Chapter 3: The Ringing Choke Converter

Sometimes a need arises to raise a lower voltage up to a somewhat higher voltage to run a needed module that requires an input greater than what a battery would provide. One example of this would be to power a silicon avalanche photo diode that usually requires 90 volts to operate for maximum sensitivity. If the load is not isolated from the battery this is an easy thing to do using a ringing choke converter. Another example is driving a piezoelectric crystal from an amplifier. This may require several hundred volts - certainly out of the range of common carbon-zinc batteries. About 60 years ago you could purchase dry cells that offered 300 volts or more that were employed to drive Geiger counters or xenon photoflash packs. Such batteries were common and readily available at your neighborhood TV repair shop. Figure 3.1 shows a 560 volt battery – about the highest made. But with the fading away of vacuum tubes such power sources have long become obsolete yet the need is still there for portable medium high voltage applications. This chapter describes an easy way to step up a low DC voltage so it can be easily used in these applications.

Figure 3.1: 560 Volt battery for photoflash applications

Ringing choke converters are one of the most basic topologies that can be made. Because the power section consists of a transistor, an inductor, a diode and one capacitor they are very reliable and can offer step-ups as high as a 100 with power levels at least to 50 Watts. We shall study the driven RCC first because it will give us the foundation we need before moving on to more complicated circuits such as those that incorporate transformers. Figure 3.1 details a SPICE schematic of a simple RCC that converts 9 volts to 30 volts. In this simple topology, a

MOSFET switch is turned on and off by a pulse of a set duration from voltage source V1. In practice this can be a pulse width modulator where the frequency and duty cycle of the pulse is easily adjusted, say be using a SG3524. To simplify the SPICE analysis a general-purpose MOSFET, the Si9420, a 200V 1 Ohm RDS is used as the switch although you could just as well insert a power bipolar transistor in that location to accomplish the same thing. To help model convergence and try to be as realistic as possible, we have incorporated a 1.0 Ohm resistor R1 in the circuit to represent the Ohmic losses that are always found in the inductor and on the circuit board. A general fast recovery diode, the 1N4937 couples the energy to the output which is read by test point TPv2. We will use a 50 uF capacitor to limit output ripple.

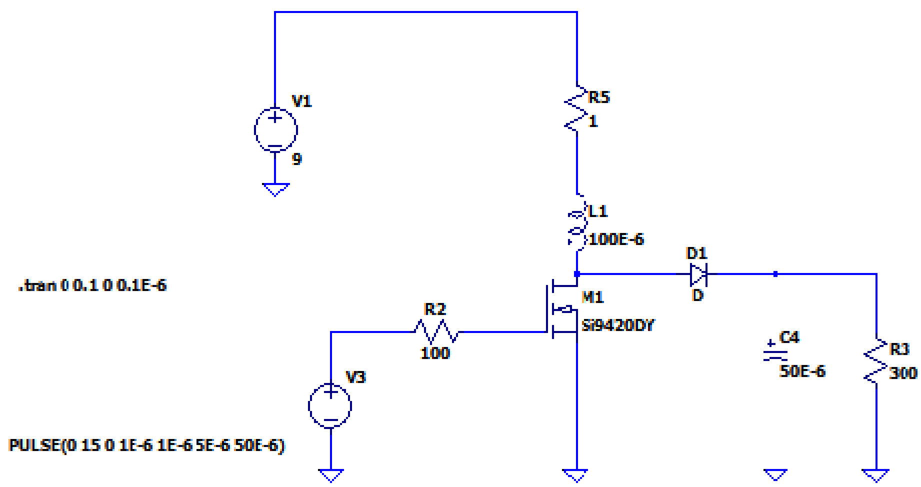


Figure 3.2: 9 to 30 volt 3W Ringing Choke Converter

Notice that there are no transformers in a true ringing choke converter. Sometimes, you may find online schematics that purport to be ringing choke converters using a transformer as the step-up mechanism. These are actually "flyback" converters and should not be confused with a ringing choke converter because a ringing choke converter uses a stand-alone inductor (choke) and not a transformer.

Looking at Figure 3.2 you will see, as mentioned earlier, a transistor switch that closes for a certain time, τ_{on} allowing current to flow into the inductor from the battery. During this time the current builds up in a linear fashion satisfying the following equation:

$$V_L = L\, di / dt \qquad (3\text{-}1)$$

where V_L is the voltage across the inductor and i, the current flowing through the inductor as a function of time, usually a ramp if powered from a DC source. Following this line of thought, if the circuit is powered from a battery, V_{BAT}, the current peak of the ramp would be given by:

$$V_{BAT} = L\, i_{peak} / \tau_{on} \qquad (3\text{-}2)$$

where τ_{on} is the on-time of the switching transistor. The highest point in the current ramp, i peak, is the value of current just before the transistor switch shuts off. The units used in this equation are Volts, Henries, Amperes and seconds. When current flows through an inductor, energy is being stored in the magnetic field according to:

$$E = \tfrac{1}{2} L\, i^2 \qquad (3\text{-}3)$$

and this energy, in Joules, is eventually transferred to the output capacitor C1 and load during the second part of the switching cycle, that is, after the transistor turns off. Referring to the schematic in Figure 3.2, when the drive pulse from V1 stops and the transistor switch opens, the voltage seen at the transistor Drain connection (Voltage Test Point 1), rings up higher than the battery voltage forcing diode D1 into a forward bias conduction, charging up the output capacitor C1 and providing energy to the load resistor R2. Because of this ring-up in voltage we have a step-up in potential above the input voltage.

One simple mechanical analogy of this circuit would be someone pulling down a mass attached to a hanging spring. When the mass is let go it suddenly shoots upwards above its equilibrium point and can transfer energy to a lever mechanism providing work output (not shown).

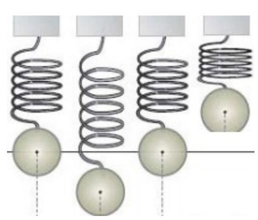


Figure 3.3: Mass-spring oscillator

The questions may be asked, why does the voltage across the inductor suddenly shoot upwards and how high does it rise? For the hanging mass analogy that is easy to see. When the weight is pulled down the spring expands and mechanical energy is stored into it. When let go, the spring pulls upwards on the weight causing it to move. Here the elastic energy stored in the spring is conveyed to the kinetic energy of the mass – a complete transfer occurring when the mass has risen to the zero equilibrium position before it was pulled down.

In the electrical case the reason for the ring up in voltage is not so obvious, but in both cases, the resultant action has to adhere to the First Law of Thermodynamics, the conservation of energy. In our circuit, the current builds up when the transistor is on and di / dt is positive. All is well until the transistor shuts off and di / dt becomes a large negative value because of the slope of the current VS time falling almost instantly to zero. Now the voltage across the inductor becomes very negative. The terminal of the inductor tha is connected to the battery becomes negative and terminal of the inductor connected to the transistor DRAIN jumps positive. Because the inductor positive voltage is now in series with the battery voltage, we get a gain in voltage where D1 couples current to the output. If diode D1 were removed from the circuit, the voltage on the DRAIN may rise up hundreds of volts, limited only by the breakdown of the switching transistor.

Current from the inductor and battery couples energy out of both components to charge capacitor C1 and provide power the load. When current stops flowing through D1 the inductor current ramps down to zero. To see this in action, let's use a SPICE simulation on a simple RCC that converts 9 volts to 30 volts in order to drive a string of UV-LEDs as a load (not shown). We shall use a TRANSIENT analysis that lasts 1 second looking at the switching voltage of the transistor drain (TPv1) as well as the inductor current (TPi3).

Design of a simple RCC: (Ultra-violet LED driver)

Specifications:	Vin:	9V (battery)
	Vout:	30 Volts
	Power:	3 Watts (will be used to drive a string of UV LEDs)
	Frequency:	20 kHz

To start our analysis, we will try an inductor value of 100 uH and an on-time, of 5 microseconds (τon = 5 uS). We will set the repetition frequency of operation to be 20KHz because it is just above the audible range. Having an on-time of only 5 uS represents a duty cycle of 10% and running a SPICE analysis will tell us what output voltage we should expect to get from this RCC. Selecting an inductance of 100 uH gives us our first cut at the design. We shall analyze the 3 Watt full load case first because if that design works, the circuit can certainly do the No-Load case as well – the on-time would just have to be made shorter or he output will rise well above 30 volts. You can assume for now that our switch is driven from a pulse-width modulator circuit where we can manually select the on-time of the switch to any value we want. Later on we can close the loop and regulate the output against any load or line changes.

