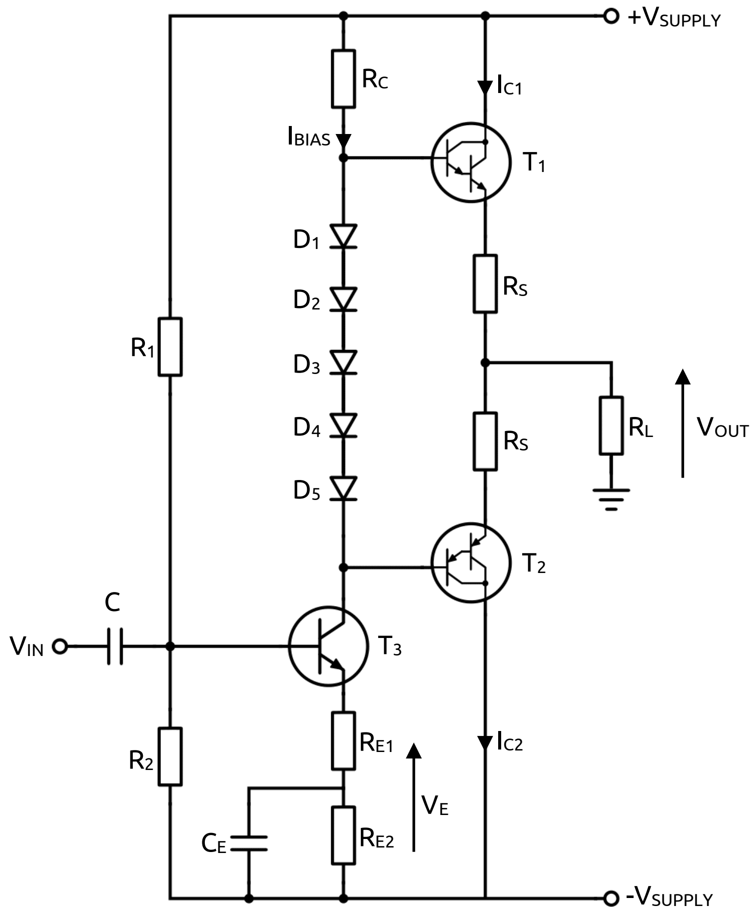
Benjamin Griffiths (160159871)

EEE223 | ASSIGNMENT 1



Specification	
Power (RMS)	100W
h_{FE} (T_1 & T_2)	700
h_{FE} (T_3)	150
V_d	0.7V
R_L	4Ω
I_{BIAS}	5mA
V_E	10V
Voltage Gain	25

To determine the symmetrical rail voltages, the maximum voltage across the load must be calculated to achieve the required power dissipation of 100W.

$$P = \frac{V^2}{R}$$

$$V_{RMS} = \sqrt{100 \times 4} = 20V$$

$$V_{PEAK} = \sqrt{2} \times 20 = 28.28V$$

Therefore, the supply rails must be at least +/- 28.28V, but I will choose **+/- 30V** since this is a nice round value and will give some headroom to overcome voltage drops across the transistors and R_S resistors etc.

Calculating the base voltage and current for T_3:

$$h_{FE} = \frac{I_C}{I_B}$$

$$I_B = \frac{0.005}{150} = 33.3\mu A$$

To ensure the base will always have sufficient current, I will set the current flowing through R1 and R2 to be 1mA.

$$R_{TOTAL} = \frac{30 - (-30)}{0.001} = 60k\Omega$$

From the specification, V_E is 10V, hence the voltage at the base of T_3 will be:

$$V_B = -30 + 10 + 0.7 = -19.3V$$

Hence:

$$1mA \times R_2 = 10.7V$$

$$1mA \times R_1 = 49.3V$$

$$R_1 + R_2 = 60k\Omega$$

Solving:

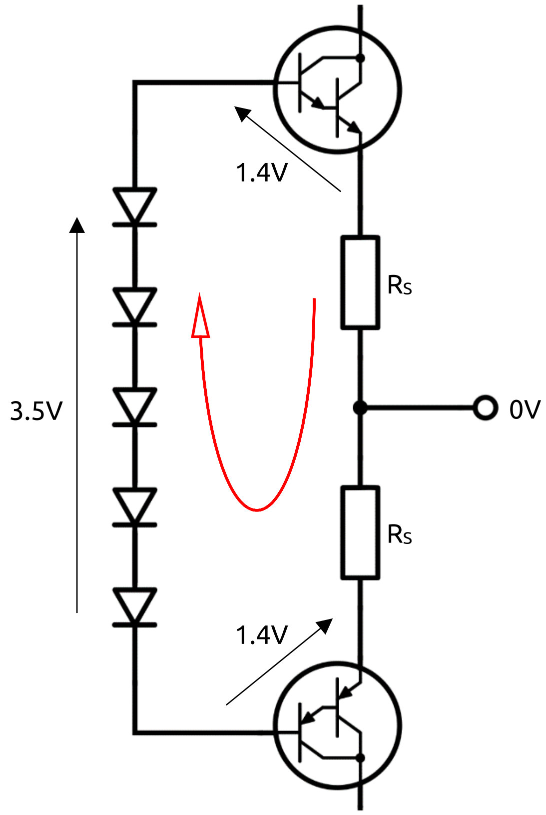
$$R_1 + \left(\frac{10.7}{1mA}\right) = 60k\Omega$$

$$\left(\frac{49.3}{1mA}\right) + R_2 = 60k\Omega$$

$$\boldsymbol{R_1 = 49.3k\Omega}$$

$$\boldsymbol{R_2 = 10.7k\Omega}$$

Summing the voltages around the power amplifier section:



$$2(1.4) + 2V_{RS} = 3.5$$

$$V_{RS} = 0.35V$$

Calculating the largest current flowing through R_s and R_L:

