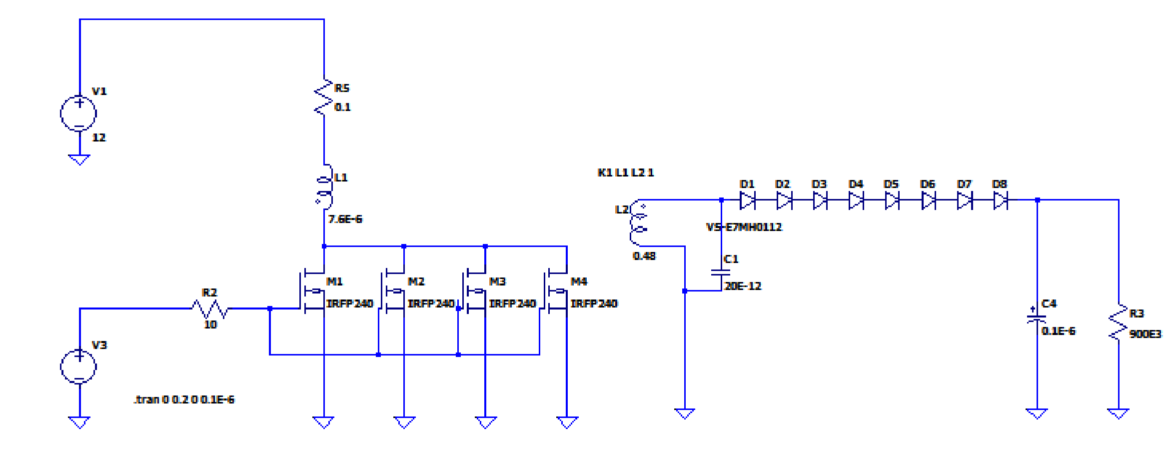

Figure 4.16: 12V to 3kV flyback converter with lower inductance values

In the schematic it shows that instead of selecting a higher current FET we just used one we know, the IRFP240, a TO-247 plastic device and simply put four in parallel. That is easier to do in SPICE than solder up four devices. Figure 4.17 details the DRAIN waveforms:

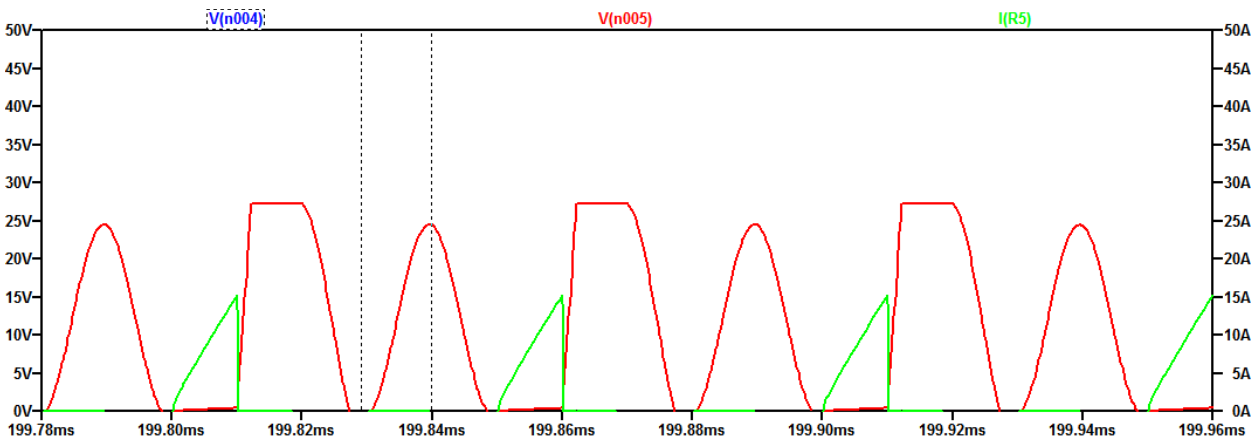

Figure 4.17: DRAIN voltage (red,left) and DRAIN current (green right) τ_{ON} = 10 µS

The output voltage for this simulation was 3.82kV indicating that we can cut back on the ON-time in an attempt to get it down to 3kV. Here is the data taken from the simulation with an ON-time of 10 µS:

$I_{PRIMARY\ PEAK}$	=	15.0 Amperes
$I_{IN\ AVG}$	=	1.56 Amperes
V_{OUT}	=	3,820 Volts

The efficiency calculates out to:

$$\eta = \text{Power out / Power in} = 16.2/(12)(1.56) = 86.6\%$$

This is much better than our 37.7% we originally had with the first design 4.8 Henry secondary inductance. In measuring the time of the ring dotted lines show a half period of ring-out, and using the easy LT SPICE calculator, the frequency of the ring seems to be 46kHz, vey close to the 51.36 we calculated earlier. Notice also that there is a lot of OFF-time where the components are loafing again. We could get rid of that by speeding up our frequency to start the ON-time just after the first ring-up, just as the DRAIN voltage is dropping.   Running the converter simulation again at a full period of 25 µS, we get the following waveforms and data:

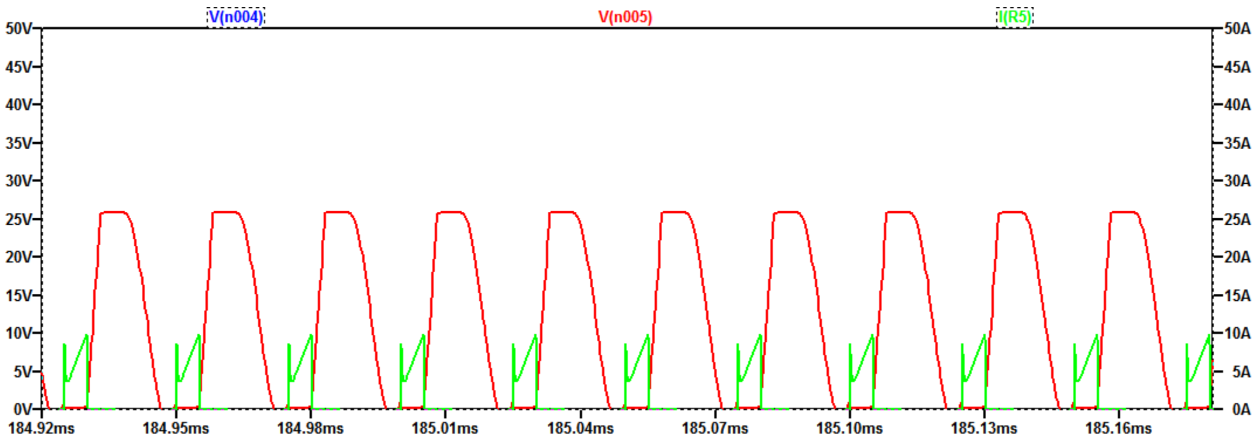

Figure 4.18: DRAIN voltage (red,left) and DRAIN current (green right) τ_{ON} = 5 µS

IPRIMARY PEAK	=	9.28 Amperes
IIN AVG	=	1.216 Amperes
VOUT	=	3,450 Volts

The efficiency calculates out to:

η = Power out / Power in = 13.22/(12)(1.216) = 90.6%

Example 5: 28V to 3,000 V 100W Flyback Converter

What happens if we increase our input voltage to 28 VDC a typical military avionics value? In this example we are also increasing the power level but keeping our secondary winding capacitance constant at 20 pF. Remembering the hard fought previous example, we should be careful because the step-up is still 107. Here are the following parameters for this converter:

VBAT	=	28 V
VOUT	=	3,000 V
f	=	20kHz (minimum)
P	=	100W

η	=	0.9 (assume)
V_D	=	3.5 (several junctions in series)
C_{SEC}	=	20 pF (will keep the same)

We will use a standard silicon HV diode with a 5kV breakdown voltage made from at least five junctions in series. Using equation 4-15 for the primary inductance, hoping that the secondary inductance will make the resonant frequency much greater than 20kHz, :

$$L_P = \eta V_{BAT}^2 / 8 f P_{OUT} \qquad (4\text{-}15)$$

We find that the primary inductance has to be:

$$L_P = 44.1 \ \mu H$$

and the secondary inductance is calculated from equation 4-16 as:

$$Ls = Lp\ ((Vout + VD)/VBatt)^2 \qquad (4\text{-}16)$$

$$L_S = 0.507\ H$$

not a very high value and almost the same as compared to the previous example. From what we learned earlier in Example 4, let's check the self resonant frequency of the secondary and the winding capacitance using the simple frequency equation:

$$f = 1/2\pi\ SQRT\ (LC) \qquad (4\text{-}19)$$

Using our values of L and C we have:

$$f = 1/2\pi\ SQRT((0.507)(20E\text{-}12))$$
$$f = 50.0\ kHz.$$

much higher than Example 4. We can certainly run at 20kHz and perhaps higher. There would be advantages to running at higher frequency especially in the area of output ripple. Using the parameters obtained above we have the schematic shown in Figure 4.19. Trying the usual ON-time of 50% we have the simulation shown in Figure 4.20.