Figure 3.4 shows the switch waveform (Voltage Test Point 1) in red and the inductor current ramps in green when we drive the switch with an on-time, (τ_{ON}), of only 5 uS. The run time of this SPICE simulation is 0.1 seconds and we have selected to show the waveforms at the end of the run. Notice how the voltage rises up (to 14.4V) for a short time certainly above the 9 volt input battery voltage. This time is called the ring-time (τring) and as shown in Figure 3.4, this is 8.9 uS long during which current is flowing from the battery, through inductor L1 and through D1 to charge the output capacitor C1 and provide power to the load. Notice during τring, the current ramps down to zero. Although we want this converter to make 30 volts from the 9 volt input it is not doing as well as we want because the output (blue) is only 14.5 volts. In this simulation, made with LT SPICE, the current peak is seen to be 0.564 Amperes.

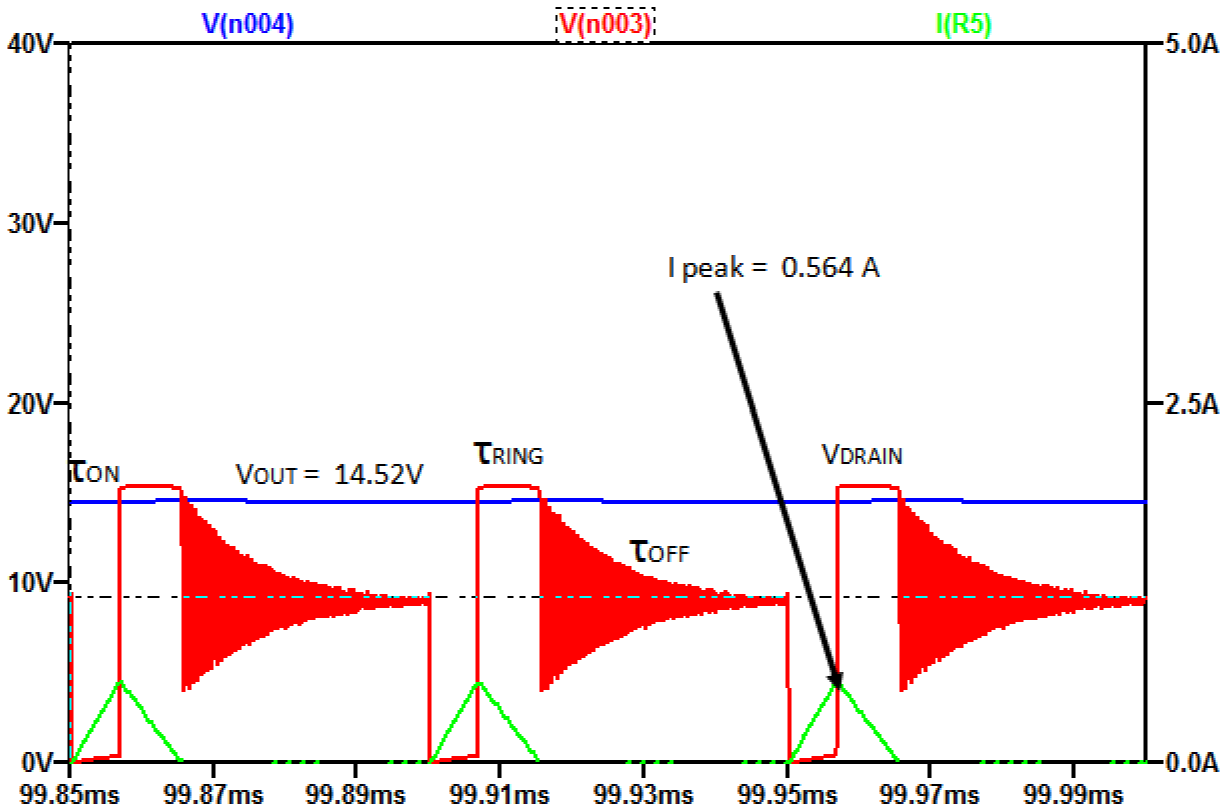


Figure 3.4: LT SPICE simulation of 9 to 30 volt RCC with τon = 5 uS
Switching waveforms: Drain (red), Vout (blue) inductor current (green)

Notice that in the DRAIN waveform, there is a τon of 5 uS, a τring of 8.9 uS, but there is also an off-time as well, τoff, of about 34 uS. This occurs when the inductor current stops flowing and the DRAIN voltage simply converges to the input battery voltage (9V). The high frequency ringing during τoff is due to the small value capacitors making up the model of the transistor switch we used (Si9420), working in conjunction with the 100 µH inductor of our circuit.

How do we get the output higher because we need 30V? One thing we can do is to increase the on-time of the transistor switch, τon, to store more energy in the magnetic field. This is easy to do and will obviously increase the current ramp peak value. Figure 3.5 shows the switch waveform for an on-time of 10 microseconds. Here, the switch voltage now rings up to 21.4

volts which is better than before but still not good enough to use in our 30 volt converter. The current ramp peak value has increased to 0.931 Amperes, the ring time τring has dropped to 7.1 uS and the off time has decreased to 31.2 uS.

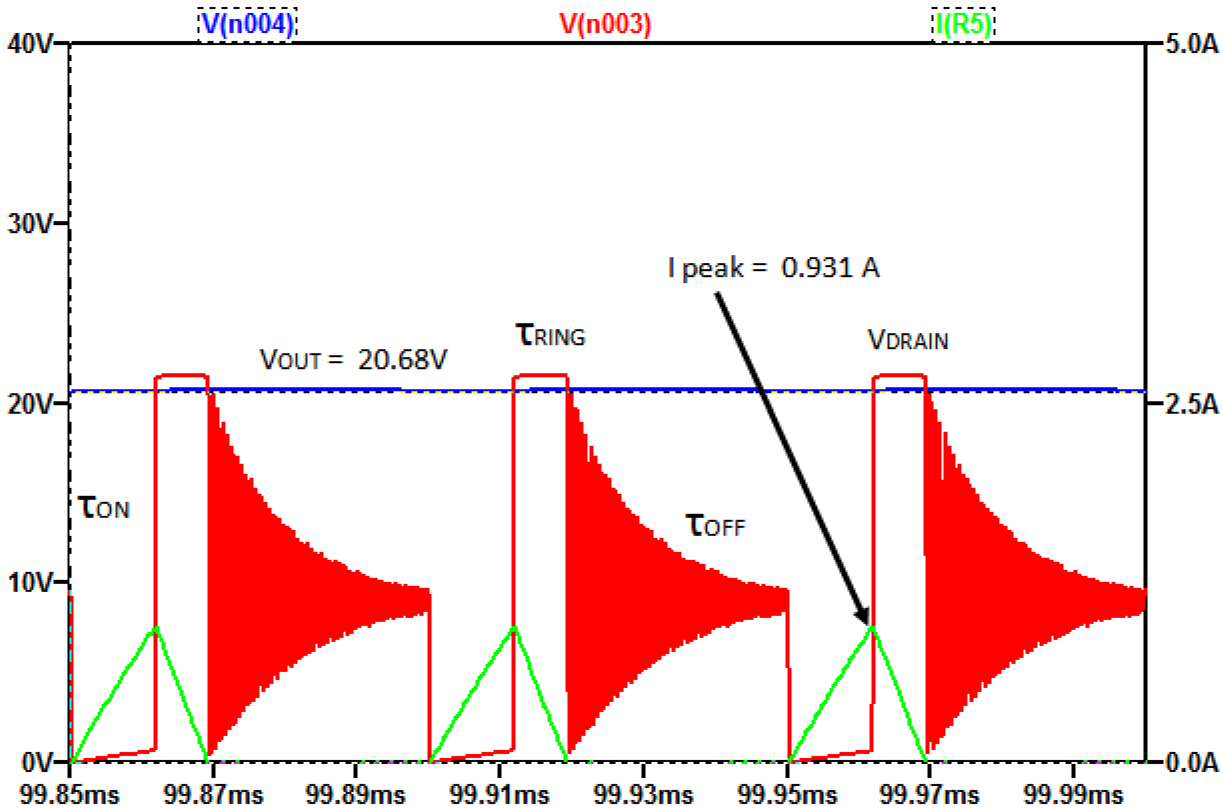


Figure 3.5: LT SPICE simulation of 9 to 30 volt RCC with τon = 10 uS
Switching waveforms: Drain (red), Vout (blue) inductor current (green)

Increasing the on-time further to 18.3 uS gives us what we are looking for, an output of 30 VDC as shown in Figure 3.6 working into a full load of 300 Ohms. The peak of the current ramp has gone up as expected and is now 1.49 Amperes..