$P_{PEAK} = 2 \times P_{AVERAGE}$ (Assuming Sinusoidal Signal)

$$I_{PEAK} = \frac{100 \times 2}{28.28} = 7.07A$$

Solving:

$$\boldsymbol{R_S = \frac{0.35}{7.07} = 50m\Omega}$$

Since the output voltage is centred at 0V, we can work out the voltage across R_C:

$$V_B\,(T_1) = 0 + 0.35 + 1.4 = 1.75V$$

$$V_{RC} = 30 - 1.75 = 28.25V$$

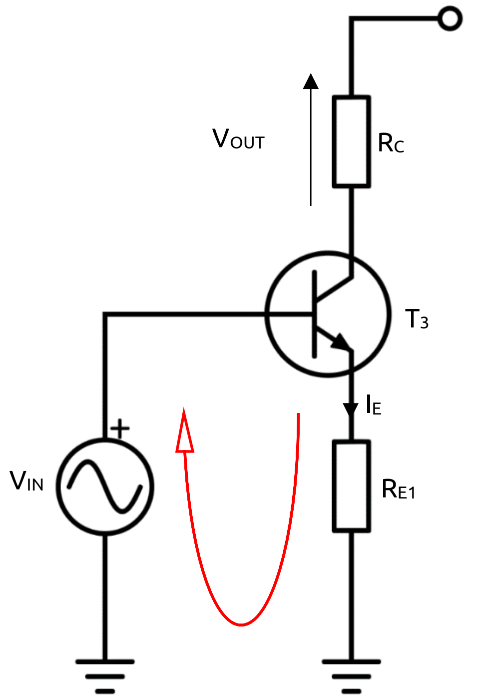
$$\boldsymbol{R_C = \frac{28.25}{0.005} = 5.65k\Omega}$$

We know the current flowing through R_{E1} and R_{E2} is 5mA, and that there is a voltage of 10V across them.

$$10 = 0.005(R_{E1} + R_{E2})$$

$$(R_{E1} + R_{E2}) = 2k\Omega$$

Summing small signal voltages around T_3:



$$V_{IN} = I_E \times R_{E1}$$

$$V_{OUT} = I_E \times R_C$$

$$\frac{V_{OUT}}{V_{IN}} = \frac{R_C}{R_{E1}}$$

Since we know the gain and the value of R_C:

$$25 = \frac{5650}{R_{E1}}$$

$$\boldsymbol{R_{E1} = 226\Omega}$$

$$\boldsymbol{R_{E2} = 2000 - 226 = 1774\Omega}$$

I will set C and C_E to both be **100µF**, since the impedance is low enough compared to R_{E2} at the lowest frequency required (20Hz).

$$X_C = \frac{1}{2\pi f C}$$

$$X_C = \frac{1}{2\pi \times 20 \times 100 \times 10^{-6}} = 80\Omega$$

Which is sufficiently low enough to bypass the AC signals.

Power dissipation in each component:

Average current through transistor.

$$P_{T1} = 3.5 \times (30 - 0.35) = 104W$$

$$P_{T2} = 3.5 \times (-30 - (-0.35)) = 104W$$

$$P_{T3} = 0.005(1.75 - 3.5 - (-20)) = 0.09W$$

$$P_D = 0.005 \times 0.7 = 3.5mW$$

$$P_{RS} = 7.07 \times 0.35 = 2.5W$$

$$P_{RE1} = (0.005)^2 \times 226 = 5.65mW$$

$$P_{RE2} = (0.005)^2 \times 1774 = 44.35mW$$

$$P_{RC} = (0.005)^2 \times 5650 = 141.25mW$$

$$P_{R1} = (0.001)^2 \times 49.3k = 49.3mW$$

$$P_{R2} = (0.001)^2 \times 10.7k = 10.7mW$$

To calculate the thermal properties of the Darlington transistors, I have decided to use the **TIP142** and **TIP 147** from ON Semiconductor for the NPN and PNP Darlington transistors respectively. [1]

Both these transistors are capable of a constant Collector current (I_C) of 10A, hence they are more than adequate for the 7A needed to achieve the needed power in the load.

Furthermore, the total power dissipation is up to **125W**, which gives some headroom above the average **104W** dissipation previously calculated.

From the specification, the Darlington transistors should not exceed a **junction temperature** of **100°C**, and the **external heatsink** should be limited to **50°C**.

From the data sheet, the junction-to-case thermal resistance is **1.0°C/W**, and we will assume a thermal washer of **$R_{C\text{-}h}$ = 0.35°C/W**.

Using the equation shown below, we can calculate the maximum thermal resistance of the heatsink necessary to achieve the **104W** power dissipation:

$$P_D = \frac{T_{j(\max)} - T_a}{R_{j-c} + R_{c-h} + R_{h-a}}$$

Where $T_{j(max)}$ is the max junction temperature, T_a is the ambient temperature, $R_{j\text{-}c}$ is the junction-case thermal resistance, $R_{c\text{-}h}$ is the case-heatsink thermal resistance and $R_{h\text{-}a}$ is the heatsink-air thermal resistance.

