

Experiment #4

Step-Up DC-DC Converter

Part 1:

Open-Loop Cascaded Boost DC-DC Converter

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ECEN 5517

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***Please note: Several components (PWM chip, gate drivers, D1) failed at the final stages of data collection on 3/20; several waveforms are missing as a result. The waveforms and data that are presented for the full circuit operation are for a 500 ohm load, instead of 250 ohm load as required to operate near 85 W. Professor Maksimovic ok-ed using this 500 ohm load data:**

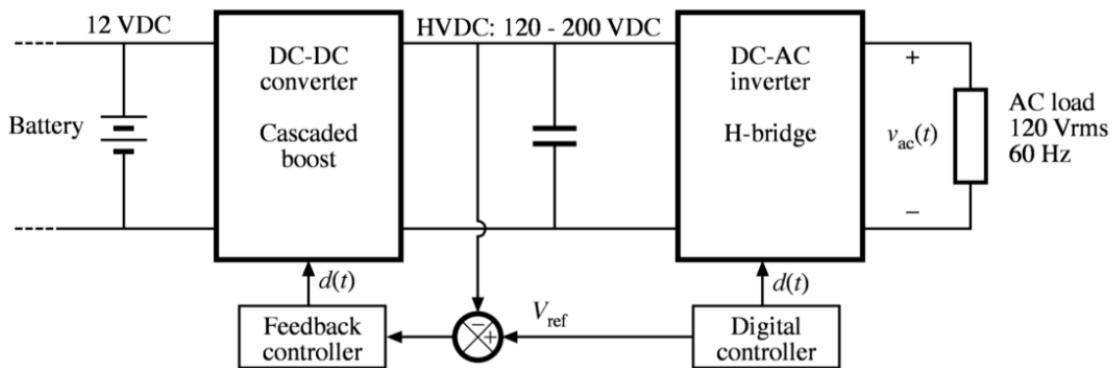


Dragan 5:53 PM

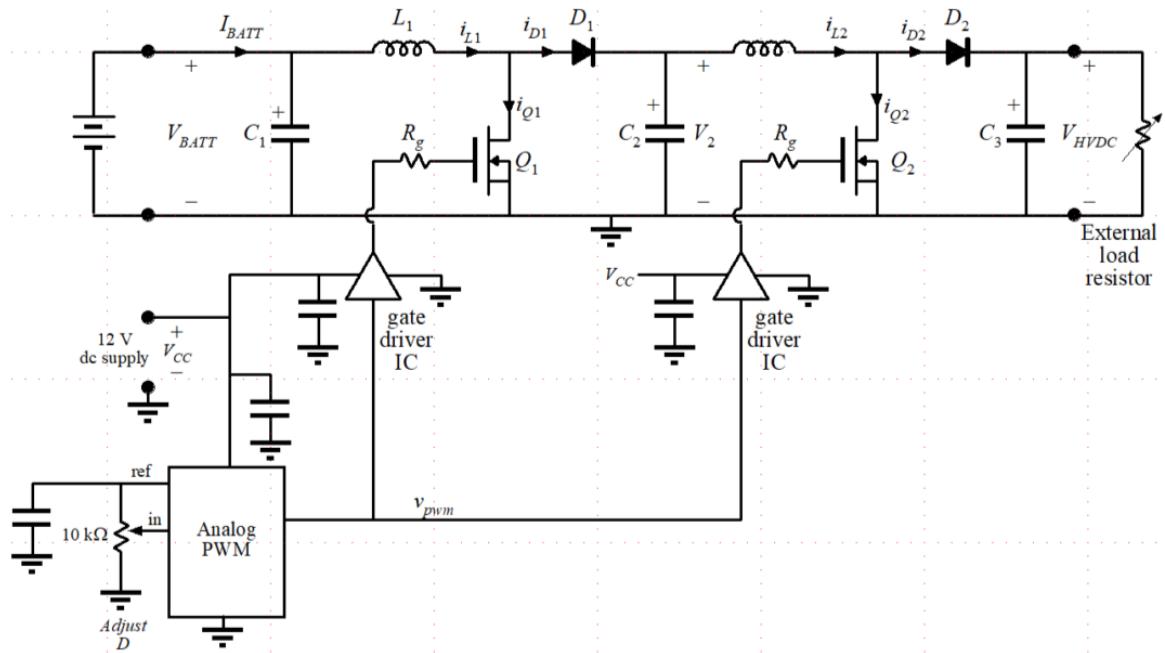
Hi Andrew, turning all results with 500 ohm will be good - please do that.

Introduction:

With the PV-to-battery dc-dc charging stage completed, we now turn to developing the dc-dc step-up converter that will power a single-phase dc-ac inverter and subsequently an AC load. The dc-dc converter should step up the ~12VDC battery voltage to 120-200VDC for input to the inverter. Our approach is to cascade two Boost dc-dc converters controlled by the same PWM signal. We chose this boost converter design for simplicity and to keep components small by having two in series. An analog PWM chip controller will be used to generate the PWM signal, and two separate gate drivers, one for each boost stage, will generate the MOSFET gate signals. Part 2 of this experiment will concern developing the closed-loop analog voltage regulator.



Block diagram of the inverter system to be designed.



Open-loop cascaded Boost converter controlled by an analog PWM

Prelab:

Choose power-stage components, estimate losses, design inductors, and design PWM circuitry.

a) Assuming identical converters, $\frac{V_2}{V_{\text{Batt}}} = \frac{V_{\text{HVDc}}}{V_2} \rightarrow \frac{V_2}{12V} = \frac{150V}{V_2}$

$$V_2 = 42.4V$$

Assuming ideal converters, $M(D) = \frac{1}{D}$

$$\frac{42.4V}{12V} = \frac{1}{1-D} \rightarrow D = .717$$

$$I_{\text{load}} = \frac{P_{\text{load}}}{V_{\text{HVDc}}} = \frac{85W}{150V} = 0.567A$$

$$I_{L2} = \frac{I_{\text{load}}}{D} = \frac{0.567A}{1-.717} = 2.00A$$

$$I_{L1} = \frac{I_{L2}}{D} = \frac{2.0A}{1-.717} = 7.08A$$

b) Choose $f_s = 150 \text{ kHz}$

$$C_3 = \frac{I_{\text{load}} \cdot D \cdot T_s}{2 \Delta V} = \frac{0.567A \cdot 0.717 \cdot \frac{1}{150 \text{ kHz}}}{2 \cdot (0.5 \cdot 150V)} = 271 \text{ nF}$$

$\hookrightarrow 5\%$ ripple

so choose 250PK33 MEFCT810X20 (33 μF, 250V)

$$C_2 = \frac{I_{L2} \cdot D \cdot T_s}{2 \Delta V} = 95.9 \text{ nF}$$

so choose UVR19222MHD (2200μF, 63V)

C₁: UHEJ E331MPD6 (330μF, 25V)

Q₁: PSMN013-100PS (100V, 13.9mΩ)

D₁: FEP16BT-E3/45 (100V, 2x8A)

Q₂: FQP11N40C (400V, 530mΩ)

D₂: FES8GT-E3/45 (400V, 8A)

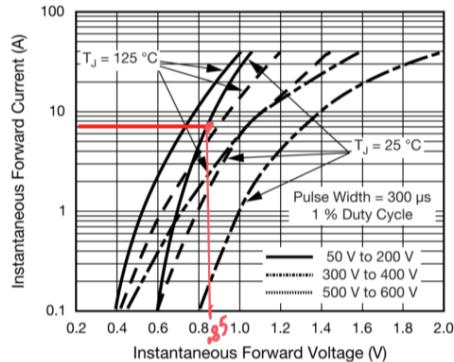
c) Conduction Losses

Q₁: R_{DSONQ1} = 10.8 mΩ at 25°C

$$P_{Loss,Q1} = (D I_{L1})^2 \cdot R_{DSON} = (717 - 7.08A)^2 \cdot 10.8 \text{ mΩ}$$

$$P_{Loss,Q1} = 0.278 \text{ W}$$

D_1 :



$$V_{fwd,D_1} = 0.85 \text{ V}$$

$$\begin{aligned} P_{loss,D_1} &= (D \cdot I_{L_1}) \cdot V_{fwd,D_1} \\ &= (1 - 0.717) \cdot 2.08 \text{ A} \cdot 0.85 \text{ V} \end{aligned}$$