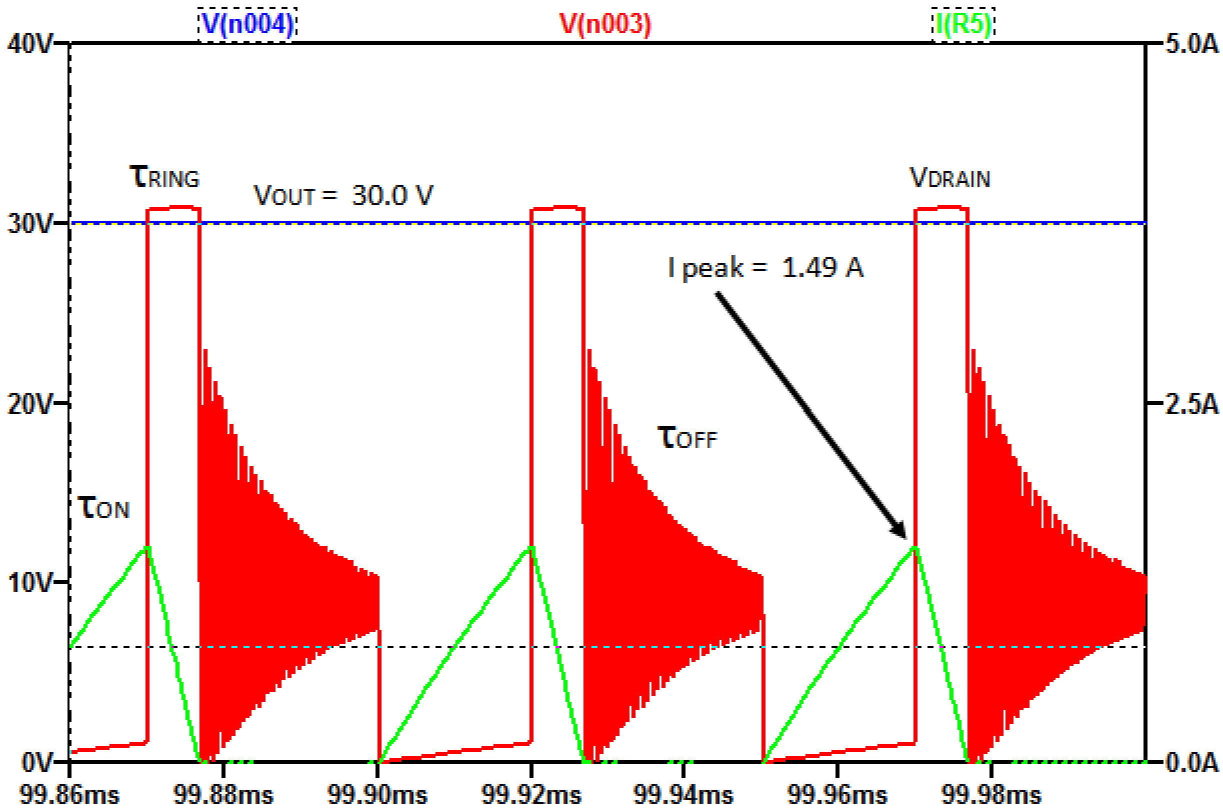


Figure 3.6: LT SPICE simulation of 9 to 30 volt RCC with τon = 18.27 uS
Switching waveforms: Drain (red), Vout (blue) inductor current (green)

Obviously, you can use this hunt and peck method to find the correct duty cycle that will give you full output under load but this will eventually become quite tedious. Also, the inductor that we selected was 100 µH. Is there a different value for our three Watt design that will give higher efficiency? Would a different choke, say 200 µH work better? How about 50 uH? The answer to this question is to look at the efficiency of the converter. A three Watt converter would draw an average of 0.334 Amperes from a 9 volt battery if the efficiency was 100% - which of course it is not. Running an LT SPICE analysis on the circuit with an on-time of 18.3 uS and looking at the average input current from the 9 volt be\battery can tell us a great deal of how we are doing. To get the average input current, select the current from the battery plot header and then, pointing to the color of that plot press the control key. Doing so reveals the average input current is 0.421 Amperes from the battery averaged over a hundred cycles. This looks reasonable.

From this the efficiency of our converter is (taking the 30.1V output into account):

$$\eta = P_{out} / P_{in} = 3 / (9)(0.421) = 79.1\ \% \qquad (3\text{-}4)$$

This is somewhat on the low side especially for a simple converter that doesn't seem to have many apparent losses. The question may be asked: where are we burning up the energy? Let's take a look. Because there is a lot of off-time when current is not flowing from the battery, we have to compensate with high peak currents to get the output power of 3 Watts. Working at 20kHz and having a 50 uS total period, our switch only operates with a 18.3 uS on-time. When current peaks are high we dissipate more energy in the transistor switch and whatever resistive losses are in the circuit, especially the 1.0 Ohm resistor we have inserted to simulate Ohmic losses. If we made the off-time equal to zero and the on-time greater, we could reduce the peak currents into our converter and perhaps pick up a few percent more in efficiency.

The following equation gives an inductor value that forces the off-time τ_{off} to be zero. That means the transistor switch will turn on again as soon as D1 is finished conducting. Let's call this inductance the inductance for zero off time: L_{ZOT}. The following equation will yield this value of inductor. We will present it now but derive it later on:

$$L_{ZOT} = R_{load}\, V_{bat}^2\, (V_{out} + V_D - V_{bat}) \,/\, 2\, f\, V_{out}^2\, (V_{out} + V_D) \qquad (3\text{-}5)$$

For our converter example, we will insert the following:

Rload	=	300 Ohms
Vbat	=	9 V
Vout	=	30 V
V_D	=	0.80 V
f	=	20,000 Hz

where we have used the dynamic voltage drop of D1, 0.80 volts, taken from the data of the first chart Figure 3.4. This gives us an inductance value for zero off-time, L_{ZOT}, of:

$$L_{ZOT} = 478 \text{ uH}$$

Figure 3.7 shows us the operation of such a circuit. The current peak is now 0.679 Amperes instead of 1.52 Amperes and the off-time has disappeared.

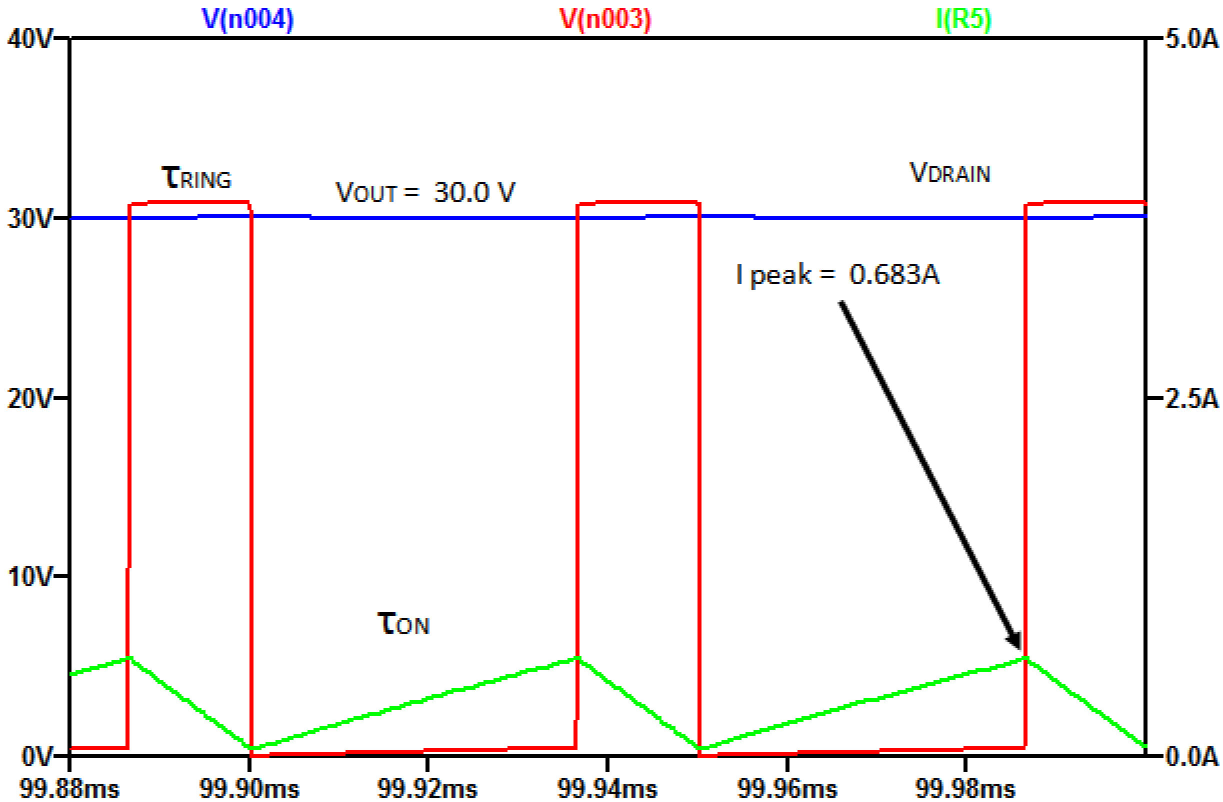


Figure 3.7: Zero off time, L = 478 uH with τon = 34.6 uS

Our LT SPICE analysis for the circuit, Figure 3.7, shows the average input current is 0.374 Amperes, even though the peak current is 0.683 Amperes. Usually a triangle ramp yields an average half of its peak value but due to slight bending of the current curve, the average current is a little higher. Let's check the efficiency of this circuit:

$$\eta = \text{Pout / Pin} = 3.00 / (9)(0.374) = 89.1\ \% \quad (3\text{-}6)$$

which is a significant improvement from the 79.1% we had before in the circuit with an excessive off time. The only difference is that the inductor has increased from 100 µH to 478 µH as well as the much longer on time, from 18.2 µS to 34.6 µS. Can we increase the efficiency even more? The answer is yes!