Re-arranging to find the maximum value of $R_{h\text{-}a}$:

$$104 = \frac{100 - 30}{1 + 0.35 + R_{h-a}}$$

$$\boldsymbol{R_{h-a} = -0.677}$$

The value obtained is **impossible**, hence it would not be appropriate to dissipate **104W through 1 transistor**. The solutions would be to run **several Darlington transistors in parallel**, say 5 for example, with a power dissipation of **20W each** to share the load. For example, if using 5 of the TIP142/TIP147 transistors instead of 1:

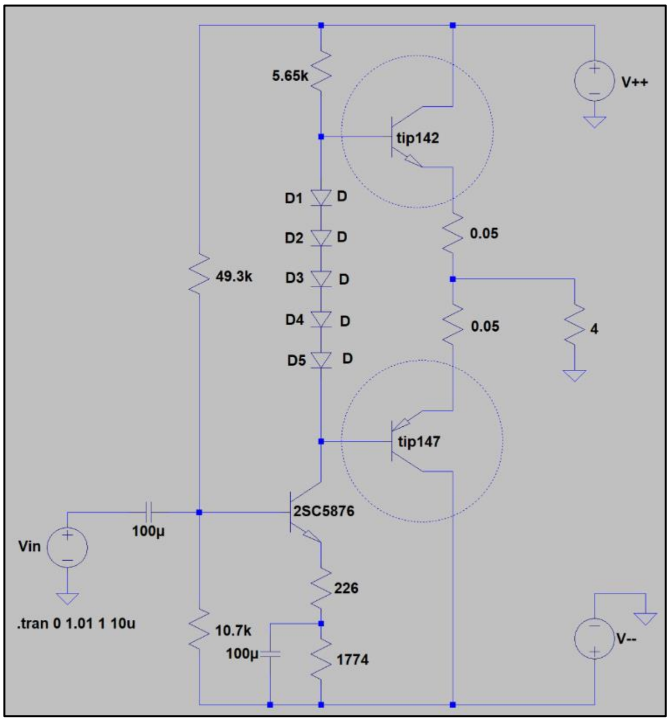
$$20 = \frac{100 - 30}{1 + 0.35 + R_{h-a}}$$

$$\boldsymbol{R_{h-a} = 2.15}$$

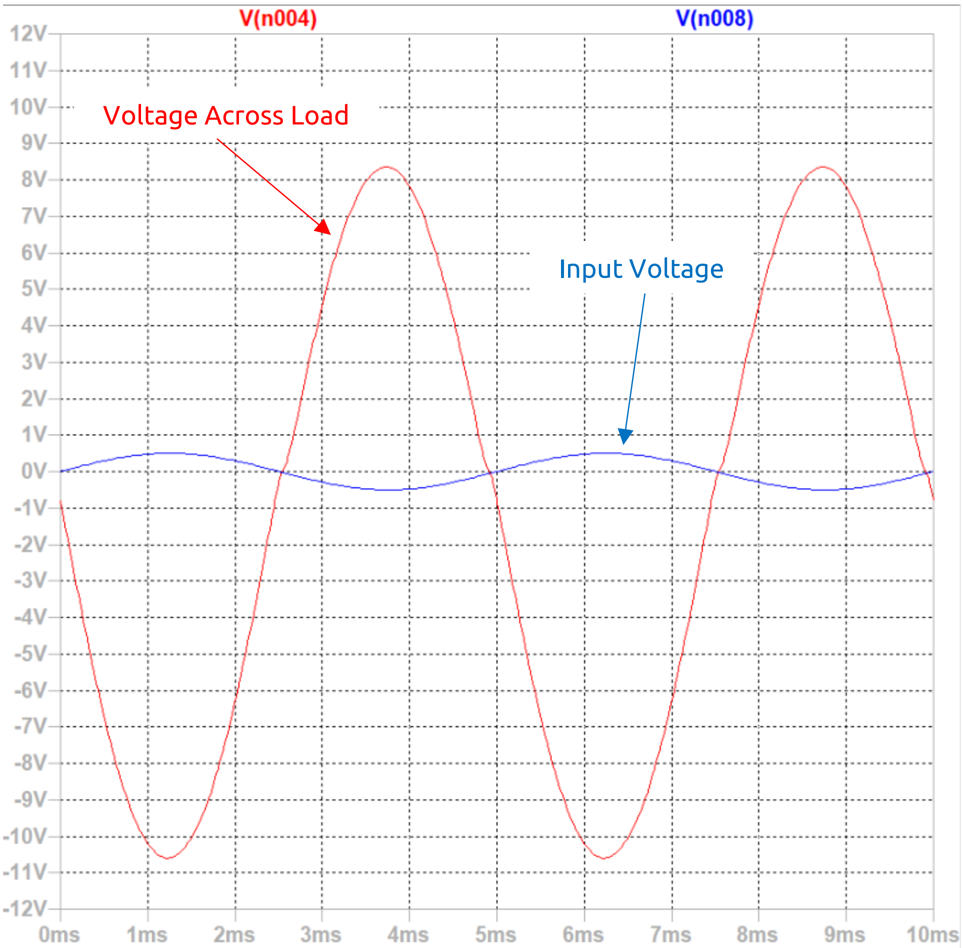
Hence, for each transistor, a heatsink with a thermal resistance **less than 2.15** will be sufficient. Realistically however, the 5 transistors would all share a common large heatsink.

Following the component values calculations, I put the circuit into **LT Spice** using models of the **TIP142/TIP147 Darlington transistors** and used a **transient analysis at 200Hz**, with a constant input voltage to check the amplifier behaved as expected.