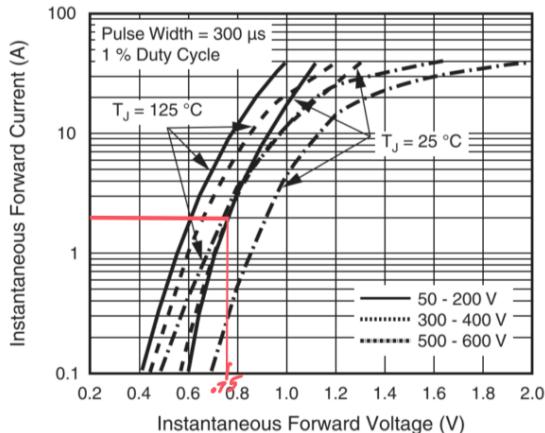
$$P_{loss,D_1} = 1.70 \text{ W}$$

Q_2 : $R_{on,Q_2} = 430 \text{ m}\Omega$ (+ppcaw)

$$P_{loss,Q_2} = (D \cdot I_{L_2})^2 \cdot R_{on,Q_2} = (1 - 0.717) \cdot 2.00 \text{ A}^2 \cdot 430 \text{ m}\Omega$$

$$P_{loss,Q_2} = 0.884 \text{ W}$$

D_2 :



$$V_{fwd,D_2} = 0.75 \text{ V}$$

$$\begin{aligned} P_{loss,D_2} &= (D \cdot I_{L_2}) \cdot V_{fwd,D_2} \\ &= (1 - 0.717) \cdot 2.00 \text{ A} \cdot 0.75 \text{ V} \end{aligned}$$

$$P_{loss,D_2} = 0.425 \text{ W}$$

d) Choose $f_s = 100\text{kHz}$

Limit to 10% ripple in current overall:

$$L_2 = \frac{V_2}{2\Delta i_L} \cdot DT_s = \frac{42.4\text{V}}{2 \cdot (0.05 \cdot 2.00\text{A})} \cdot 717 \cdot \frac{1}{100\text{kHz}}$$

$$L_2 = 1.52\text{ mH}$$

$$L_1 = \frac{V_{Batt}}{2\Delta i_L} \cdot DT_s = \frac{12\text{V}}{2 \cdot (0.05 \cdot 7.08\text{A})} \cdot 717 \cdot \frac{1}{100\text{kHz}}$$

$$L_1 = 122\text{ }\mu\text{H}$$

e)

$$D_{max} = \frac{t_c}{t_c + t_d} \quad T = \frac{1}{f_s} = t_c + t_d$$

$$0.95 = \frac{t_c}{t_c + t_d} \quad \frac{1}{100\text{kHz}} = t_c + t_d$$

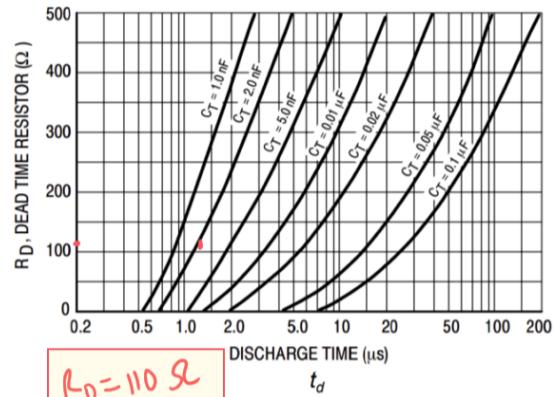
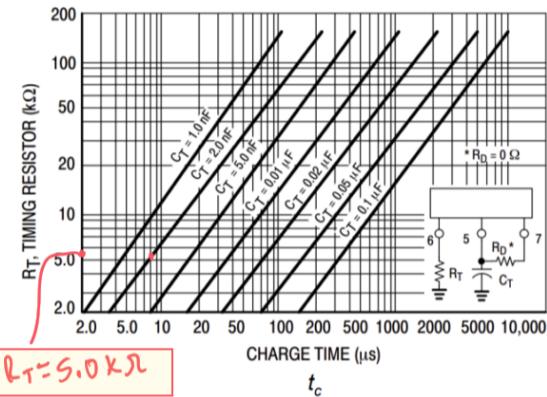
$$t_d = 0.176 t_c \quad \rightarrow \frac{1}{100\text{kHz}} = 1.176 t_c$$

$$t_c = 8.50\text{ }\mu\text{s}$$

$$t_d = 1.50\text{ }\mu\text{s}$$

Choose $C_f = 2.2\text{nF}$

Choose $C_T = 2.2 \mu F$



$$V_{max, sawtooth} = 3.34 V$$

$$V_{min, sawtooth} = 0.975 V$$

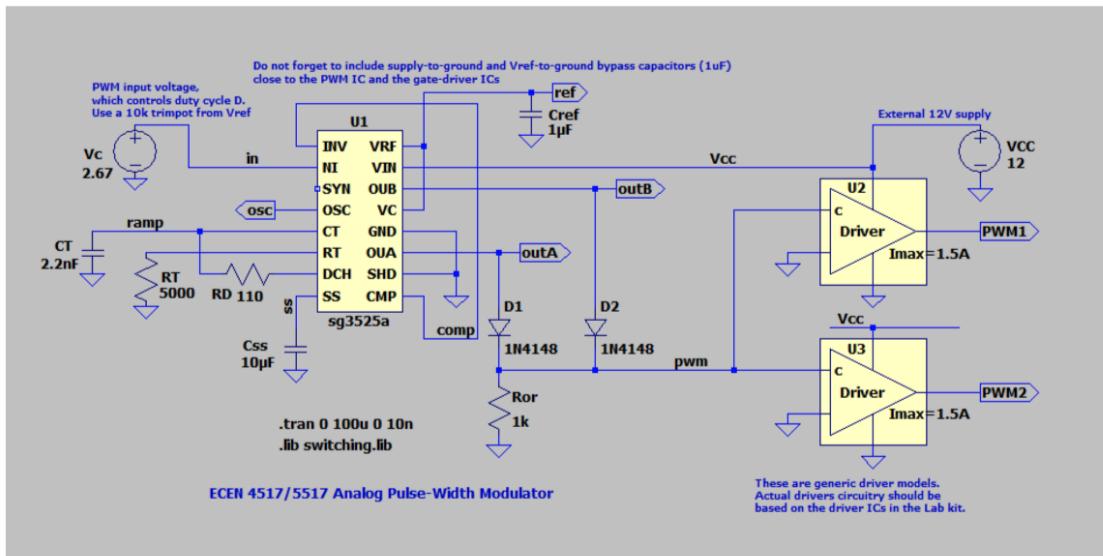
$$D = 0.717$$

$$V_C = (V_{max} - V_{min}) D + V_{min}$$

$$V_C = (3.34 V - 0.975 V) \cdot 0.717 + 0.975 V$$

$$V_C = 2.67 V$$

LTspice circuit



Waveforms



$$t_{c,sm} = 7.41 \mu s \quad 7.39 \mu s$$

$$t_{d,sm} = 0.84 \mu s \quad 0.73 \mu s$$

$$T = \frac{1}{f_s} = 8.25 \mu s \rightarrow f_s = 121 \text{ kHz}$$

$$D_{max} = \frac{t_c}{t_c + t_d} = 0.898$$

Redeney:

$$R_c = 6.03 \text{ k}\Omega \quad (4.7 \text{ k}\Omega + 1 \text{ k}\Omega + 320 \text{ }\Omega)$$

$$R_d = 267 \text{ }\Omega \quad (220 \text{ }\Omega + 47 \text{ }\Omega)$$

Step 2: Analog PWM Circuit

The TI 3525A PWM Controller chip is used to generate the PWM signal.

The PWM circuit diagram is shown below.