By not letting the switch current ever go to zero and forcing it to have a "DC level", we can actually increase the efficiency further. By operating in what is called: "continuous current mode" and preventing the current from ever decreasing to zero by selecting the proper inductance and on time value we can have a system that has the highest efficiency and lowest switch peak current possible along with several other distinct advantages as will be explained. As mentioned, this is achieved by always having a DC current level with a ramping AC component of current riding on top. As most designs do, we will limit this to AC component to +/- 10%. There are three good reasons for doing this: First, this allows peak current from the battery to be much smaller and making inductor selection with regard to magnetic material saturation less complex. Secondly, because the peak currents are lower, there will be less dissipation in both the transistor switch and our output diode. Third, because of the DC component of input current, there will be less EMI noise generated in the input circuitry because input currents there will not go to zero.

To achieve "continuous current operation" we use a simple relationship that exists for such designs. If the current ramps are *not allowed to go to zero* the equation for the voltage step-up in a RCC is given by:

$$V_{out} + V_D = V_{in} / (1 - \tau_{on} f) \qquad (3\text{-}7)$$

where f is the frequency of operation, a parameter usually known at the start of the design and V_D is the diode drop as before. By rearranging this we can find the required on-time to be:

$$\tau_{on} = (1/f)\,(V_{out} + V_D - V_{in}) / (V_{out} + V_D) \qquad (3\text{-}8)$$

We will use equation (3-8) for now and derive it later on in this chapter. Notice that the value of inductor is not even involved in this equation as long as we work in the continuous current mode because there is no off time. The relationship only involves voltages and duty cycle and can yield us the necessary on-time for full load operation.

For our 9 to 30 volt converter, we can work this equation finding that the on-time required is 35.4 uS for the 30 volt output with a 3 Watt load when working in the continuous current mode. We have assumed a dynamic output diode drop of 0.8 volts again as typical for a fast recovery device. Just for the record, if we had used a Schottky diode we would have had less voltage drop, perhaps as low as 0.4 volts but unfortunately we would have had to contend with the large diode capacitance in our SPICE circuit. For now, using a simple PN junction diode is easier to work with but you can substitute a Schottky diode later on in your analysis. After finding the on-time of the switch we can determine an inductor value that will allow us to work in this mode and this is easy to do using the next section.

The average input current is related to the output load divided by the efficiency: Assuming a power transfer efficiency of at least 94% (because we can get 89.1% in the non-continuous mode), we can find the input power to the actual converter:

Power in: Power out / η = 3 / 0.94 = 3.19 W (3-9)

Since the input battery voltage is 9V, this indicates an average DC input current of 0.354 Amperes taking efficiency into account. For reasons that will be explained shortly, a continuous current mode RCC is typically operated with its AC peak to peak current component +10% and minus 10% of the average input current. That is, riding on the DC input current component will be an AC current waveform adding and subtracting 10%. As we follow the development of our design, with the DC component of input current of 0.354 Amperes, we calculate what we want the current to be going plus and minus 10%:

Input current to the converter: 0.354 +/- 0.0354 (3-10)

This means, as the switch closes the current will rise 10% along the current ramp to 0.3894 Amperes during τon and drop as a ramp 10% below the average DC input current to 0.3186 Amperes after the switch opens and the diode begins conduction, that is, during a time $\tau_{ring.}$

Using the inductor relationship of equation 3-1, we can find the value of inductance that allows this change to happen when a voltage of 9 volts is placed across it for an on time of 35.4 µS
:

Vbat	=	L di/dt
9	=	L (0.3894 - 0.3186)/35.4 uS
L	=	4,500 uH

Now, this inductance is much larger than in our earlier discontinuous current designs and must be able to handle at least 0.3894 Amperes without saturating the magnetic material. Figure 3.8 shows the result of our LT SPICE analysis.

Figure 3.8 shows that the ring-up is 30.65 volts and the converter make a little under 30 volts. LT SPICE simulation gives us an input current average of: 0.358 Amperes, yielding an efficiency value of:

η = Pout / Pin = 2.97 / (9)(0.358) = 92 % (3-11) a tad bit higher than the previous example. Notice the DC component of input current and the AC component riding on it. The only drawback of operating in the continuous current mode is that now the inductor value is nearly ten times larger for a gain in efficiency of only a few percent. Having a continuous input current does help the efficiency and reduces EMI generated by the input section because the current never goes to zero.

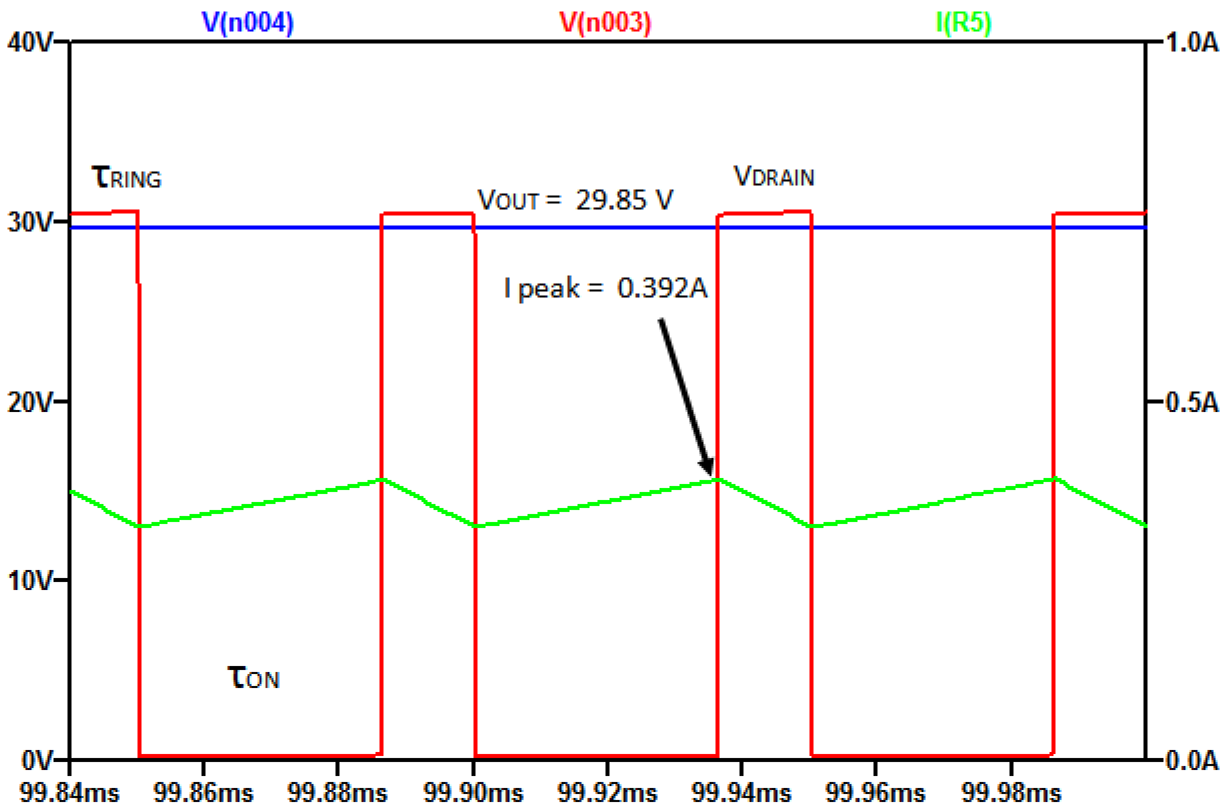


Figure 3.8: Continuous current mode operation on-time = 35.4 uS

Looking back on equation (3-7), notice that the step-up of an RCC, when operating in the continuous current mode seems to be independent of both load resistance and inductance value. Although they are all related, we can plot the step up of an RCC when operating in the continuous current mode, neglecting semiconductor voltage drops from switch and diode.