LTSpice Schematic

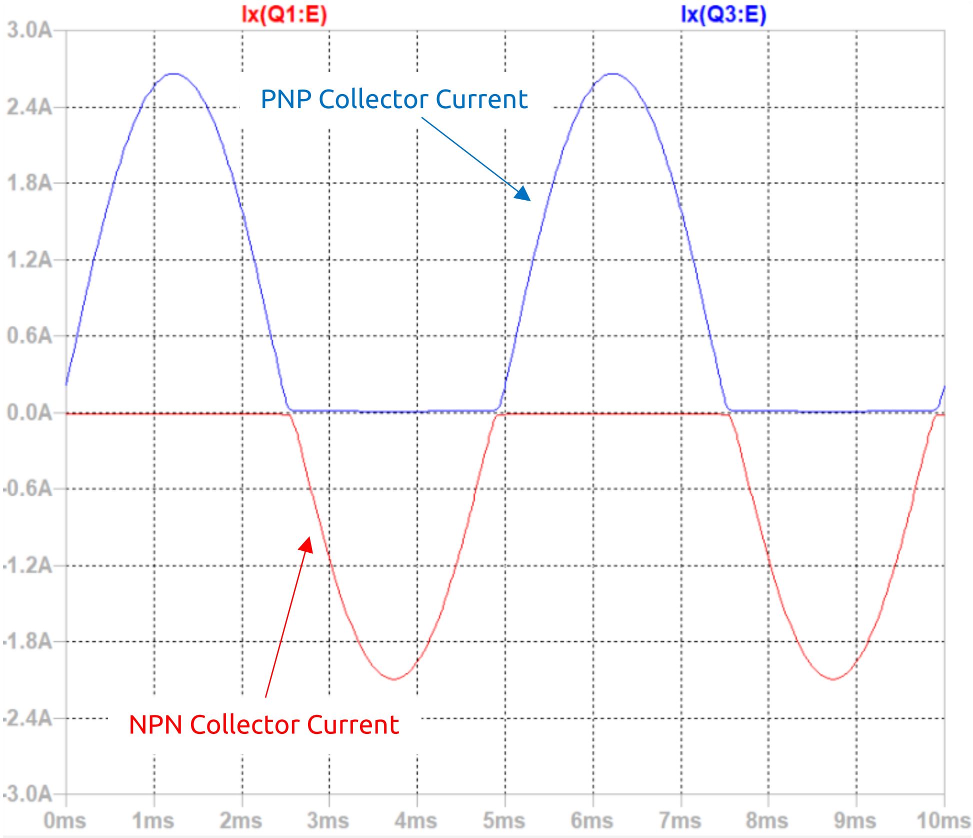


Input vs output voltage



The current through each power section transistor can be seen in the graph below, where the current switches from one transistor to the other at the 0A crossover point, but **never actually reaches 0A**. This mean the push-pull power amplifier is **successfully operating in Class A** mode since the transistors are always or slightly on.

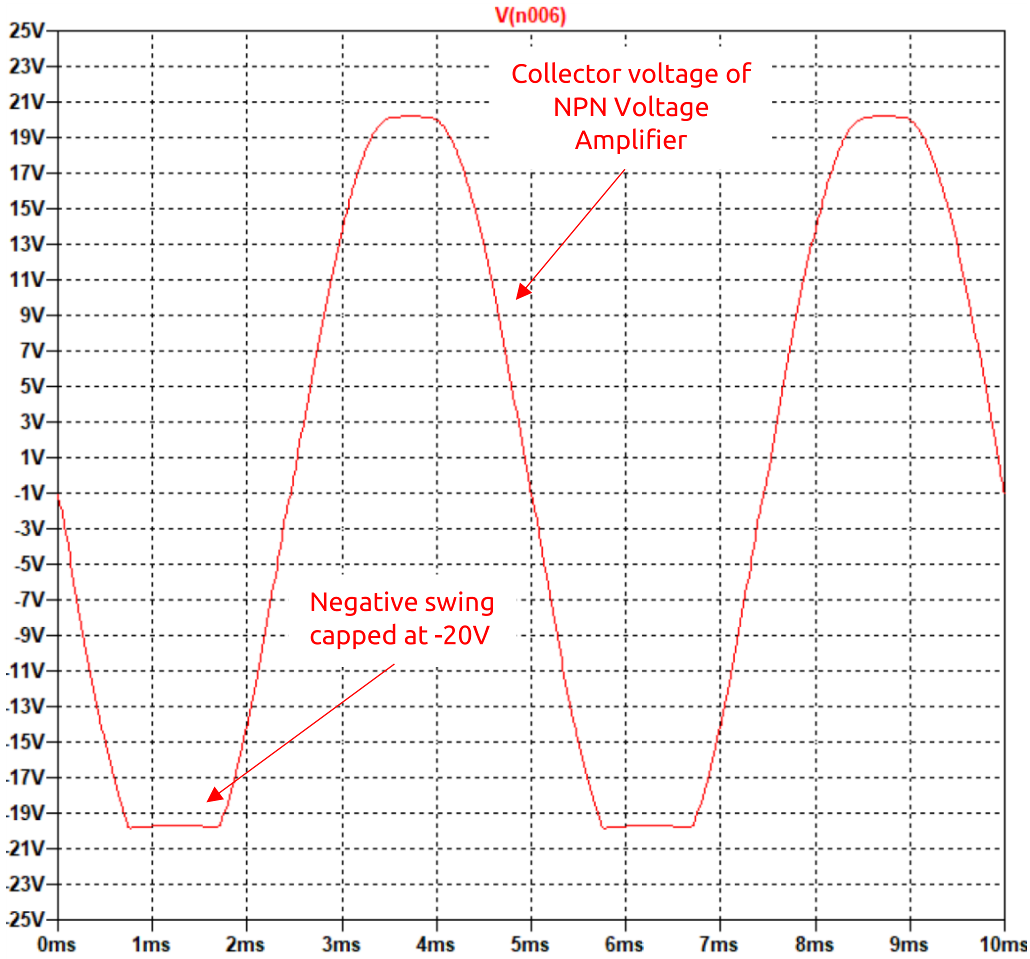
Current through each Darlington Transistor



As seen in the previous 'input vs output voltage' graph, **at low input voltages (<0.8V)**, the amplifier correctly amplifies the input voltage by approx. 21 (not quite the specified gain of 25), and there is **minimal cross-over distortion**.