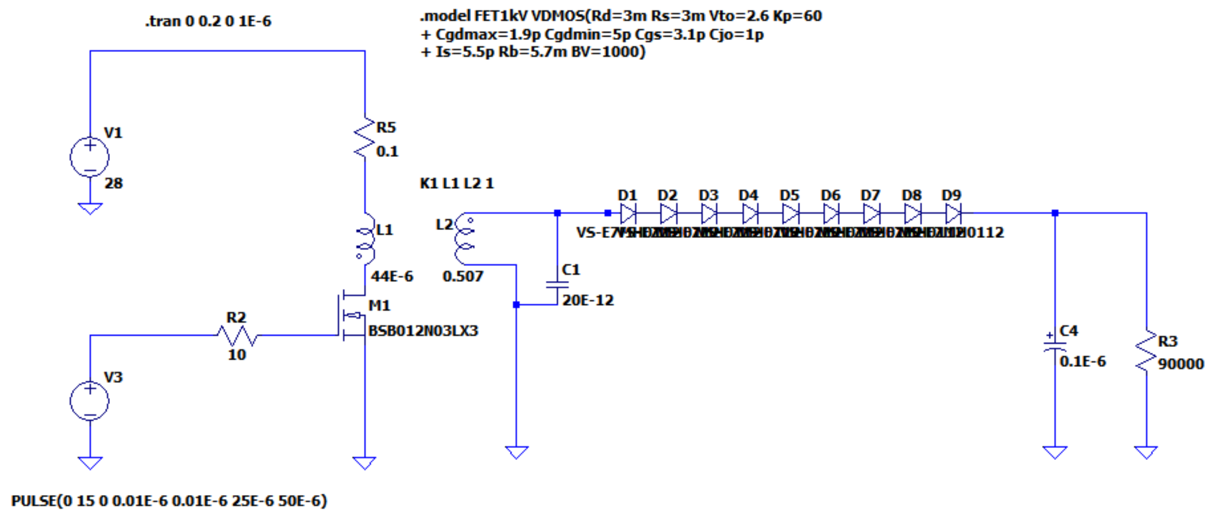

Figure 4.19: 28V to 3,000V converter run with ON-time of 25 µS, f = 20kHz.