For our 9 to 30 volt converter a gain of 3.33 was needed, indicating a duty cycle (D/C) of 70% and this is what we basically have:

D/C = time on of switch / full period

D/C = 35.4 µS / 50 µS = 70.8%

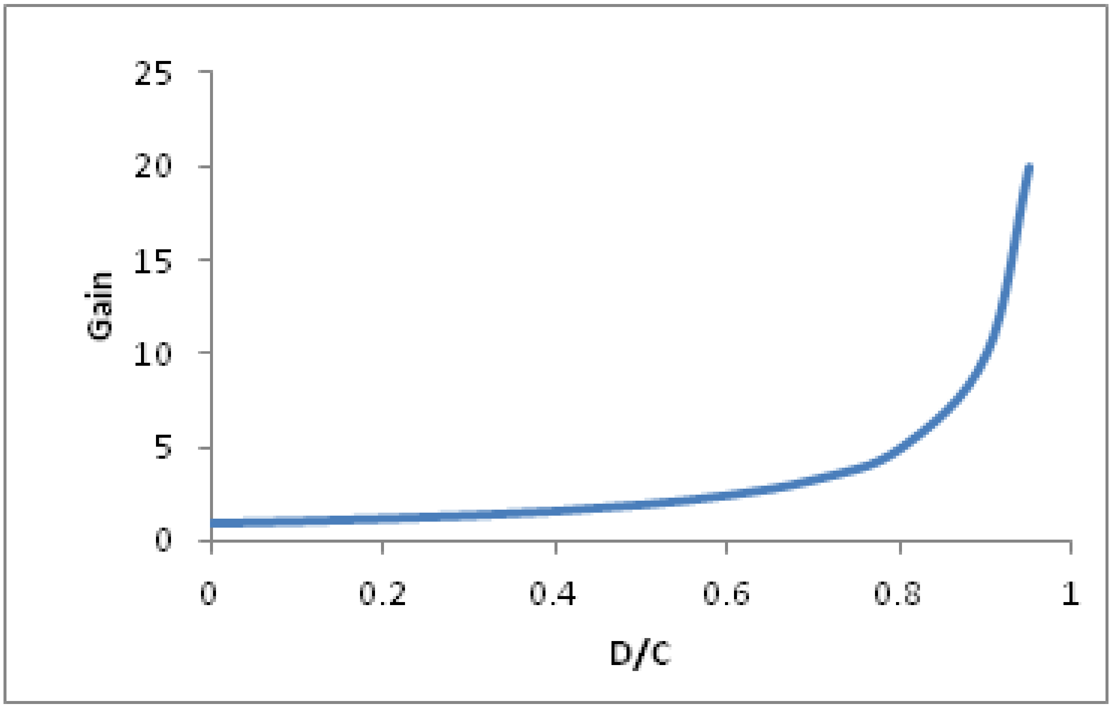


Figure 3-9: Step-up of RCC when in the *continuous current mode* of operation

For practical reasons, the duty cycle should be limited to 95% maximum when operating in the continuous current mode because ring-up voltages take some time to reach their full value (due to component and stray capacitance) and the turn-on of the switch starts to encroach on the rising voltage as duty cycles increase towards 100%. If higher step-ups are required it is best to use a transformer (Chapter 4). Also, if the circuit is controlled by a PWM, the pulse width on-time must never be allowed to go to 100% because the output will drop to zero and the circuit will latch up – burning up the switch and inductor in the process. .

Going back to Figure 3.7, the circuit for zero off time, the increasing current ramp during τon does not have the same slope as the decreasing ramp during τring. This is because there are different voltages across the inductor at those two different times. However, since we are dealing with one inductor and the same peak current, both are related. The ON-time equation deals with the input battery voltage (Vin): (from Equation 3-2):

,

$$Vin = L\, di/dt = L\, i\, peak / \tau\, on \qquad (3\text{-}2)$$

and the ring-time equation deals with the output, diode drop minus the battery voltage Vin:

$$Vout + V_D - Vin =, \quad L\, di/dt = L\, i\, peak / \tau\, ring \qquad (3\text{-}12)$$

since the frequency of operation , f, is simply the sum of τ on and τ ring,

$$1 / f = \tau\, on + \tau\, ring \qquad (3\text{-}13)$$

we have, knowing that both L and i peak are the same:

$$(Vin)(\tau\, on) = (Vout + V_D - Vin)(\tau\, ring) \qquad (3\text{-}14)$$

we can rearrange the above to:

$$Vout + V_D = Vin\, (\tau\, on + \tau\, ring) / (\tau\, ring)$$

because $$1/f = (\tau\, on + \tau\, ring)$$

$$Vout + V_D = Vin\, (1/f\, \tau\, ring)$$

$$Vout = Vin\, (1/f)(1/(1/f - \tau\, on)) - V_D$$

Finally: $$Vout = Vin\, (1/(1 - f\, \tau\, on)) - V_D$$

Thus we have derived equation 3-7.

To determine the value of inductor for the continuous current mode, L_{CCM}, the following equation is derived from the above relationships and may come in handy:

$$L_{CCM} = \eta\, Vin^2\, \tau on / 0.2\, Pout \qquad (3\text{-}15)$$

where η is the efficiency of the circuit (assume a value and test it out via simulation).

No Load

What happens if we remove the load from our converter while it is running? If that happens, the PWM had better correct the on-time pulse width or the output voltage will rise destructively upwards. Our driven converter must be inside of a closed loop system for the output to regulate. Figure 3.10 shows the waveforms if we increase the load from 300 Ohms to 300,000 Ohms and the on time stayed the same as before 35.4 uS. Because out load resistor is now thousand times higher, the time to reach a steady state value will increase for our analysis as well. In the past we were able to run a transient SPICE analysis in 0.1 seconds and yield meaningful date. That was because the time constant of our output circuit, 50 uF and 300 Ohms was 15 milliseconds. Now, with a load of 300K ohms the time constant is 15 seconds meaning you may have to run the analysis for many minutes to get to steady state. On trick would be to lower the capacitor value down to, say 0.1 uF and run the analysis for a longer time, say 1 second. Because we have lowered the load the ripple, with this amount of capacitance will not interfere in our analysis.

Figure 3.10 shows the SPICE simulation output for our circuit at no load. As you can see the output voltage has risen to 227 volts, which is *above the breakdown voltage of the 200V transistor* we are using and the duty cycle must be throttled back to prevent this, that is the PWM must be operated in a closed feedback loop.

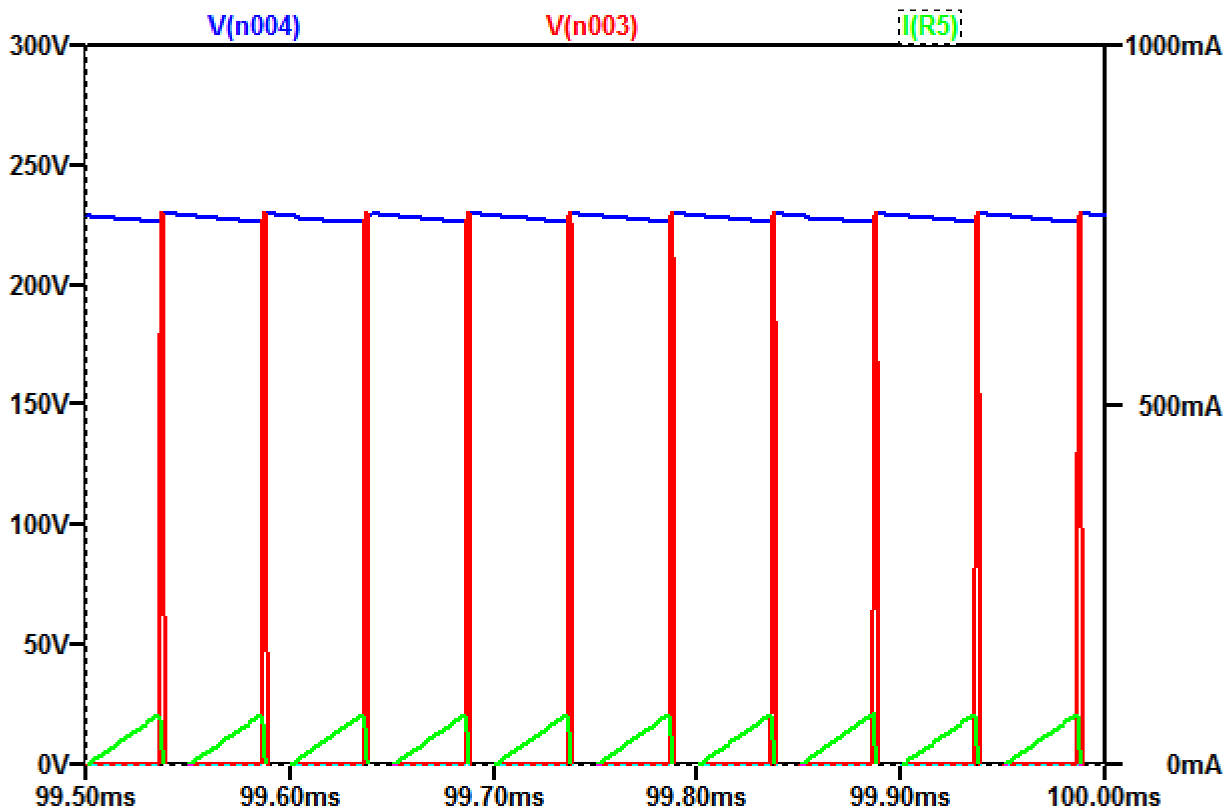


Figure 3.10: Drain waveform for Load = 300K Ohm, and on-time of 35.4 µS

At no load condition, the on-time would have to lower to 1.0 µS for the output to remain at 30 VDC with our large 4,500 µH inductor. Many designers include a 10% pre-load to stabilize the feedback loop because tiny pulse widths can be notorious for promoting feedback loop instabilities due to noise jitter. Putting a 10% preload however lowers down our maximum efficiency to only 90% and is retrograde to the gains we sought by going to the continuous current mode ringing choke converter. Figure 3.11 shows the waveforms that will happen when our PWM has adjusted the output back to 30.0V without a 10% preload.