However, when the voltage is **further increased (>0.8V)**, the combination of the **low rail voltages** and the **10V voltage** across the emitter resistors **limits the negative swing** of the voltage amplifier to **-20V**, as seen below.

<u>Output voltage of NPN BJT voltage amplifier</u>



Hence the amplifier is **unable to drive the load at 100W without significant distortion!**

A possible solution would be to **replace** the PNP BJT based voltage amplifier with a **dual-rail op-amp**, such as the **OPA551** by Texas Instruments, which can run on +-30V rails and wouldn't suffer from the -20V limit.

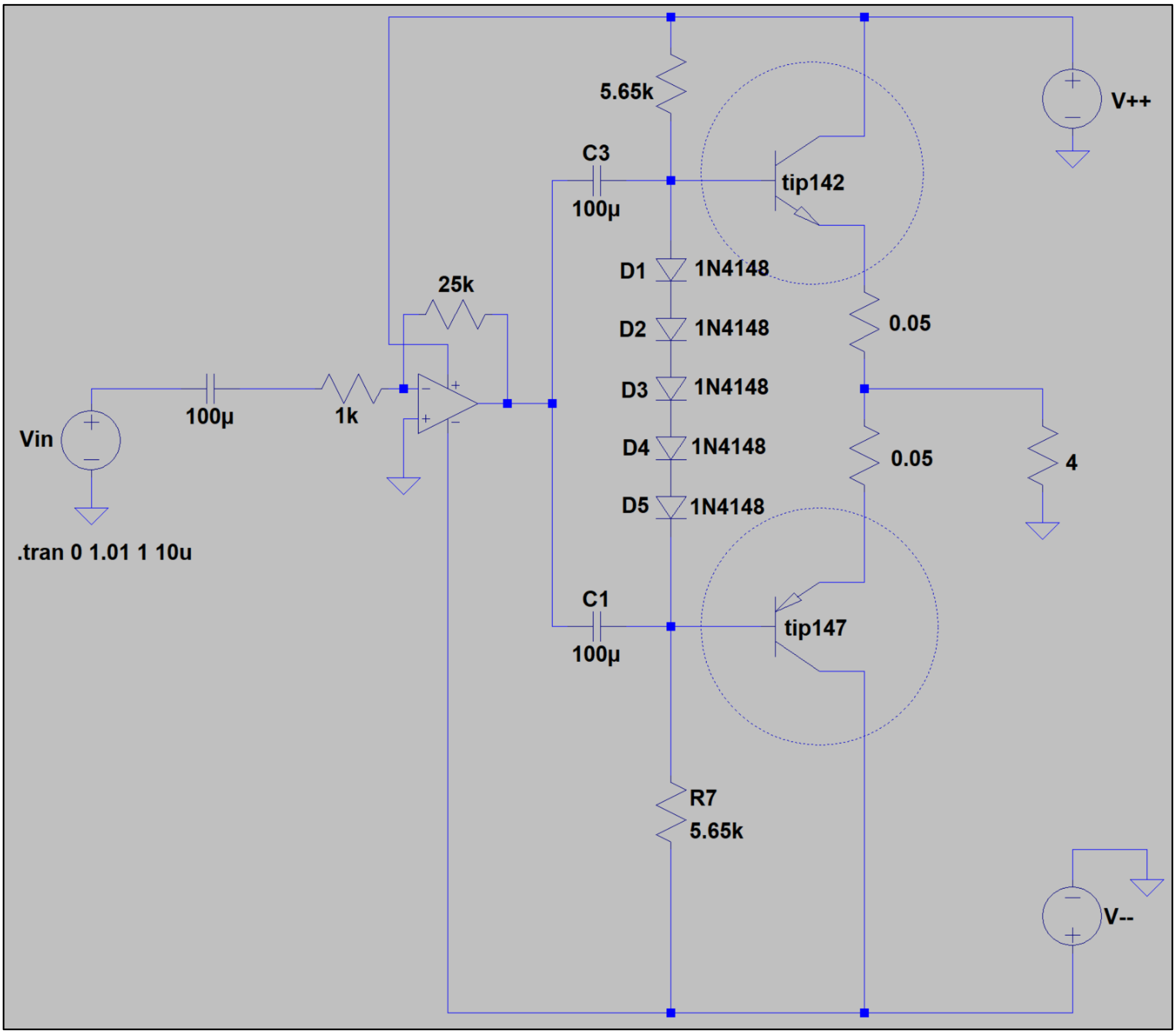
Below is the **new schematic** for the circuit with a **OPA551 op-amp** as the voltage amplifier. The **two 5.65kΩ resistors** provide the 5mA of quiescent current while making the **base voltages** of the **NPN** and **PNP** Darlington transistors **+1.75V** and **-1.75V** respectively.

The **two 100µF capacitors** act to pass the **input signal onto the bias** of the transistors inputs **(coupling capacitors)** so that each transistor can receive the signal at the necessary DC bias.