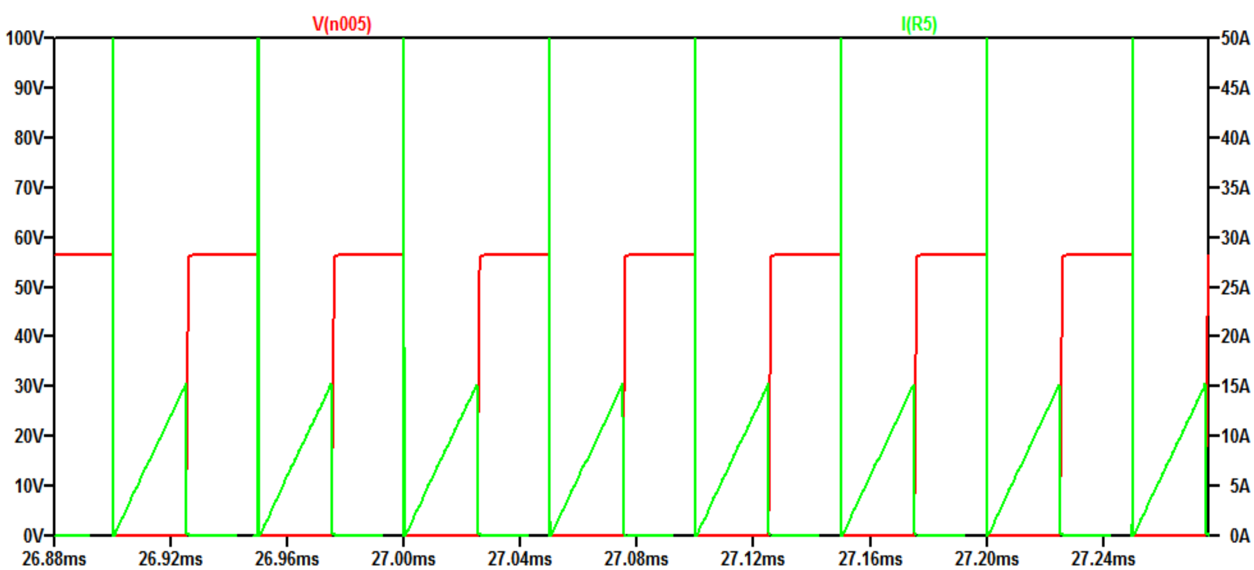

Figure 4.20: 28V to 3,000V converter. DRAIN voltage (red,left) and DRAIN current (green right)

IPRIMARY PEAK	=	15.3 Amperes
IIN AVG	=	4.138 Amperes
VOUT	=	3,058 Volts

Efficiency:	=	89.6%

Example 6: 300V to 6,000 V 1KW Flyback Converter

In this example we are also increasing the power level to 1 kilowatt and raising the output voltage to 6kV. The converter must charge up a 10 uF capacitor to obtain a moderate ripple. The end use of this power supply is to drive a high power magnetron microwave tube. Because the transformer is much larger that any other we have used we will increase our secondary and stray capacitance to 50 pF. We will still keep our frequency at 20kHz.

V_{BAT}	=	300 V
V_{OUT}	=	6,000 V
f	=	20kHz (minimum)
P	=	1000W
η	=	0.9 (assume)
V_D	=	10 (many junctions in series)
C_{SEC}	=	30 pF (large transformer – more capacitance)

We will have to use a high current fast recovery diode with at least a 10kV breakdown voltage made from at least 16 HV diodes in series. Using equation 4-15 for the primary inductance:

$$L_P = \eta V_{BAT}^2 / 8 f P_{OUT} \qquad (4\text{-}15)$$

We find that the primary inductance has to be:

$$L_P = 506 \ \mu H$$

and the secondary inductance is calculated from equation 4-16 as:

$$Ls = Lp\ ((Vout + VD)/VBatt))^2 \qquad (4\text{-}16)$$

$$Ls = 0.20\ H$$

Again, checking the resonant frequency we have, using the simple frequency equation:

$$f = 1/2\pi\ SQRT\ (LC) \qquad (4\text{-}19)$$

Using our values of L and C we have:

$$f = 1/2\pi\ SQRT((0.20)(30E\text{-}12))$$

$$f \quad = \quad 64.9\ kHz.$$

We can certainly run at 20kHz and perhaps higher. Using the parameters obtained above we have the schematic shown in Figure 4.21 using an ON-time of 50%..

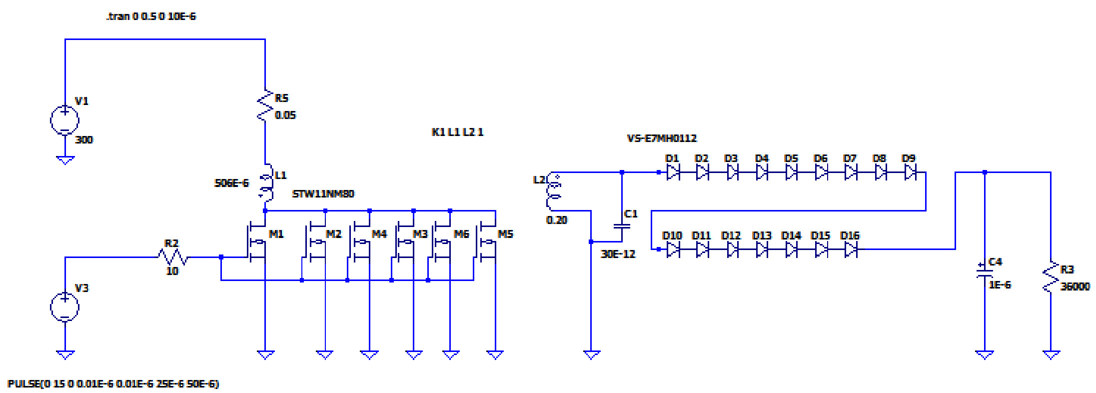

Figure 4.21: 300V to 6,000V 1kW converter run with ON-time of 25 μS, f = 20kHz.