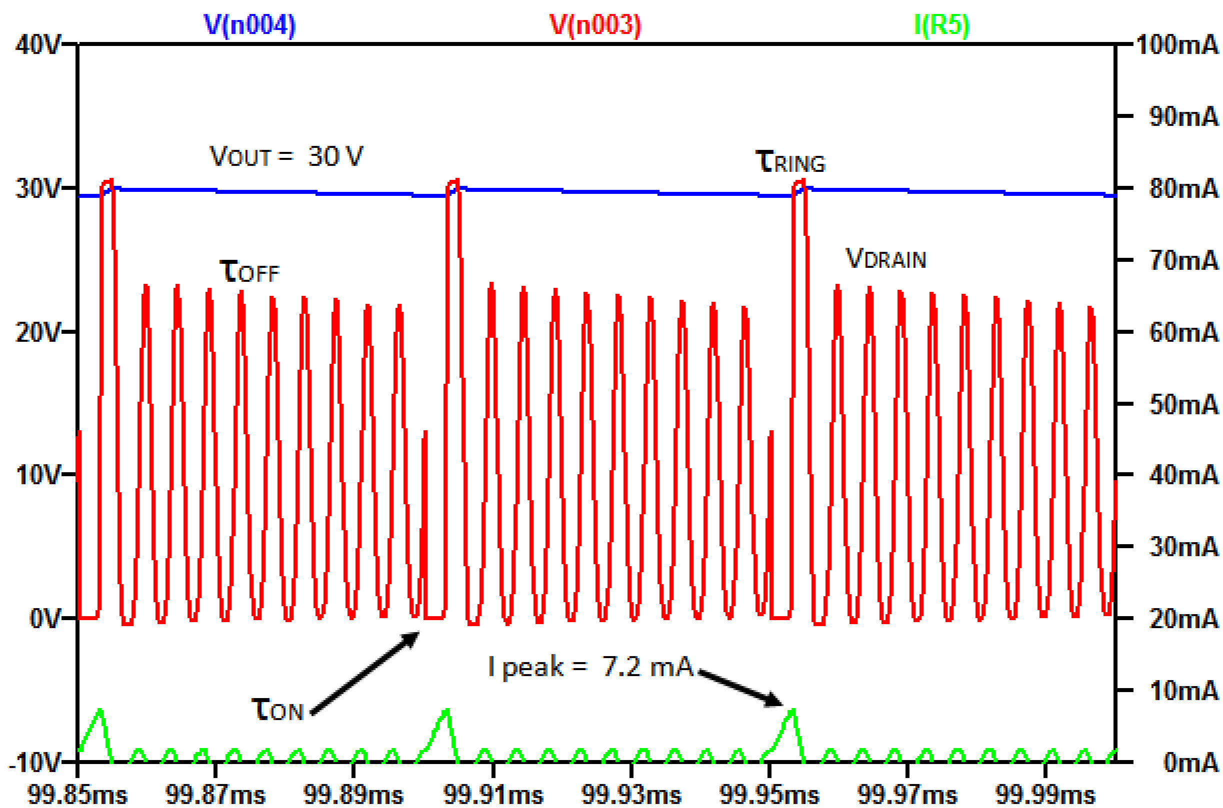


Figure 3.11: Converter producing 30V output with 1 µS on time No Load

Notice the large amount of ringing that now occurs when an off time suddenly appears. The large inductance resonates with the capacitance in the transistor model and will appear as shown. This is a source of EMI an should be squelched by appropriate snubbers. Each ring consumes energy from the battery to charge and discharge the "hidden" capacitors.

Rules for a continuous current mode RCC are as follows:

1. Using equation 3-8 determine the on-time needed to make full load
2. Using equation 3.15 determine the inductor value.
3. Run a transient SPICE analysis at full load and reasonable run time (1 sec).
4. Check the output voltage – make sure it can provide full load at low line.
5. Using the average of the simulated input current, check the efficiency.

Power Transfer Efficiency

In our SPICE simulations we have run efficiency determinations using the values of input current that our program gives. We can actually calculate the efficiency of our design because there are only three locations of power dissipation within the RCC circuit:

1. Switching transistor losses (R_{DS} if using a FET) + overlap
2. Resistive losses due to any series resistance
3. Output diode conduction losses

The efficiency is depends upon the load:

$$\eta = Pout/Pin \qquad (3\text{-}15)$$

For the converter we have been designing we can calculate what efficiency we should be getting. Let's look closely at each loss mechanism.

Switch Loss: Most MOSFET data sheets list the R_{DS} for their device. By taking the average current during the on-time, usually half of the peak current value since we are dealing with a ramp function, we can easily list the loss due to the switching transistor.

$$P\ switch_{RDS} = (i_{AVG\ INPUT}{}^2\ R_{DS})(\tau on)(f)$$

where i_{AVG} is the average current during the on-time interval and $(\tau on)(f)$ the ratio of time the transistor dissipates this power. The transistor we used, a Si9420DY hasd a listed R_{DS} of 1 Ohm. In our converter running at full load the average input current was seen in our simulation as 0.358 A, with a τ_{on} of 35.4 μS, this loss is:

$$P\ switch_{RDS} = 0.091\ \text{Watts.}$$

Another power loss mechanism associated with the transistor switch occurs when voltage and current overlap. Normally when the switch turns on the voltage is usually at a high value. Current
starts to flow as a linear ramp and energy is stored in the magnetic field of the transformer. When the switch turns off and the voltage rises quickly upwards, it occurs at the highest point in the current ramp. In an ideal world the transistor would shut off instantly, in nanoseconds, but due to charge carriers flowing through resistive crystal structures this doesn't happen exactly, many FET's require at least a microsecond to turn off during which the voltage and current overlap. We can come up with a simple relationship that gives the instantaneous power burned away in the transistor die as a function of time:

$$P\ switch\ overlap\ (t) = Vmax\ i\ peak\ (t / \tau_{fall} - t^2/\tau_{fall}^2)$$

which shows a parabolic relationship as a function of time.
To integrate this over the overlap period, that is from $t = 0$ to $t = \tau_{fall}$, we get the total energy loss as:

$$Energy\ loss = 0.166\ Vmax\ i\ peak\ \tau_{fall}$$

Because this happens at the operational frequency, we have a power loss:

$$P\ switch\ overlap\ (t) = 0.166\ Vmax\ i\ peak\ \tau_{fall}\ f$$

In our last converter design Vmax was seen as 30.65V and the peak current was seen as 0.393 A. We will use a value of 1 μS for τ_{fall} a reasonable value that is commonly used, especially when operating at only 20kHz. This gives the following loss as:

.

$$P_{switch\ overlap} = 0.04 \text{ Watts}$$

Ohmic Loss: It is a good thing to take into account any series resistance in the circuit. This includes the resistance of the inductor and copper traces on the printed circuit board.

$$P_{Ohmic} = (I_{AV\ INPUT}{}^2 R_{series})(\tau_{on}\ f)$$

We did this by assuming the choke and its interconnections had a resistance of 1.0 Ohm. The loss for this mechanism is:

$$P_{Ohmic} = 0.091 \text{ Watts}$$

exactly the same as the R_{DS} of the transistor only because the resistance is the same value which is usually is not.