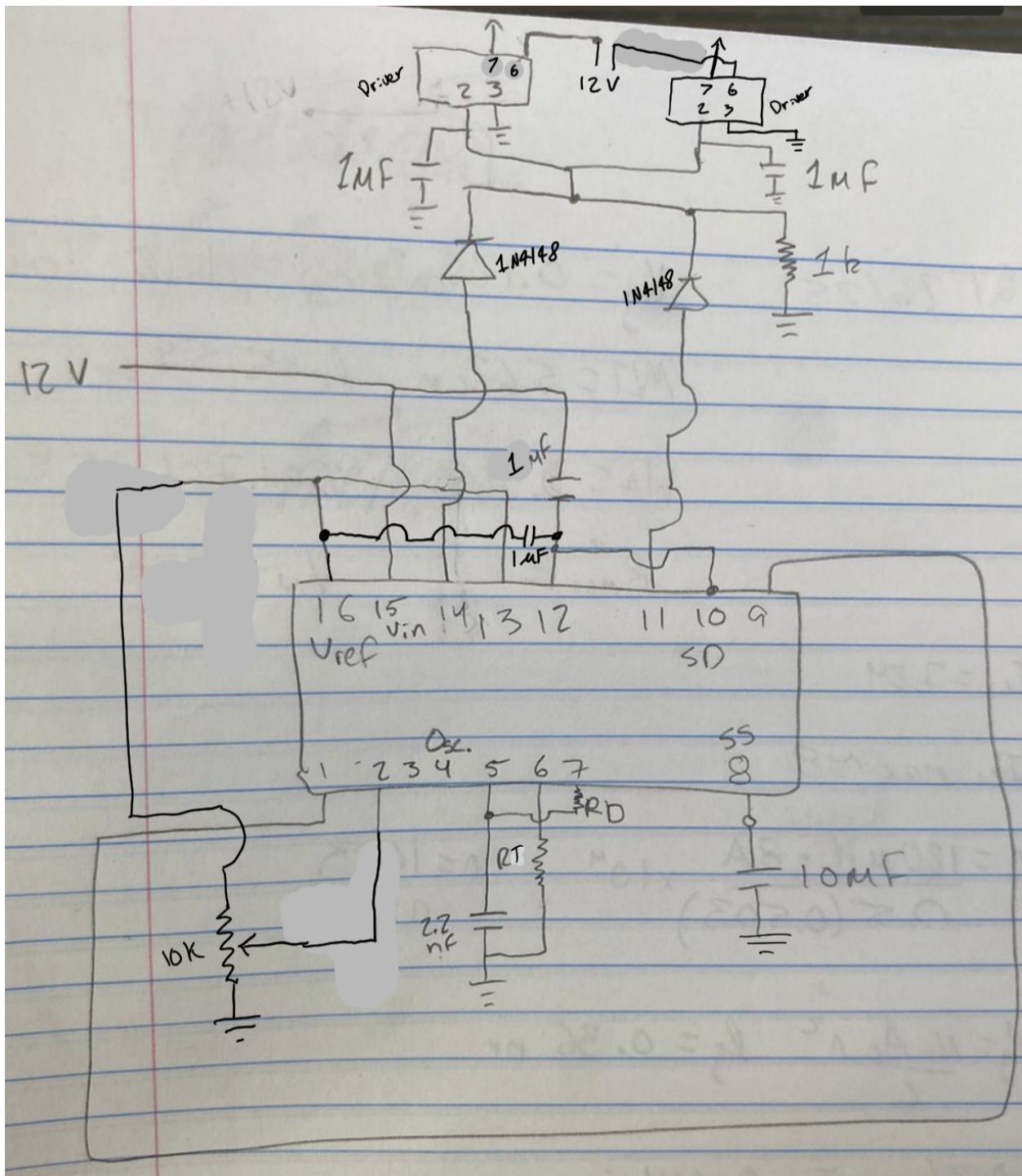


Figure 1: Complete PWM circuit diagram

R_T and R_D were redesigned from our prelab after testing our original values of $R_C=6.03$ kohm and $R_D = 267$ ohm, and finding that D_{MAX} was around 0.70. We increased R_C to 10 kohm, and achieved a D_{MAX} of .809, with a switching frequency of 66 kHz at the gate driver input.

The circuit was constructed and tested. The resulting waveforms and voltages at 60% duty cycle are shown below:

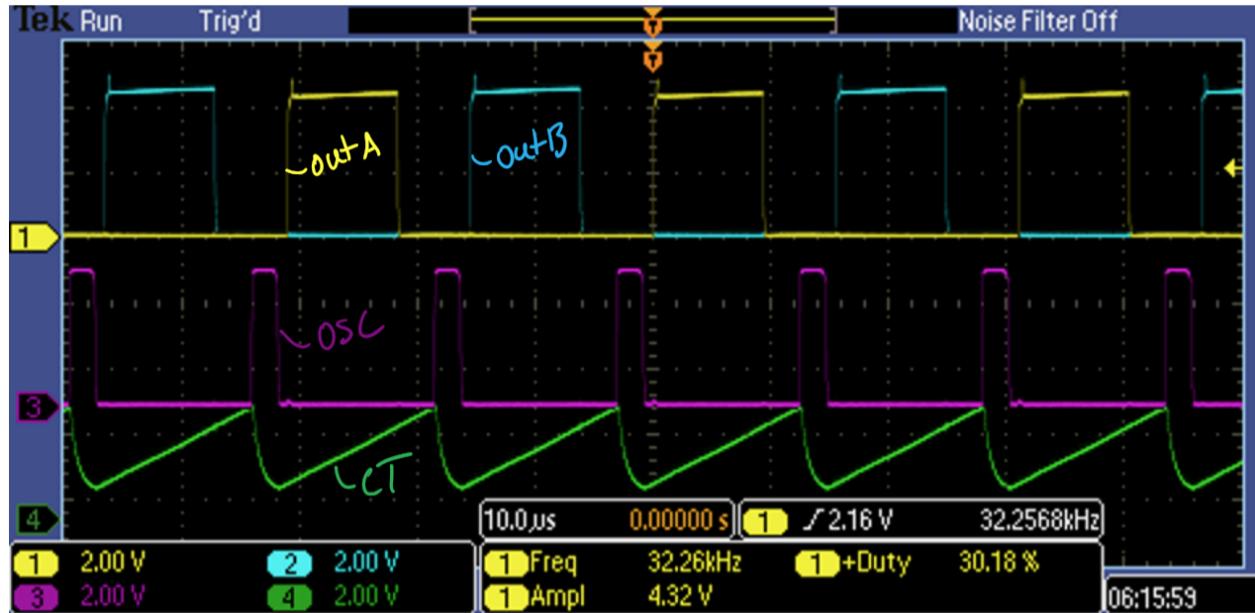


Figure 2: Waveforms for OUTPUT A (yellow), OUTPUT B (blue), OSC (purple), and CT (green) at 60% duty cycle

The waveform for the overall PWM circuit output was not obtained before the circuit failed, but a 60% duty cycle pulsed waveform was observed. The waveform at the MOSFET gates was 12 Vpp, with the same duty cycle.

| Pin | Measured Voltage (V) |
|-----|----------------------|
| 1 | 5.32 |
| 2 | 5.32 |
| 6 | 3.88 |
| 8 | 5.0 |
| 9 | 5.4 |
| 10 | 0 |
| 12 | 0 |
| 13 | 5.32 |
| 15 | 12.6 |
| 16 | 5.8 |

Table 1: DC Voltages on remaining pins of PWM chip

This circuit was simulated in LTSpice as shown below:

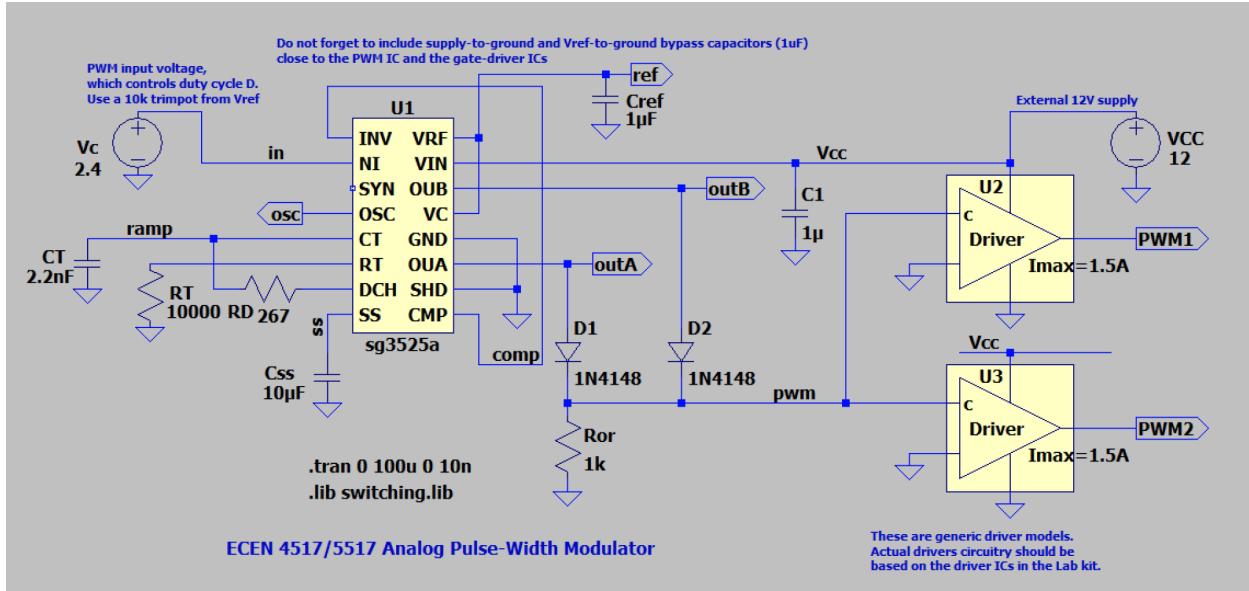


Figure 3: LTSpice simulated circuit

The following waveforms resulted from a transient simulation:

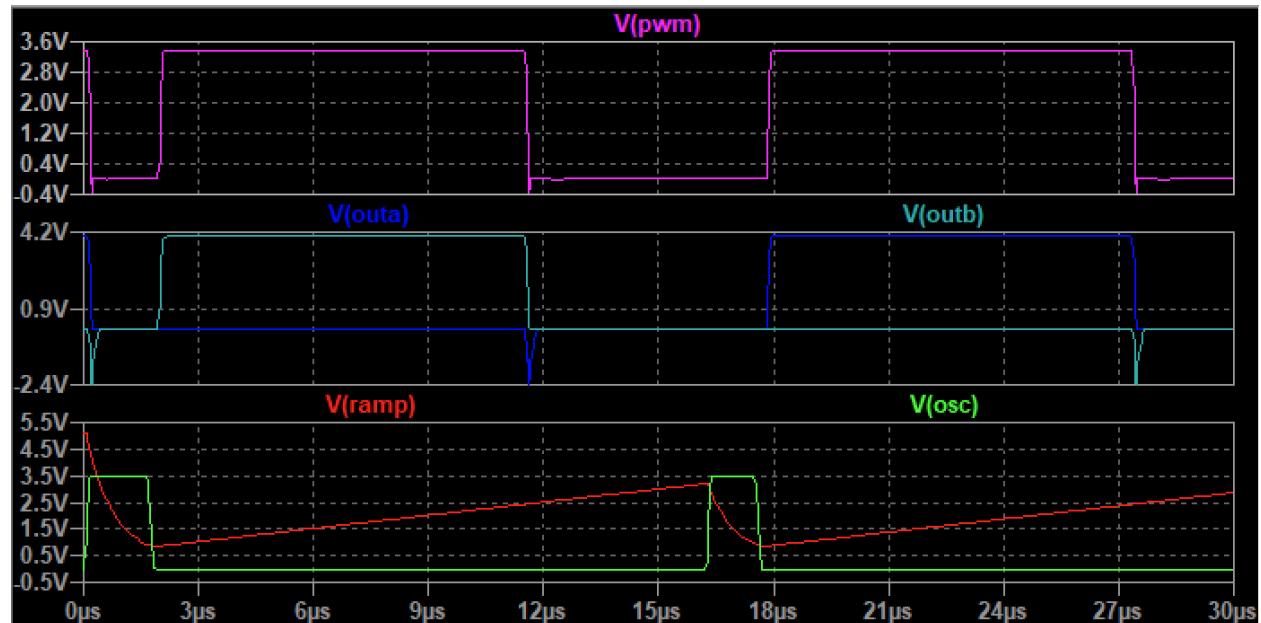


Figure 4: Transient Simulation results of circuit in figure 3

The simulation resulted in a PWM frequency of 63 kHz, with a D_{MAX} of 90.1%. The measured results differ from the simulated results due to the fact that the actual PWM output does not track exactly with the charging time - there is a delay between the start of the charging time and the positive pulse of the PWM. This results in a lower actual D_{MAX} than in simulation.

Step 3: Inductor Design and Construction

The two inductors were designed for 100kHz switching frequency, a B_{MAX} of 0.25 T, and 1W maximum resistive power loss.

Inductor L_1 was designed for an inductance of 120 uH.

Inductor L_2 was designed based on the fact that the smaller core has a greater height, which would decrease the number of layers (and thus proximity loss). A design with the larger core had resulted in a very large number of turns. Inductor L_2 was designed for an inductance of 760 uH.

| Inductor | Core | Number of turns | Air gap | Wire gauge | Measured inductance at 100 kHz | Measured series resistance |
|----------|----------|-----------------|---------|------------|--------------------------------|----------------------------|
| L_1 | PQ 26/25 | 17 turns | 14 mil | AWG 16 | 130 uH | 11.7 mohm |
| L_2 | PQ 26/25 | 37 turns | 10 mil | AWG 20 | 848 uH | 70.2 mohm |