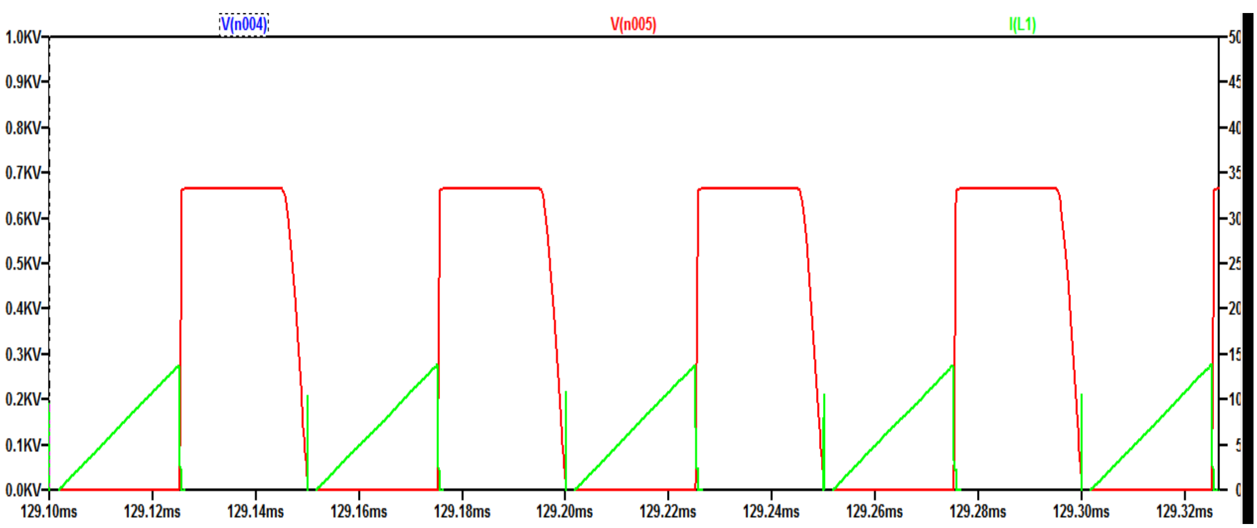

Figure 4.22: 300 to 6,000V converter. DRAIN voltage (red,left) and DRAIN current (green right)

In schematic Figure 4.22 notice the use of parallel FETs to provide the current capability. The following parameters were obtained from the simulation:

$$I_{PRIMARY\ PEAK} \quad = \quad 14.7\ \text{Amperes}$$

$I_{IN\ AVG}$	=	3.72 Amperes
V_{OUT}	=	6,058 Volts
Efficiency:	=	91.3%

When transformer coupling k is less than one

In all of our LT SPICE simulations in this chapter we assumed that the coupling in the transformer was exactly 1.0. This made the analysis run faster and gave a good chance of getting a convergent answer. As you will see, sometimes SPICE programs will jam up and not deliver an answer because they get stuck in a certain iteration they can't get out of. Internal math problems such as dividing by zero are common internal stopping points especially when dealing with transformers. Sometimes adding small series resistances (as we do) helps the simulation move forward. All SPICE programs have this difficulty and those with saturating magnetic cores are the most problematic.

Once you get a circuit that seems reasonable you should attempt to reduce the K factor and see what happens because it gets you closer to the real world situation. All transformers will be a little less than ideal when constructed no matter how much care goes into their design. This occurs especially with high voltage transformers because there is a tradeoff between coupling and voltage breakdown. In addition, high voltage power supplies are especially prone to transformer problems because the secondary turns are usually a great deal higher than the primary amount. Why is this K value less than unity?

Looking at the E core transformer shown in Figure 4.23, the primary winding is located close but not exactly on the secondary winding. It seems to reason that some of the magnetic paths of the primary simply do not couple to the secondary. When this occurs we have some built-in inductance in series with the primary circuit and that is called for one reason or another "leakage inductance". Because this flux is created but does not couple to the secondary it stores energy that cannot be utilized by the secondary winding and sent to the output load. We have see with Figure 4.20 and Figure 4.21 what deleterious effects extra inductance does to our converter. Leakage inductance just gets in the way of the power transfer.