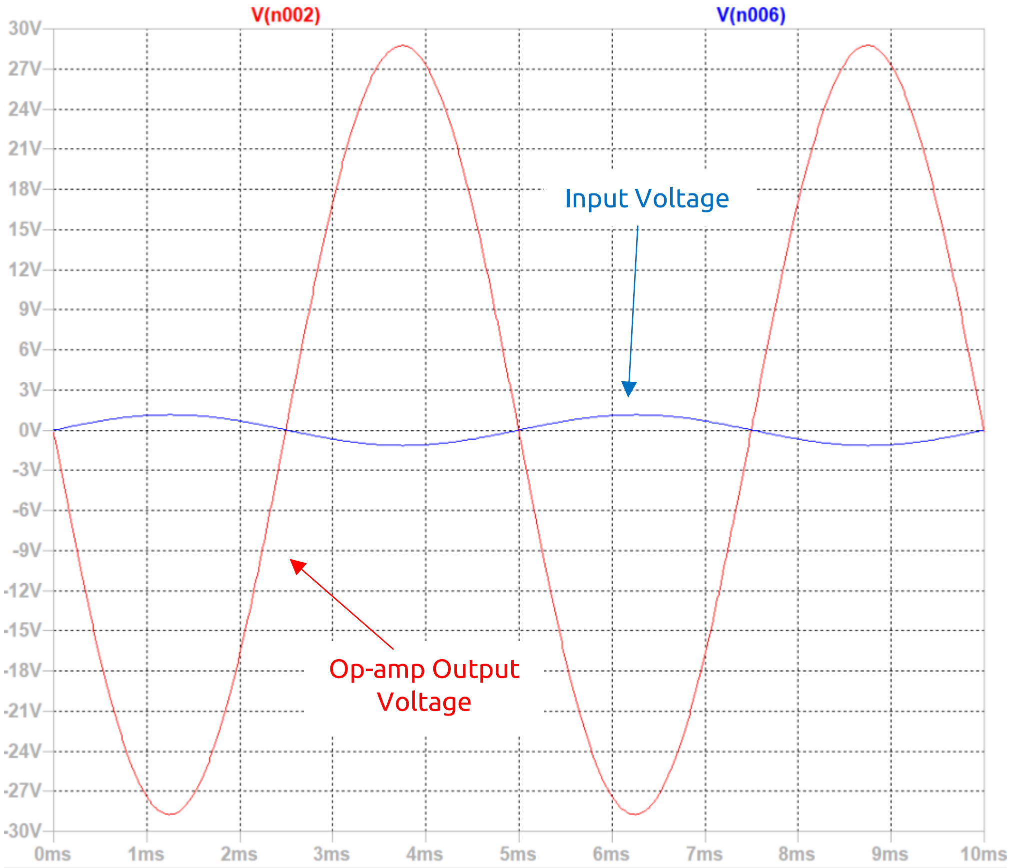
The result of these changes enables the output of the amplifier to reach the desired **+-28V** at **7A** through the loud, as shown in the figure below named 'Input vs output voltage of op-amp'.

The downside is that high voltage op-amps are not the cheapest around.