Diode Loss: When the rectifier diode is forward biased, the voltage drop in its PN junction leads to another power loss that depends upon the output current ramp:

$$P_{diode} = (V_D)(I(t))$$

The instantaneous power dissipation in the forward biased diode as it conducts the linearly decreasing current into the load and output capacitor is given by:

$$P_{diode}(t) = (V_D)(i_{peak}) - ((V_D)(i_{peak})\ t / \tau_{ring}$$

another parabolic relationship. This integrated over the cycle ($t = 0$ to $t = \tau_{ring}$)

when the current is flowing through the diode yields an energy loss of:

$$\text{Energy loss during the cycle:} = \tfrac{1}{2}(V_D)(i_{peak})(\tau_{ring})$$

But because this is happening 20,000 times a second, we need to take that into account by multiplying by the frequency:

$$P_{diode} = \frac{1}{2} (V_D)(I_{peak})(\tau_{ring})(f)$$

In our converter with τ_{ring} from Figure 3.8, of 14.6 µS (just 50 µS – τ_{on}), we have the diode loss as:

$$P_{diode} = 0.045 \text{ Watts}$$

Summing up all of these losses:

P switch$_{RDS}$	=	0.091 Watts.
P switch overlap	=	0.04 Watts
P Ohmic	=	0.091 Watts
Pdiode	=	0.045 Watts
Total Power loss	=	0.267 Watts

For an output power of 3 Watts, the input must be at least 3.267 yielding an efficiency of operation of:

$$\eta = \text{Power out / Power in} = 3/3.267 = 91.8\ \%$$

in close agreement with our simulation equation 3-11 that gave 92%. It looks like the greatest loss is due to the 1.0 Ohm we inserted into the schematic to take account of the inductor resistance and the lowest for this converter seems to be the transistor switch overlap. This may not always be the case because higher input voltages may cause higher overlap dissipation.

If a bipolar transistor is used you will have to insert an equivalent R_{DS} value you can devise from the device saturation voltage and peak current. This should not be too hard to do.

Obviously, these four dissipations have many terms but it gives a reasonable view of where the circuit is losing power. Sometimes other equations can be found. Pressman, in his classic book: *Switching and Linear Power Supply Power Conversion Design* (1977) uses the simplified equation:

$$\eta = Vin / Vin + 2$$

which is certainly easier to use but gives a value of:

$$\eta = 9 / 9 + 2 = 81.8\%$$

for our converter example. This is a lot lower but was written way before power MOSFETs were available.

We shall now work several examples of the ringing choke converter and calculate the important parameters and check them with SPICE simulations to see how they correspond.

Design examples:

Example 1:
A power converter that steps up 12 volts to 50 volts at 2 Amperes is required to run a bank of ten white ultra-bright LEDs placed in series. This will be used on a Coast Guard boat as a spot light so the battery voltage may drop as low as 11.0 VDC and be as high as 13.6 VDC when the engine is running. .

We will use the RCC topology running at 100 kHz in the continuous current mode. From equation 3-8 we can determine the on-time of our converter.

$$Vout + V_D = Vin / (1 - \tau on\ f) \qquad (3\text{-}8)$$

Since this is a 100 Watt converter, it will be advantageous to use a Schottky diode that drops as least amount of voltage as possible. High voltage Schottky diodes with a drop of only 0.5 volts at 2 Amperes are available. We will neglect the capacitance of the Schottky diode at first but in a SPICE analysis we will include it. From equation 3-8 the on time is calculated:

$$\tau_{on} = 7.82\ \text{uS}$$

where the low line voltage of 11.0 volts was used. This will give us the longest on time needed. From this, working in the continuous current mode, we can calculate the inductance to be:

$$L_{CCM} = \eta\, V_{bat}^2 \tau_{on} / 0.2\, P_{out} \qquad (3\text{-}14)$$

$$L_{CCM} = 42.6\ \mu\text{H}$$

Which, if we assume an efficiency of 90%, and an input voltage of low line 11.0V. As a capacitor, we will use a value of 20 uF giving us a ripple voltage of 1 Vpp. Because the input voltage is only 12 volts, every volt of input counts, especially when we will be drawing in over ten Amperes. The series resistor in our analysis we used before, now has to be reduced down to 0.1 Ohms instead of the 1.0 Ohm we had in previous 3 Watt 30 volt converter. In addition, we have to replace our switching transistor, with something more hefty due to the increase in current as well. We will select the IRF240, a TO-247 device known for it's maximum current of 50 Amperes. Figure 3.12 is the schematic. The output diode is a Schottky type rated 100V at 10 Amperes. Because we are trying to work at 100kHz, our gate drive must be able to switch the gate voltage much faster than we allowed for the earlier example. This time we will lower the rise and fall time of the gate pulse to 0.1 µS. We should also reduce the 100 Ohm gate resistor to a lower value, say 10 Ohms.

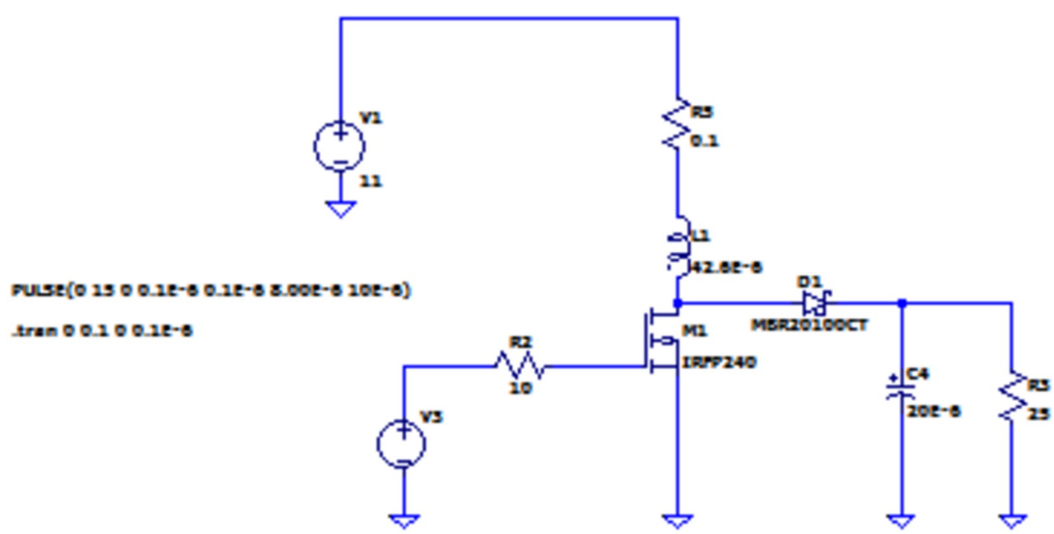


Figure 3.13 shows the waveforms when operating at 11 volts input. Here ring-up voltage only reaches 48 volts during the simulation, shy 2 volts. This can be corrected by either a slightly smaller inductor or longer on time/ The current ramps for this simulation indicate a peak current through the 42.6 uH inductor of 11.3 Amperes. The average input current was seen as 10.49 Amperes. This indicates an average efficiency of 92.2 / 115.4 = 80.0%, somewhat lower than desired.

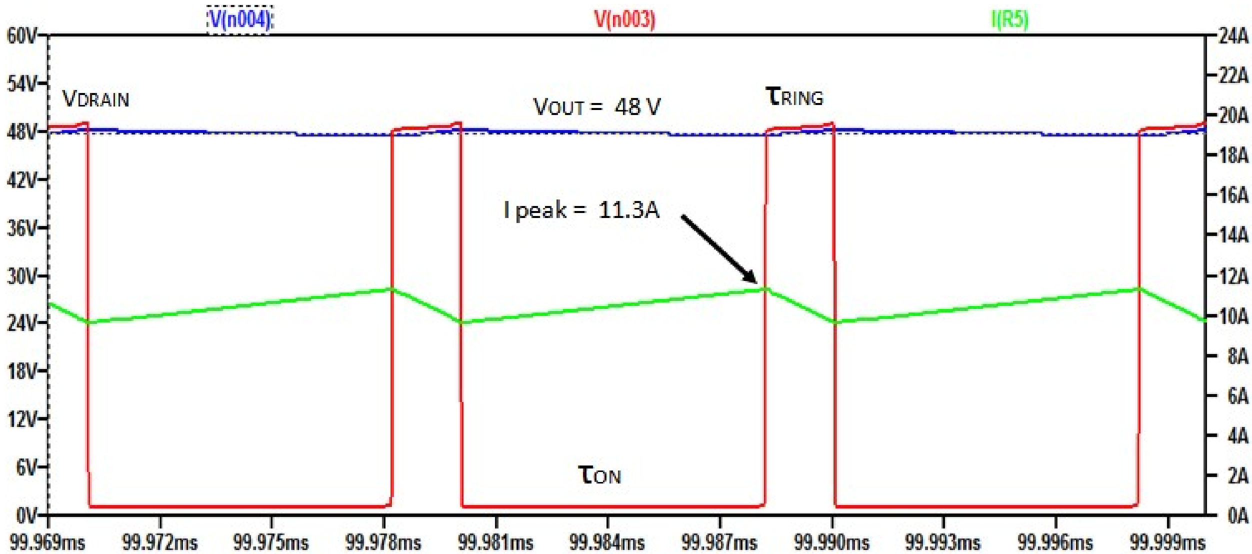


Figure 3.13 12 to 50 volt RCC converter Drain (red) Vout (blue) Iin (green)

calculating losses from the earlier methods discussed we find, with the following data:

Vin	11.0 volt
Vout	48.0
R_{DS}	0.18 Ohm
$i_{AVG\ INPUT}$	10.49 Amperes
τon	7.82 µS
f	100 kHz
Vmax	48.5
τfall	0.1 µS (assume)
i peak	11.34 Amperes
Rseries	0.1 Ohm
V_D	0.5
τring	2.18 µS

Power loss equations:

P_{RDS} loss = $(i_{AVG\ INPUT}{}^2 R_{DS})(\tau on)(f)$

P switch overlap (t) = $0.166\ V_{max}\ i\ peak\ \tau_{fall}\ f$

P Ohmic = $(I_{AV\ INPUT}{}^2\ R_{series})(\tau\ on\ f)$

Pdiode = $\frac{1}{2}(V_D)(I\ peak)(\tau ring)(f)$

Calculated losses using the converter data:

P switch$_{RDS}$	=	15.46 Watts.
P switch overlap	=	0.91 Watts
P Ohmic	=	8.61 Watts
Pdiode	=	0.62 Watts
--		
Total Power loss	=	25.60 Watts

Giving us a power efficiency of:

$$\eta = \text{Power out / Power in} = 92.16/(92.16 + 25.60) = 78.2\ \%$$

Remembering that the output was less than 100 Watts this is close to our simulation of 80%. As can be seen in the above summary of losses the highest loss is due to the RDS of the switching transistor and second by the losses associated with the added series resistance. Because we have assumed a limited switch overlap of current and voltage dueing the switch shut-off, this dosen't seem to add too much but we must make sure that the transistor fall time is indeed 0.1 µS. Because we are using a Schottky diode, the power dissipation there seems low compared to the other components. It also has to be mentioned that the resistive Ohmic loss is that just due to pure DC resistance and does not take into account any AC resistive losses due to skin effect or proximity effect which may, if the choke is not wound correctly, increase this loss factor an order of magnitude. This is discussed in Chapter 5 under Magnetics.

Example 2:
A power converter that steps up 170 volts to 600 volts at 1 mA is required to run a Geiger Counter tube in its plateau region. Because this converter is to be run directly off the rectified

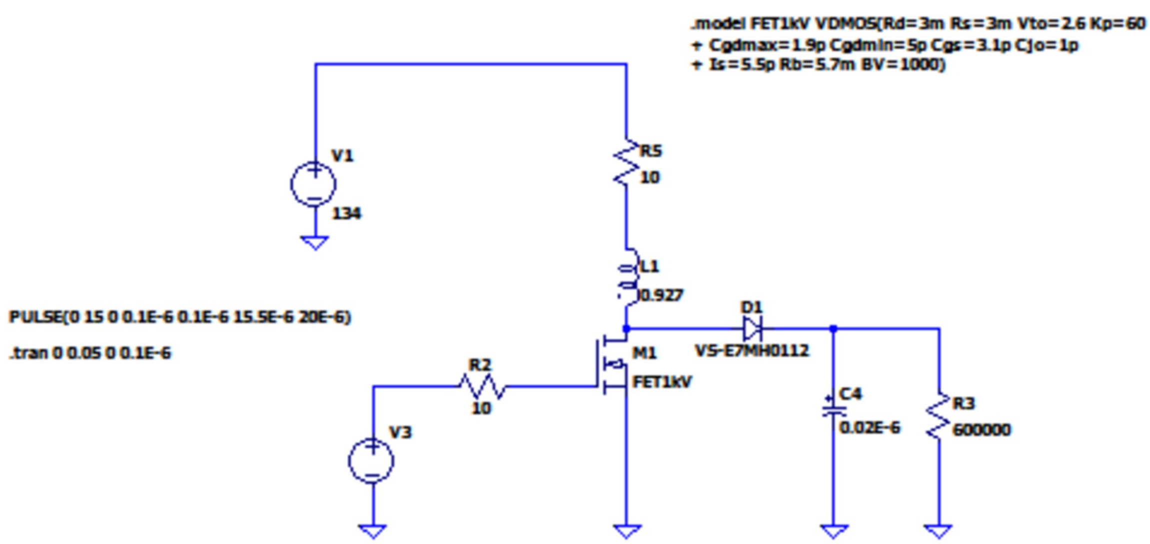
120 VAC line great care must be taken to isolate the voltages and returns from ground. This Geiger Counter to be constructed as a simple wall plug-in module to be just placed in a convenient outlet and used as a SCRAM device in radiation environments such as a nuclear power plant or university lab housing nuclear equipment. The electronics will be mounted in a plastic sealed box with only a audio alert siren mounted on the bottom through the plastic. The step up to 600 volts can easily be achieved by using a 1 kV MOSFET as the switching element along with a 1kV fast recovery diode. The specification says that the AC line may drop as low as 95 VAC during brown outs and the converter still needs to work even during brown outs for safety reasons. This means that the input DC voltage after rectification may drop to 134 VDC at low line and the converter must still make 600 volts. Because the specification for output power is only 0.6 Watts this should not be much of a problem design. We will use the RCC topology running at 50 kHz. From equation 3-8, we can find the on-time of our converter.

$$V_{out} + V_D = V_{in} / (1 - \tau_{on} f) \qquad (3\text{-}8)$$

Since this is a 100 kHz, 0.6 Watt converter, the high voltage diode must be a fast recovery type, and we will assume a V_D of 0.8 volts. Using the low line voltage of 134 V_{IN}, we find:

$$\tau_{on} = 7.75 \text{ uS}$$

This will give us the longest on time needed. From equation 3-14, we can calculate the required inductor value, and because our Vin is a rather high voltage we will only assume an efficiency of 80%. The series resistor for this choke is mde to be 10 Ohms.



The drain waveforms are shown in Figure 3-15 along with the inductor current waveform. LT SPICE shows that the average input current is 6.6 mA. This, with an input voltage of 134 volts and output is 669V gives a circuit efficiency of 0.745 / (134)(0.066) = 84.3 %. Rather good considering we have a 10 Ohm series resistance. Notice in the model of the transistor (which you can make), all of the capacitance terms are reduced in value to a typical power FET because our mathematical analysis does not take them into account.

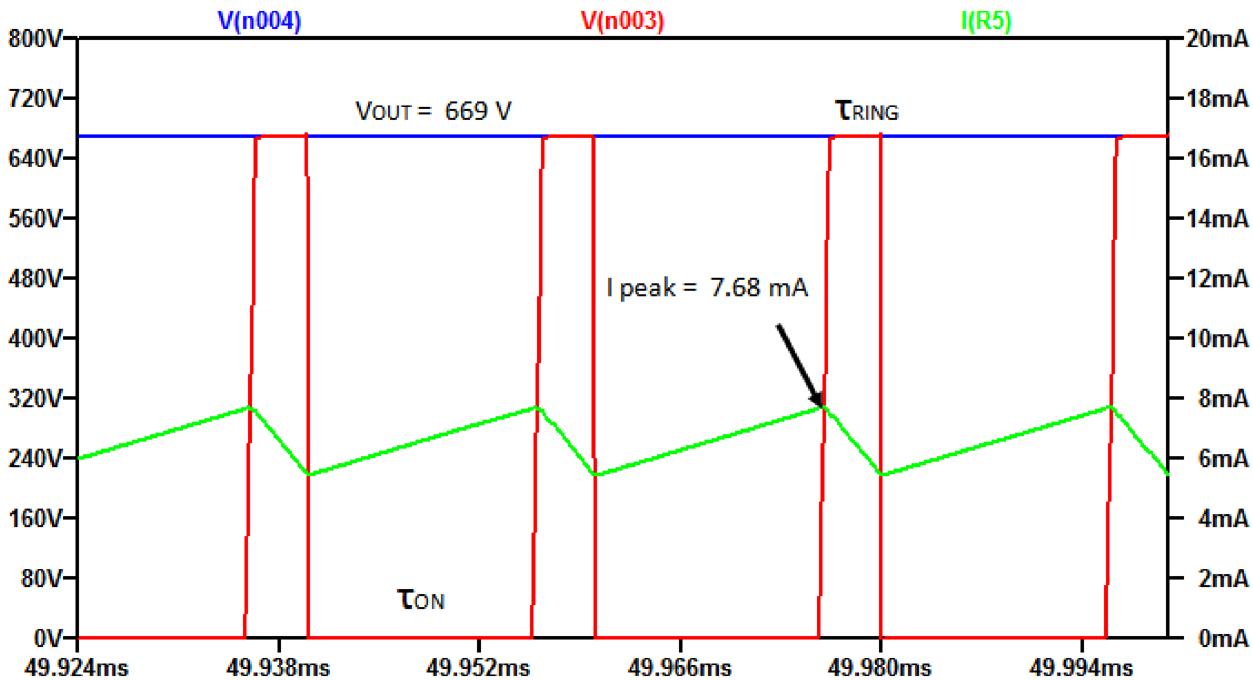


Figure 3.15: 134v to 600v converter. Vout (blue), V_{DRAIN} (red), i_{DRAIN} (green)