Table 2: Inductor design parameters

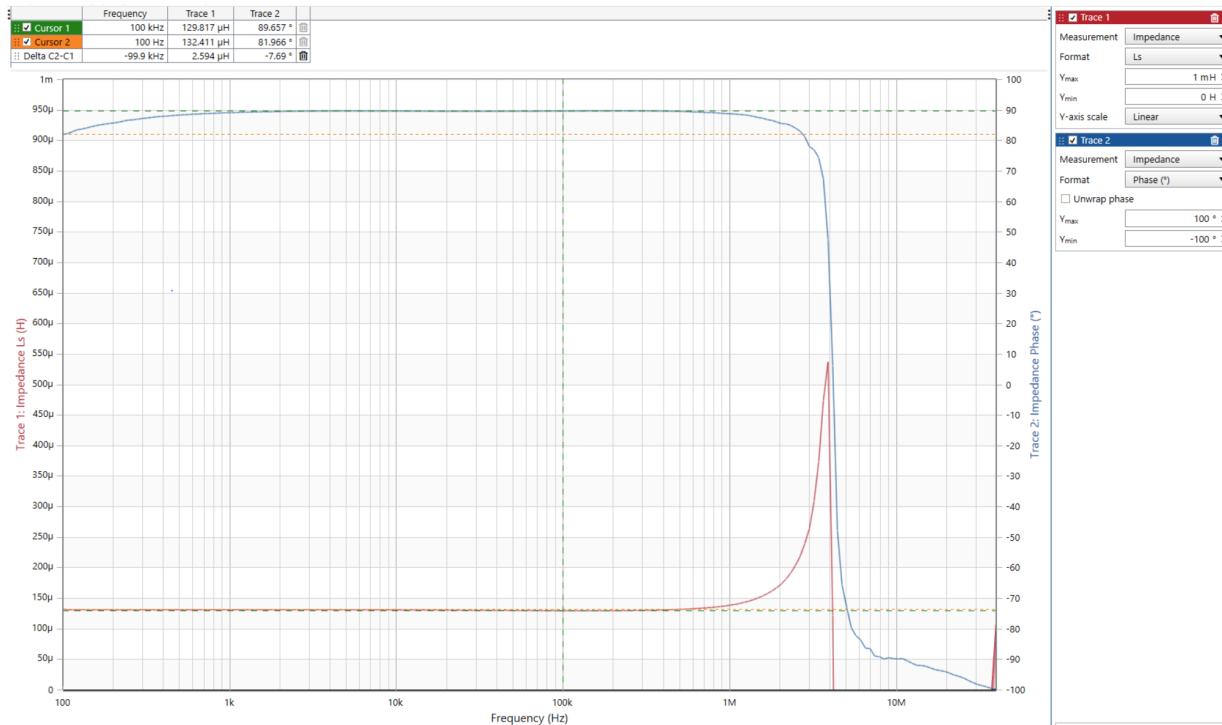


Figure 5: Inductor L_1 bode analyzer plot

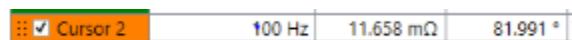


Figure 18: Bode analyzer resistance of Inductor L_1

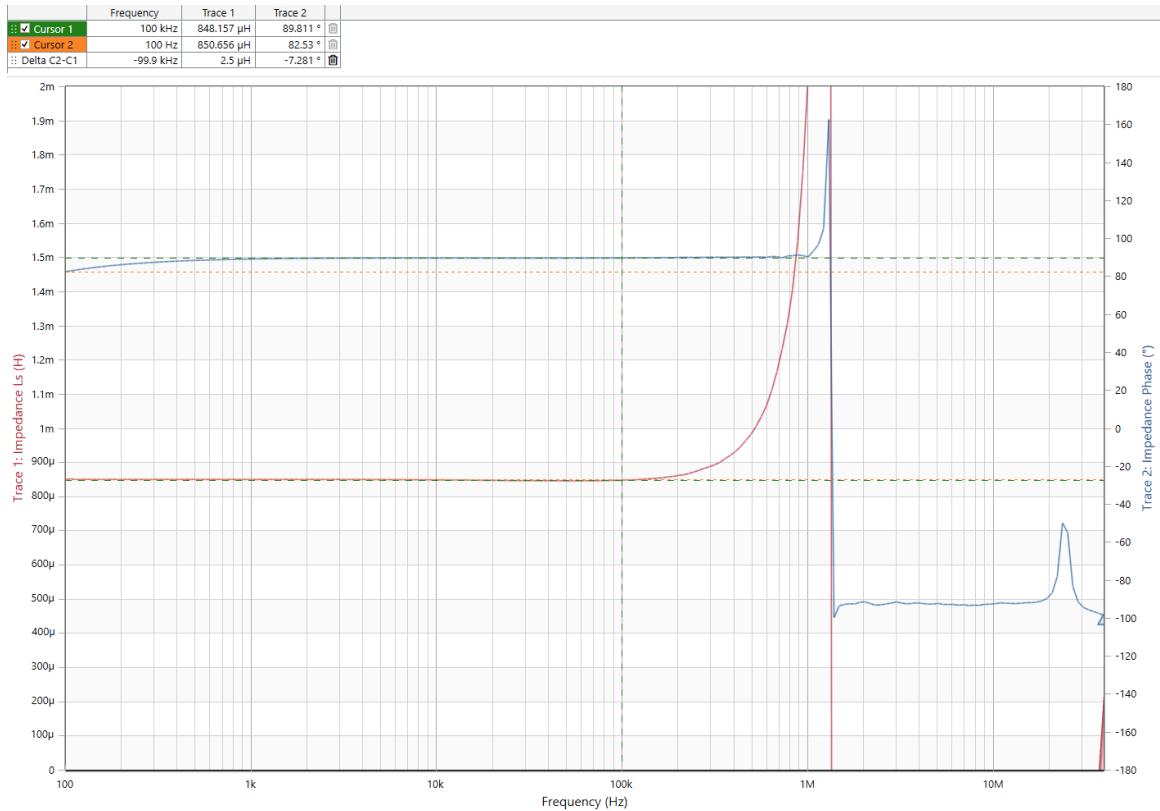


Figure 6: Inductor L_2 bode analyzer plot

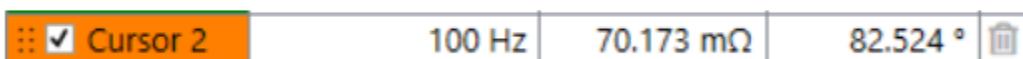


Figure 17: Bode analyzer reading of inductor resistance for L_2

We designed our inductors before completing the PWM circuitry, thus the reason for designing the inductors for 100 kHz switching frequency when we ended up operating at 66 kHz. We decided to see if our inductors would maintain an unsaturated state at nominal operating conditions. When testing with 12.55 V input and 500 ohm load, the inductors did not saturate. However, when switching to 250 ohm load to operate around 85 W input power, our inductors saturated and efficiency dropped considerably. We will need to redesign our inductors for the next part of the lab, as we reconstruct the rest of our damaged circuit.

Step 4: Cascaded Boost Power Stage Construction

The cascaded boost power stage was constructed and connected to the existing PWM circuit.

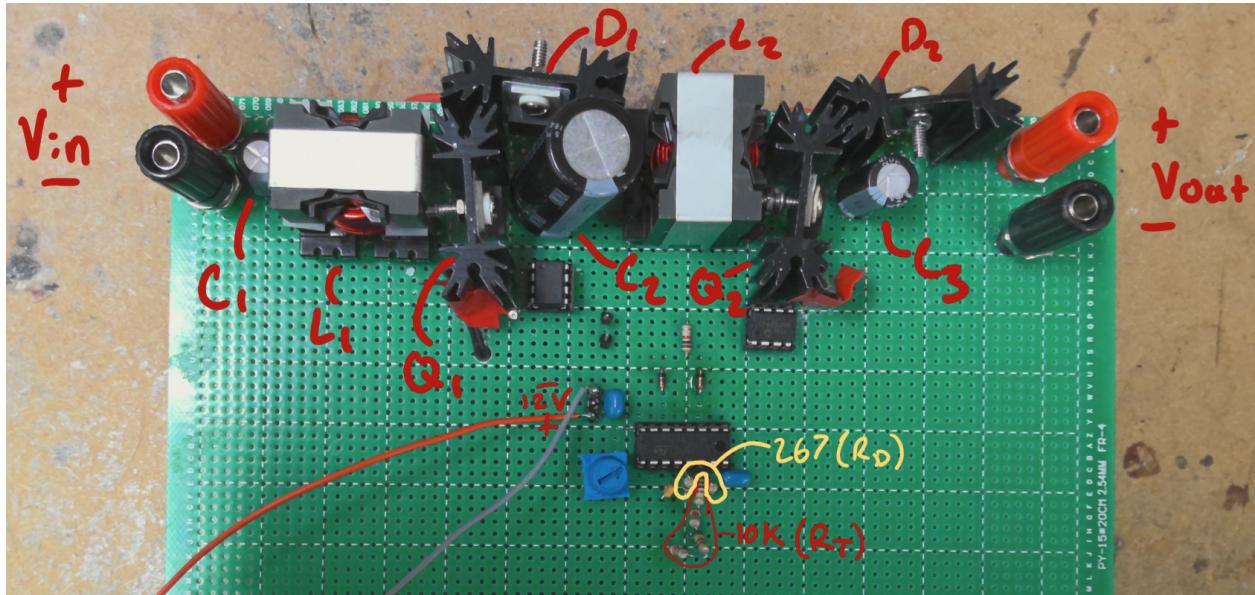


Figure 7: Top of fully constructed cascaded boost converter

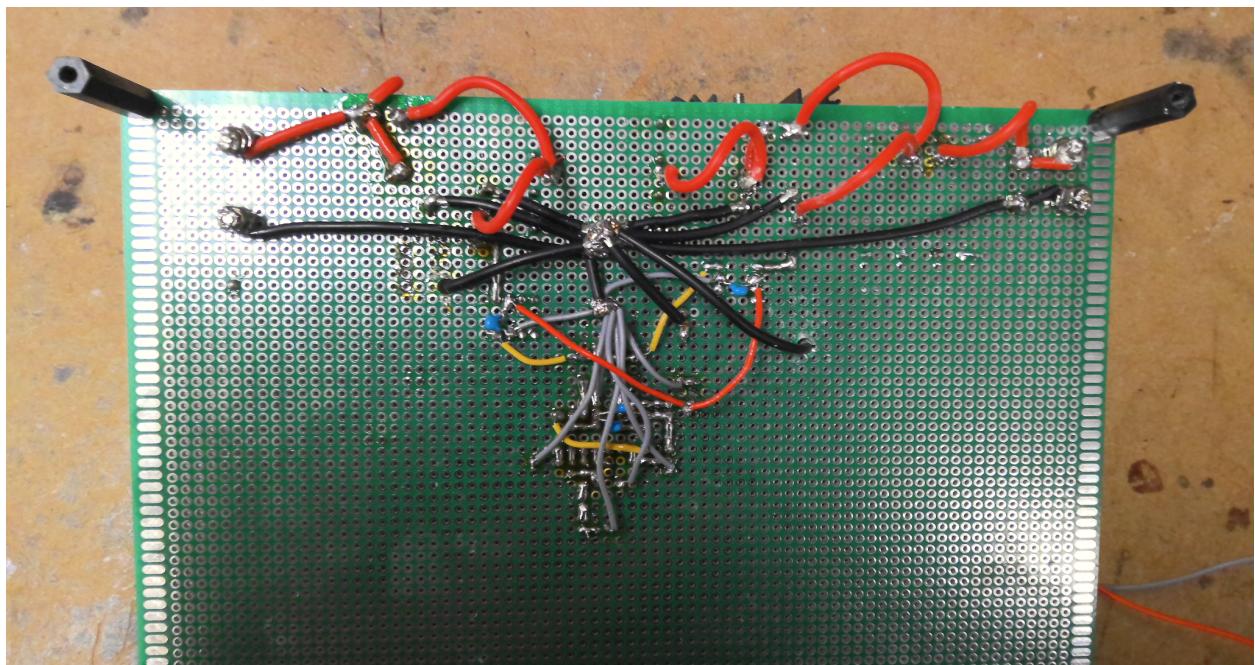


Figure 8: Bottom of fully constructed cascaded boost converter

Step 5: Cascaded Boost Converter Testing

The input and output currents and voltages of the converter were measured with a 500 Ohm load at 71.7% Duty Cycle, shown below:

Input Voltage: 12.55 V

Input Current: 4.10 A

Output Voltage: 153.3 V

Output Current: 0.295 A

$$\text{Input Power} = \text{Input Voltage} * \text{Input Current}$$

$$\text{Input Power} = 51.46 \text{ W}$$

$$\text{Output Power} = \text{Output Voltage} * \text{Output Current}$$

$$\text{Output Power} = 45.22 \text{ W}$$

$$\text{Efficiency} = \text{Output Power} / \text{Input Power}$$

$$\text{Efficiency} = 87.9\%$$

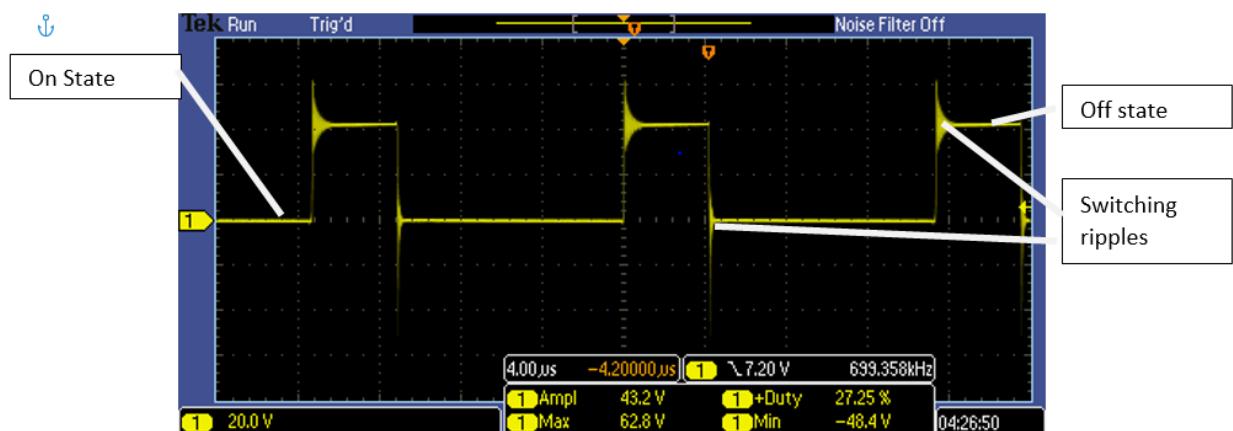


Figure 9: Drain to Source Voltage waveform of transistor Q1

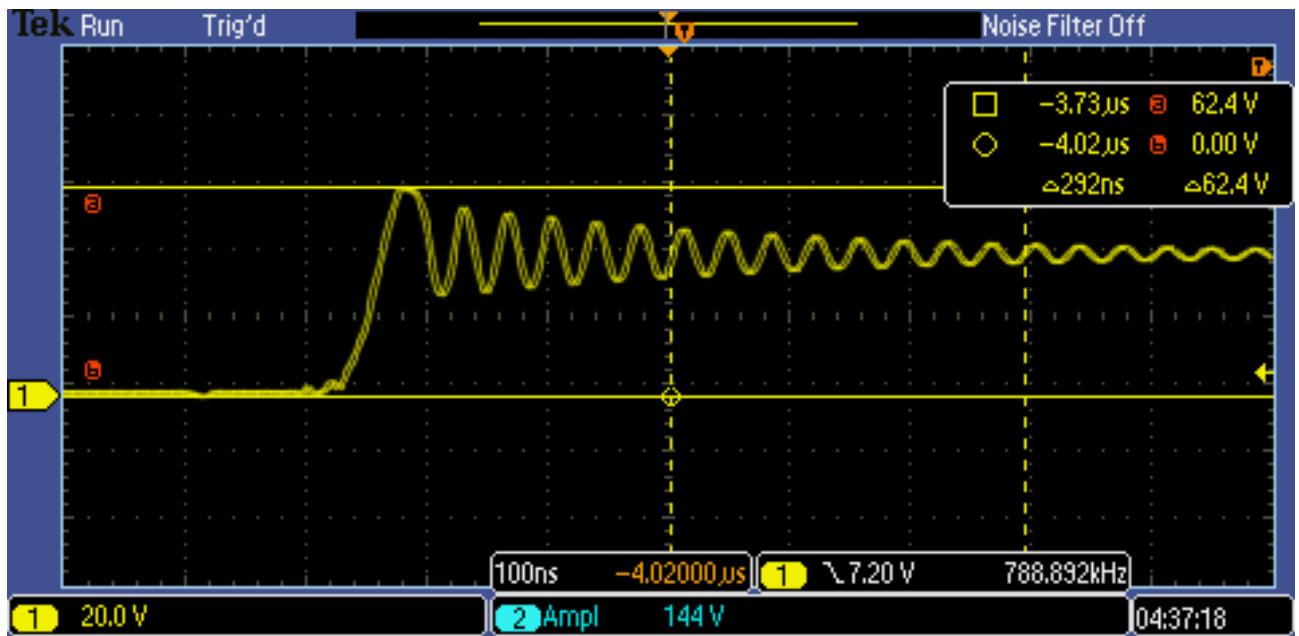


Figure 10: Magnified view of the Drain to Source Voltage waveform of transistor Q1