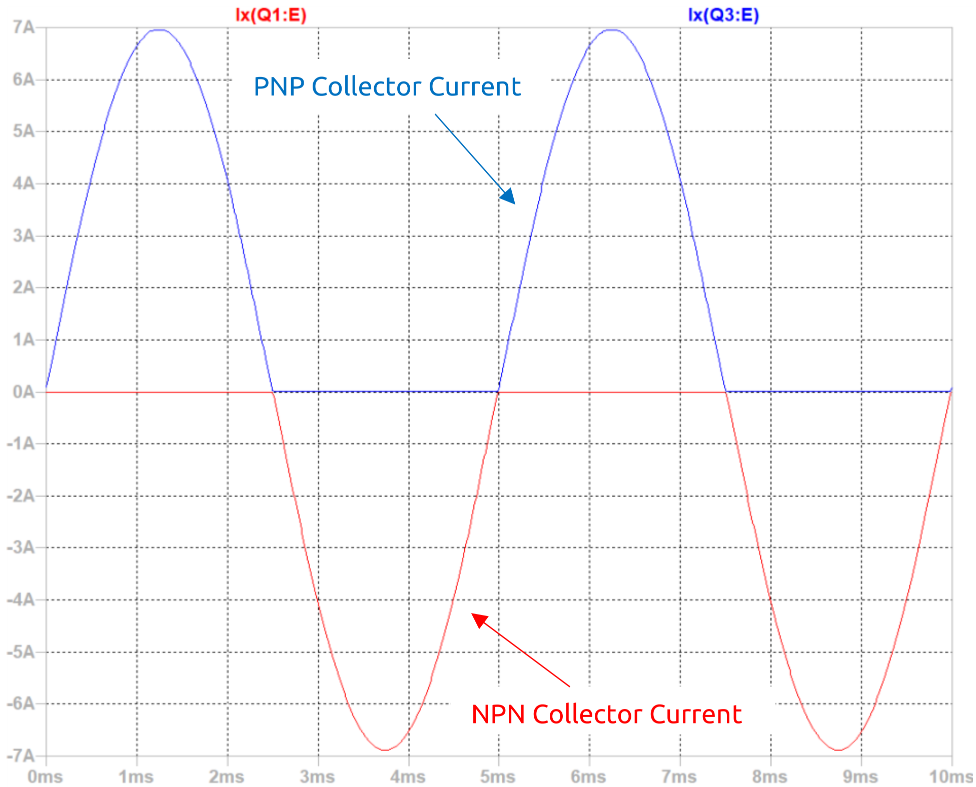
<u>New customised circuit schematic</u>



Input vs output voltage of op-amp



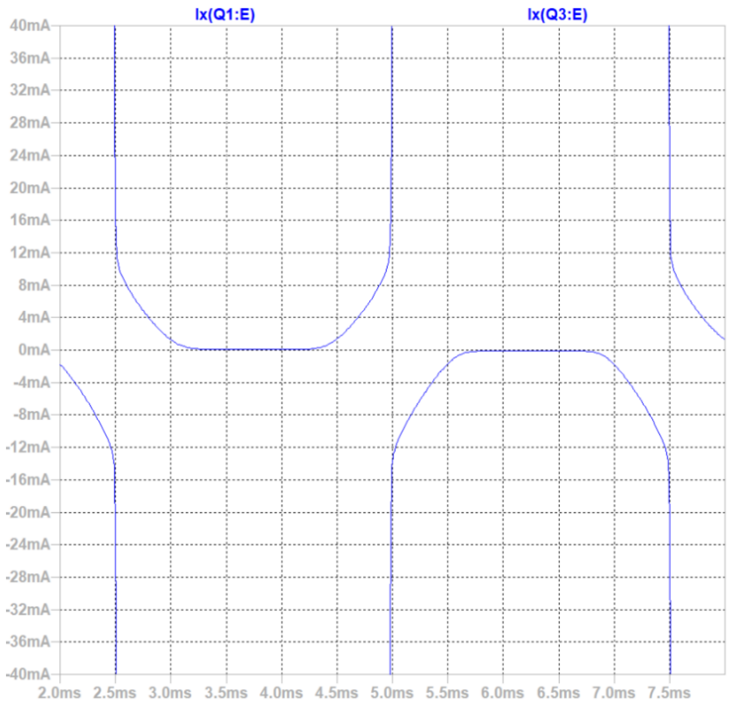
Current through each Darlington Transistor



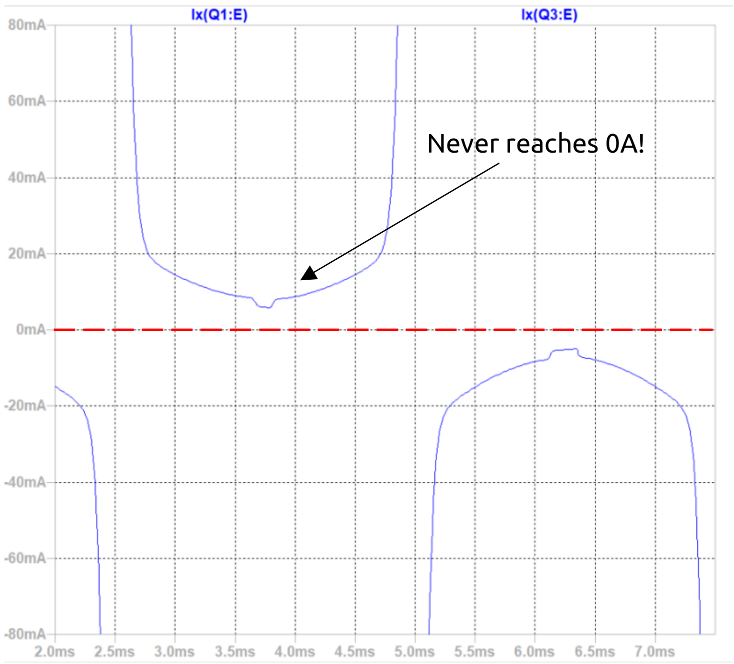
However, upon **closer inspection**, as shown in the diagram below, the current actually **does reach 0A** when the Darlington transistors are not in use – hence **not being a Class A amplifier**.

To rectify this, I put an extra diode between the two Darlington transistors to bump up the voltage difference between their bases, the result is shown below, where around **5-10mA of conduction is maintained**, hence **operating in Class A**.