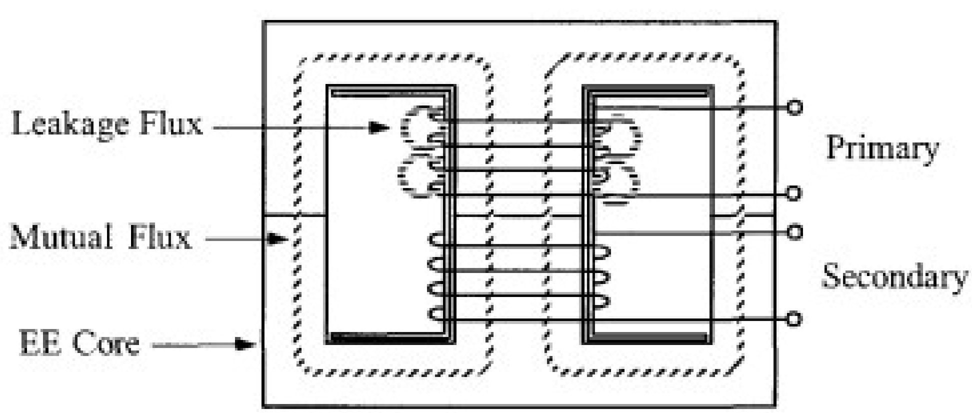

Figure 4.23: Primary flux not coupled to secondary causing "leakage inductance"

The relationship between leakage inductance and the K coupling coefficient is:

$$L_{LEAK} = L_P (1 - k^2) \qquad (4\text{-}12)$$

Where L_{LEAK} is the leakage inductance as seen reflected to the primary and L_P the primary inductance. It is usually very hard to calculate leakage inductance just from physical measurements because you are dealing with geometry of turns not coupling to others. It is however very easy to measure the leakage inductance of a transformer if you have an inductance bridge – all you need are a couple of clip leads. Measuring the leakage inductance with respect to the primary is accomplished in the following manner:

Measuring Leakage Inductance:

Steps:

1. Using low frequency to minimize capacitance effects (if frequency can be controlled, use a frequency around 100 Hz), determine the primary inductance, L_P of the transformer.
2. Short the secondary – the inductance seen at the primary winding should drop way down.
3. This new value as seen on the meter is the leakage inductance of the transformer with respect to the primary.

Most low voltage transformers (60 Hz and step-down 20kHz types) show leakage inductance values less that 1% of the primary inductance. This calculates to a K value of 0.995. If the leakage inductance is 5% of the primary, as most high voltage transformers are, the K value drops to 0.975. The farther the secondary is spaced from the primary the greater the leakage inductance. If the secondary winding was in the next room, the leakage inductance would be exactly the primary inductance and K would be zero indicating there is no coupling at all.

Winding Capacitance:

As we see, the winding capacitance plays a very important role in a flyback converter especially at light loads with high gains. It will form sinusoidal waveforms on the switch when operating at NO-LOAD conditions. The University of North Carolina published a paper several years ago detailing the calculation of both leakage inductance and winding capacitance:

https://coefs.charlotte.edu/mnoras/files/2013/03/Transformer-and-Inductor-Design-Handbook_Chapter_17.pdf

It lists several entities that add to stray winding capacitance:

1. Capacitance between turns
2. Capacitance between layers
3. Capacitance between windings
4. Stray capacitance due to terminations

Medhurst did a lot of work during the 1940's on inter-winding capacitance of air-cored single layer solenoids. An equation that is an extension of Medhursts work is:

$$C\ell = 4\ell\epsilon_o\epsilon_{rx}\,(12k_C\,(\epsilon_{ri}/\epsilon_{rx}+1)+1)\;/\;\pi\mathrm{Cos}^2\psi$$

where $$k_C = 0.106D^2/\ell^2 + 0.717439D/\ell + 0.933048(D/\ell)^{3/2}$$

and D is coil diameter, ℓ is the coil length, $\psi = \tan^{-1}(p\pi D)$, p is the pitch or turn spacing, ϵ_o permittivity constant 8.854 E-12 F/m, ϵ_{ri} is relative permittivity of the coil form material (2.56 for polystyrene, for example), and ϵ_{rx} is relative permittivity of space external to the coil (probably just air).