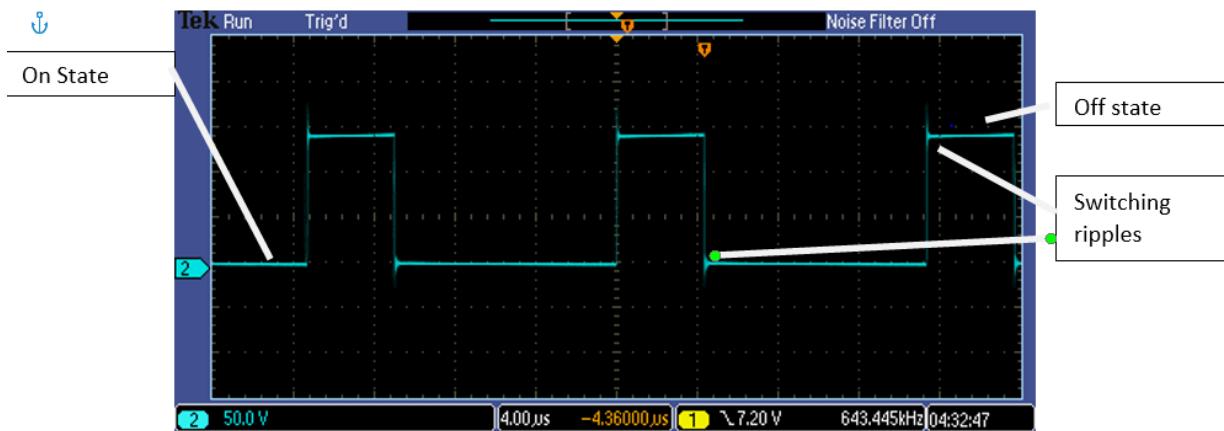


Figure 11: Drain to Source Voltage waveform of transistor Q2

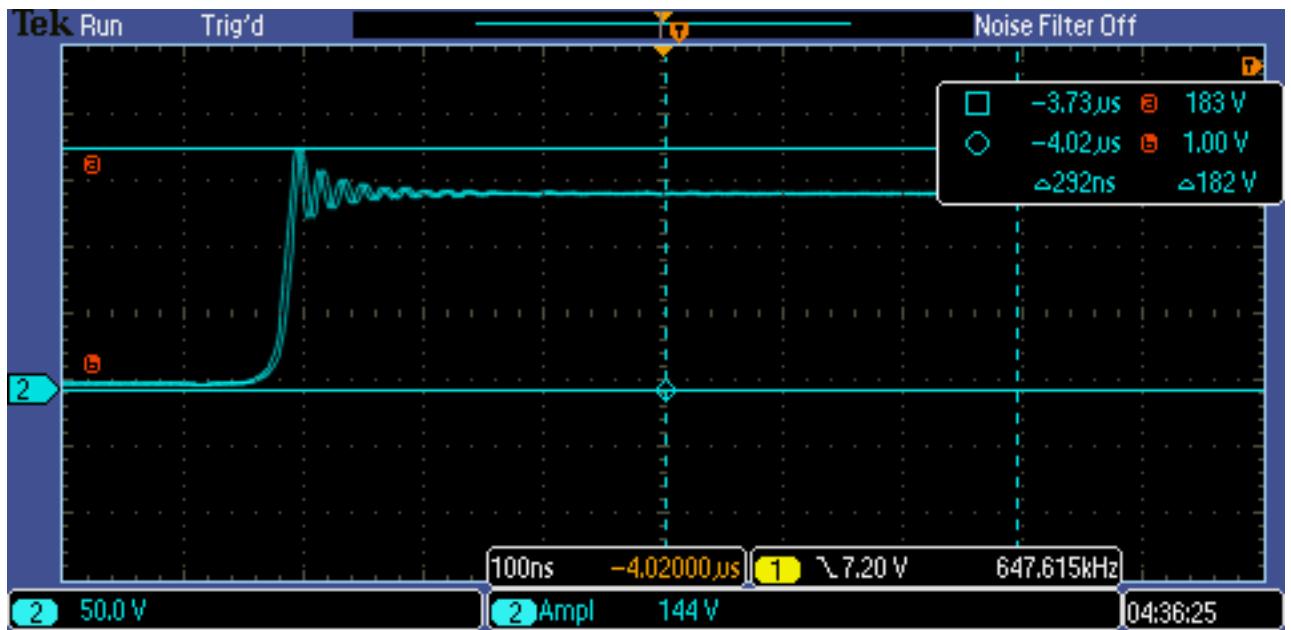


Figure 12: Magnified view of the Drain to Source Voltage waveform of transistor Q2



Figure 13: Inductor Current waveforms, with the Inductor L1 current waveform shown in purple and the Inductor L2 waveform shown in green

The difference between the minimums and maximums of the Inductor current waveforms were measured with a 250 Ohm load, and the inductor current ripple can be calculated by dividing this value in half.

Inductor L1 measured difference between minimum and maximum: 1.12 A

Inductor L1 Current ripple = 1.12 A / 2

Inductor L1 Current ripple = 0.56 A

Inductor L2 measured difference between minimum and maximum: 568 mA

Inductor L2 Current ripple = 568 mA / 2

Inductor L2 Current ripple = 284 mA

Step 6: Loss Budget

Diode and Transistor Conduction Losses:

The specifications for our Diode Forward Voltage Drop(Vf) and the Transistor on-resistance(Ron) are listed below.

Q2 Ron = 0.53 Ohms

Q1 Ron = 25 mOhms

D1 Vf = 1.3 V

D2 Vf = 1.3 V

The conduction losses for the Transistors can be calculated as

$$P_{\text{loss}} = (I_{\text{rms}})^2 * \text{Ron}$$

And the power losses for the Diodes can be calculated as

$$P_{\text{loss}} = I_{\text{rms}} * Vf$$

Unfortunately, at 6 PM on the due date of this lab report, our converter experienced a serious failure. This left us unable to take measurements of the diode and transistor currents directly. However, we were able to take measurements of the inductor currents. The Q1 and Q2 transistor currents are the currents of Inductors L1 and L2 respectively when the transistors are conducting in the interval from 0 seconds to DTs. Similarly, the D1 and D2 diode currents are equal to the currents of Inductors L1 and L2 respectively when the MOSFETs are not conducting, i.e. from the interval between (1-D)Ts and Ts, where Ts is the time of one switching cycle.

The equation for the I_{rms} values of the currents can be calculated as:

$$I_{rms} = I\sqrt{D}\sqrt{1 + (1/3)*(\Delta i/I)^2}$$

Where I is the DC component of the current and Δi is the current ripple. We can observe from Figure X that the DC component of the inductor L_1 is roughly 4 Amps, and the DC component of inductor L_2 is 1 Amp. Using the values for the inductor current ripples calculated in step 3, and with a D value of 0.717 for the transistor D values and 0.283 for the diode D values, the RMS currents can be calculated and the results are shown below.

$$I_{rmsQ1} = 3.40 \text{ A}$$

$$I_{rmsQ2} = 0.86 \text{ A}$$

$$I_{rmsD1} = 2.13 \text{ A}$$

$$I_{rmsD2} = 0.54 \text{ A}$$

The Conduction losses for each component can then be calculated as follows:

$$P_{lossQ1} = (4.24 \text{ A})^2 * 0.53 \text{ Ohms} = 6.13 \text{ W}$$

$$P_{lossQ2} = (0.86 \text{ A})^2 * 0.025 \text{ Ohms} = 0.018 \text{ W}$$

$$P_{lossD1} = (2.67 \text{ A}) * 1.3 \text{ V} = 2.77 \text{ W}$$

$$P_{lossD2} = (0.54 \text{ A}) * 1.3 \text{ V} = 0.70 \text{ W}$$

Giving a total conduction loss from the transistors and diodes of

$$P_{loss,Conduction} = 9.62 \text{ W}$$

Switching Losses:

When the diode in each boost converter stops conducting current, there is a small reverse current. This causes a current spike through each transistor in the corresponding boost converter

before the transistor voltage drops. During this period, the transistor conducts current while maintaining a voltage across it, causing switching losses.

The waveforms of the voltage and current of Transistor 2 were measured in order to estimate this switching loss, shown below.

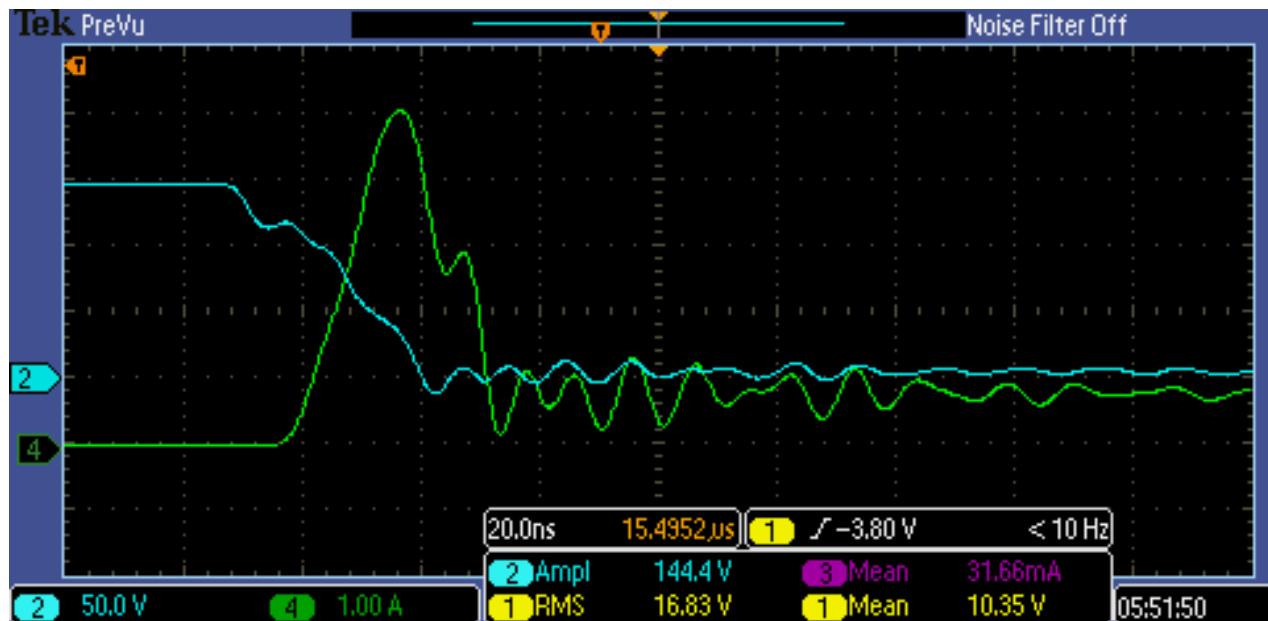


Figure 14: Oscilloscope measurement of the Transistor Q2 current and voltage during the switching period when the transistor begins conducting

Using the oscilloscope cursors,, estimates of linear equations defining the current and voltage behavior during the switching period were created, using the moment the diode current becomes non-zero as $t = 0$, shown below.

$$\text{For } 0 < t < 23.2 \text{ ns}, V = (120 \text{ V}) - (5.17 * 10^9 \text{ V/s}) * t$$

$$\text{For } 0 < t < 19.6 \text{ ns}, I = (2.62 * 10^8 \text{ A/s}) * t$$

$$\text{For } 19.6 \text{ ns} < t < 23.2 \text{ ns}, I = 5.14 \text{ A} - (2.11 * 10^8 \text{ A/s}) * t$$

By integrating $V*I$ over the period from $0 < t < 23.2 \text{ ns}$, we can estimate the energy lost in the transistor Q2 as $2.67 * 10^{-6}$ Joules every switching cycle. By multiplying this by the switching frequency of the converter, we can estimate the switching power loss as

$$P_{\text{onSw2}} = 2.67 * 10^{-6} \text{ Joules} * 66 \text{ kHz}$$

$$P_{\text{onSw2}} = 0.176 \text{ W}$$

Our Converter failed before we were able to take measurements of the switching waveforms of Q1, but we can estimate the Power loss by taking the on-state current and off-state voltage of the transistor, multiplying them by the turn-on delay time listed in the transistor datasheet, and multiplying that by the switching frequency of the converter. The on-state current will be the minimum current of Inductor L1, 3.44 Amps. The turn-on delay time is 20.7 ns. The off state voltage of the transistor is 43.2 Volts, as shown in Figure 9.

$$P_{\text{onSw1}} = I_{\text{on}} V_{\text{off}} t_{\text{delay}} f_s = (3.44 \text{ A}) * (43.2 \text{ V}) * (20.7 \text{ ns}) * (66 \text{ kHz})$$

$$P_{\text{onSw1}} = 0.20 \text{ W}$$

A similar estimation can be made for the switching losses turning off transistor Q1, using the maximum current of Inductor L1, 4.56 Amps, and the turn-off delay time in the data sheet of 52.5 ns.

$$P_{\text{offSw1}} = I_{\text{on}} V_{\text{off}} t_{\text{delayoff}} f_s = (4.56 \text{ A}) * (43.2 \text{ V}) * (52.5 \text{ ns}) * (66 \text{ kHz})$$

$$P_{\text{offSw1}} = 0.68 \text{ W}$$

And the switching losses when turning off transistor Q2, using the maximum current of Inductor L2, 1.284 Amps, the off-state Voltage of 144 V shown in Figure 11, and the turn-off delay time of 81 ns listed in the datasheet, can be estimated as

$$P_{\text{offSw2}} = I_{\text{on}} V_{\text{off}} t_{\text{delayoff}} f_s = (1.284 \text{ A}) * (144 \text{ V}) * (81 \text{ ns}) * (66 \text{ kHz})$$

$$P_{\text{offSw2}} = 0.99 \text{ W}$$

This gives a total switching loss from both transistors of

$$P_{\text{lossSw}} = 2.046 \text{ W}$$

Inductor Loss Calculations:

The copper loss can be estimated using the measured series resistance and the RMS current in the inductors.

$$P_{\text{Loss, DCCU, L1}} = R_{S, L1} * I_{\text{RMS, L1}}^2 = 11.7 \text{ mohm} * (4.01 \text{ A})^2 = 0.188 \text{ W}$$

$$P_{\text{Loss, DCCU, L2}} = R_{S, L2} * I_{\text{RMS, L2}}^2 = 70.2 \text{ mohm} * (1.03 \text{ A})^2 = 0.072 \text{ W}$$

According to the datasheet, the core loss of the PQ 26/25-N95 cores is less than 3.60 W at 100 kHz, 25 °C.

To calculate the increased copper loss factor due to the proximity effect we can use the equation

$$F_R = \frac{P}{P_{dc}} = \frac{1}{3} \left(\frac{h}{\delta} \right) (2M^2 + 1)$$

Where h is conductor thickness, δ is penetration depth of the conductor, and M is the number of layers in the inductor. We can more accurately estimate the h/δ factor by considering a layer of round conductors as an effective foil conductor as

$$\varphi = \frac{h}{\delta'} = \sqrt{\eta} \sqrt{\frac{\pi}{4}} \frac{d}{\delta}$$

$$\text{Where } \eta = \sqrt{\frac{\pi}{4}} d \frac{n_\ell}{l_w}$$

n_l is the number of turns of round wire in a layer, and l_w is the layer width.

Additionally,

$$\delta = \sqrt{\frac{\rho}{\pi \mu f}}$$

For L_1 (AWG 16, $d = 0.129$ cm, 2 layers, $n_l = 8.5$ turns, $l_w = 11.6$ mm)

$$\delta = \sqrt{1.724 * 10^{-6} \text{ ohm} * \text{cm} / (\pi * 4 * \pi * 10^{-7} \text{ H/m} * 66 \text{ kHz})} = 0.0257 \text{ cm}$$

$$\begin{aligned} h/\delta' &= \sqrt{\sqrt{\pi/4} * 1.29 \text{ mm} * 8.5 \text{ turns} / 11.6 \text{ mm}} * \sqrt{\pi/4} * 0.129 \text{ cm} / 0.0257 \text{ cm} \\ &= 4.07 \end{aligned}$$

$$F_R = \frac{1}{3} * 4.07 * (2*2^2 + 1) = 12.2$$

$$P_{\text{loss}} = P_{\text{loss,CU,L1}} * F_R$$

$$P_{\text{loss}} = 0.188 \text{ W} * 12.2$$

$$P_{\text{loss,CU_TOTAL,L1}} = 2.29 \text{ W}$$

For L₂ (AWG 20, d = 0.0813 cm, 4 layers, n_l = 9.25, l_w = 8.08 mm)

$$\delta = \sqrt{1.724 * 10^{-6} \text{ ohm} * \text{cm} / (\pi * 4 * \pi * 10^{-7} \text{ H/m} * 66 \text{ kHz})} = 0.0257 \text{ cm}$$

$$h/\delta' = \sqrt{\sqrt{\pi/4} * 0.813 \text{ mm} * 9.25 \text{ turns} / 8.08 \text{ mm}} * \sqrt{\pi/4} * 0.0813 \text{ cm} / 0.0257 \text{ cm}$$

$$= 2.54$$

$$F_R = \frac{1}{3} * 4.07 * (2 * 2^2 + 1) = 27.94$$

$$P_{\text{loss}} = P_{\text{lossDC}} * F_R$$

$$P_{\text{loss}} = 0.072 \text{ W} * 27.94$$

$$P_{\text{Loss,CU_TOTAL,L2}} = 2.01 \text{ W}$$

Total Losses:

$$P_{\text{Loss,TOTAL}} = P_{\text{loss,CU_TOTAL,L1}} + P_{\text{Loss,CU_TOTAL,L2}} + P_{\text{Loss,Core,L1}} + P_{\text{Loss,Core,L2}} + P_{\text{LossSw}} + P_{\text{loss,Conduction}}$$

$$P_{\text{Loss,TOTAL}} = 2.29 \text{ W} + 2.01 \text{ W} + 2 * 3.6 \text{ W} + 2.046 \text{ W} + 9.62 \text{ W} = 23.17 \text{ W}$$

The calculated losses are much greater than the 6.24 W of losses observed when operating the boost converter.

Possible Improvements:

A large amount of our losses come from the conduction losses of the transistors and diodes. As these are caused by the circuit currents, they are unavoidable. The other large area of power losses were the inductor copper losses, which were greatly exaggerated by the proximity effect. Redesigning our inductors to reduce the number of layers would be the easiest way to reduce this effect.