Reference for this is "The self-resonance and self-capacitance of solenoid coils" by David Knight. The equation is 5.3 on p 25, but there is a lot of detail of its development. This entire paragraph was found online at www.stackexchange.com. Looks like a hard equation to use. The better method is to just wind up a transformer and measure it's self resonant frequency.

How to measure winding capacitance in a high voltage transformer

Here are the steps:

1. First determine the inductance of the secondary of the transformer using the lowest frequency possible to minimize and capacitance effects.
2. Drive the transformer primary with an oscillator.
3. Using your scope probe placed near the output of the transformer (not connected to it), look for a resonant peak as you sweep frequency.
4. Using the simple resonant frequency formula, determine the winding capacitance C.

$$f = 1/2\pi\ \mathrm{SQRT}\ (LC) \qquad (4\text{-}19)$$

Values of 5 to 100 pF are common depending of course on the size of the transformer. The reason not to connect to the transformer is the probe capacitance, which might be as high as 10pF will add to the value.

Analysis of an OLD Television HV Flyback Power Section
As shown in the first figure of this chapter, the old CRT high voltage power supply that accompanied all black and white television receivers in the 1948 – 1965 time period utilized a specially wound step-up transformer having very small secondary capacitance. Caled the "flyback transformer" it operated at the horizontal sweep frequency of the picture tube, 15.734 kHz allowing some people to actually hear it working. This transformer was specially designed by its construction to have low winding capacitance and to resonate at the fifth harmonic, approximately 75kHz.

Figure 4.24 shows a typical waveform associated with the HV flyback transformer section of a typical B&W television set, powered from a 300 volt supply. The ON-time was very long, the full horizontal width of the beam travel and the ring-off occurred quickly. Primary inductance is 150 mH and secondary inductance is 1 Henry. Stray capacitance is 10 pF.

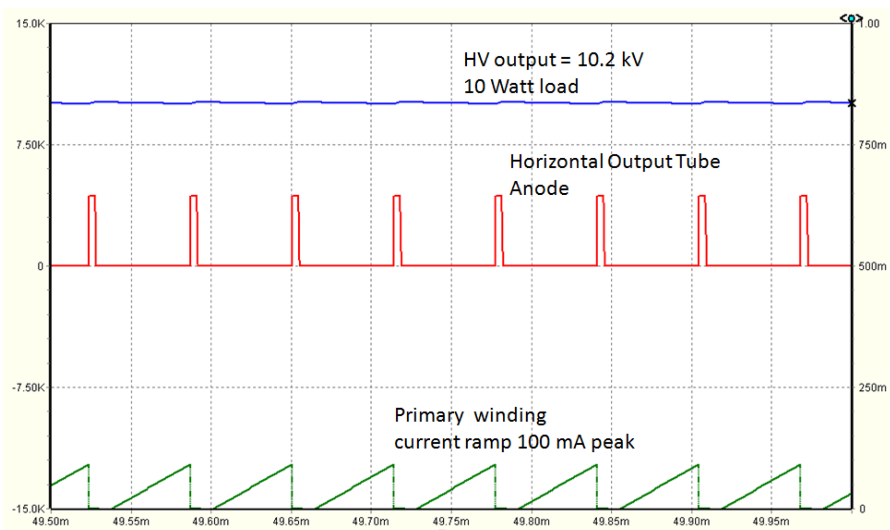

Figure: 4-24: Waveforms in old HV flyback transformers

The horizontal driver tube not only supplied current to move the electron beam across the front of the screen but it provided the drive to the high voltage flyback transformer. The green ramp details the current through the tube, almost 100 mA peak as the primary of the high voltage transformer increased in energy. When the tube was shut off the red waveform shows the huge pulse of voltage found on the anode of the tube. Blue trace shows the output after a simple half wave rectification usually in the vicinity of 10kV.

When color televisions made their debut in the late 1960's, much higher accelerating potentials were needed, usually up to 30kV with powers exceeding 20 Watts to overcome the intrusive shadow mask within the color picture tube structure that guided the electrons from the three electron guns to the proper color dot but intercepted much of them in the process. The color CRT was a finicky thing and required precise voltages so as not to de-focus the beam. To regulate the high voltage output a 30kV shunt tube, usually a 6BK4 was used right on the HV output.