Original crossover current

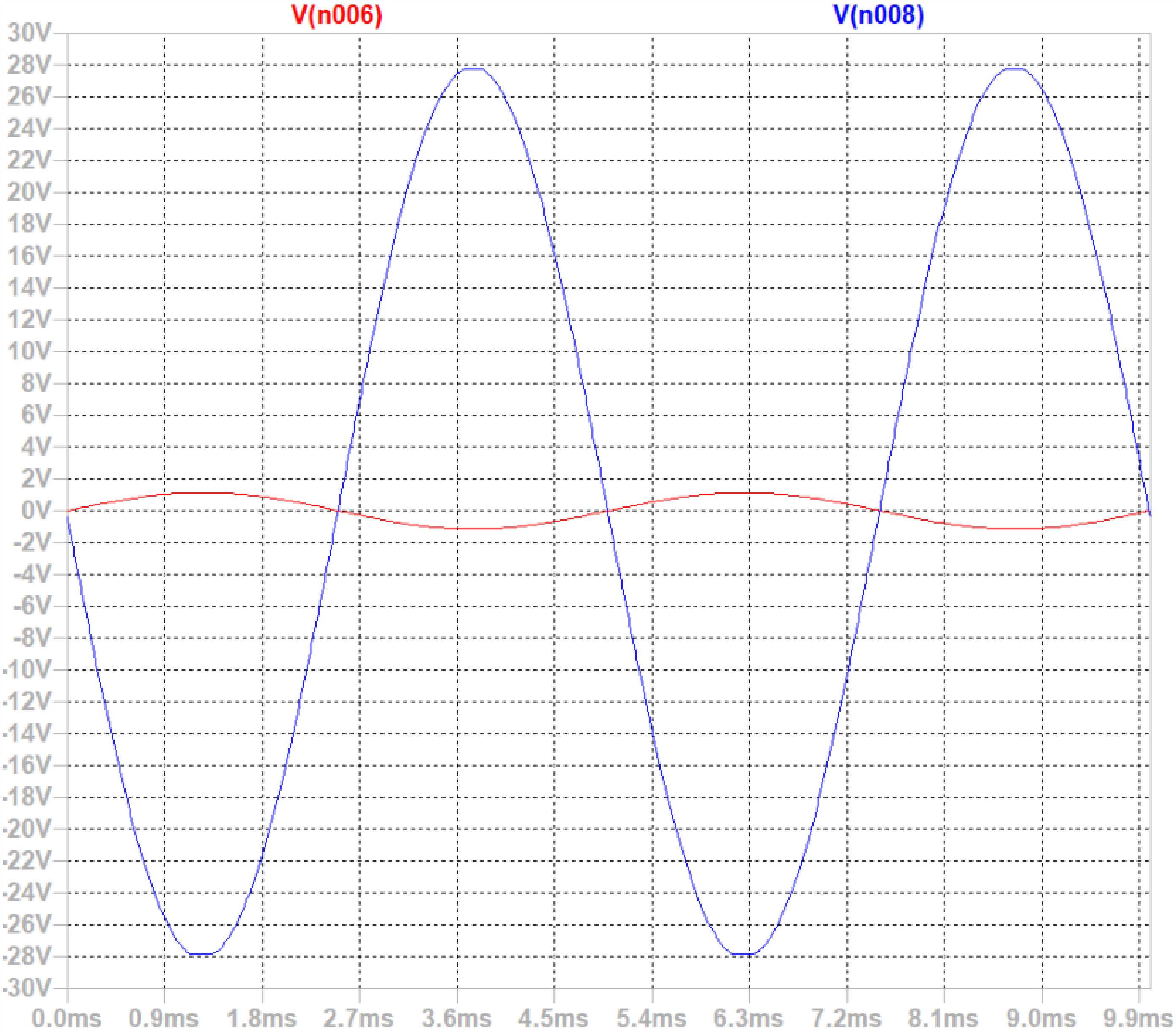


Crossover current with added diode



The figure below shows the **final circuits input voltage vs output** (voltage across load), proving that the **updates made fixed the issues** associated with the original circuit.

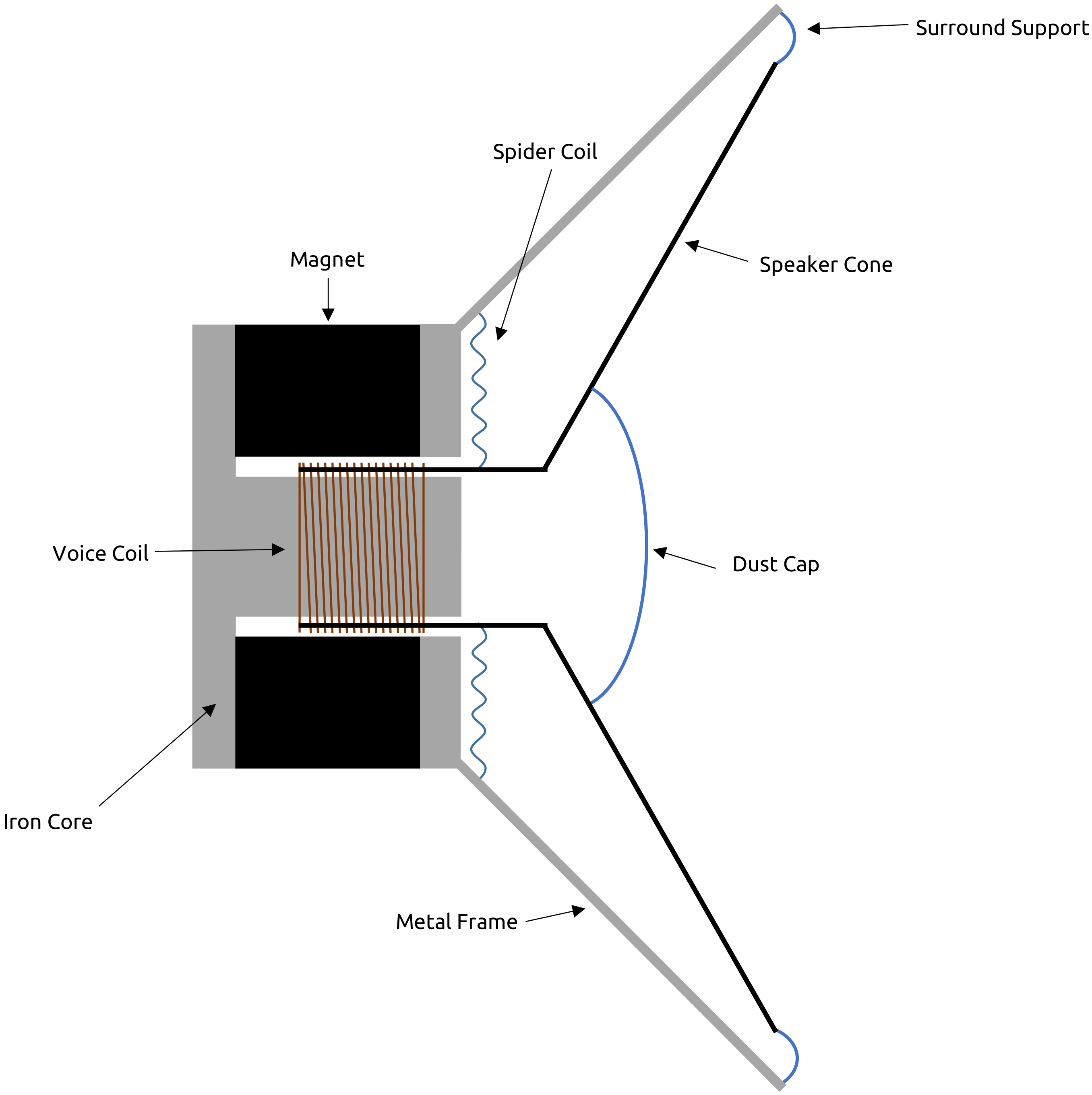
Final circuit input voltage vs output voltage



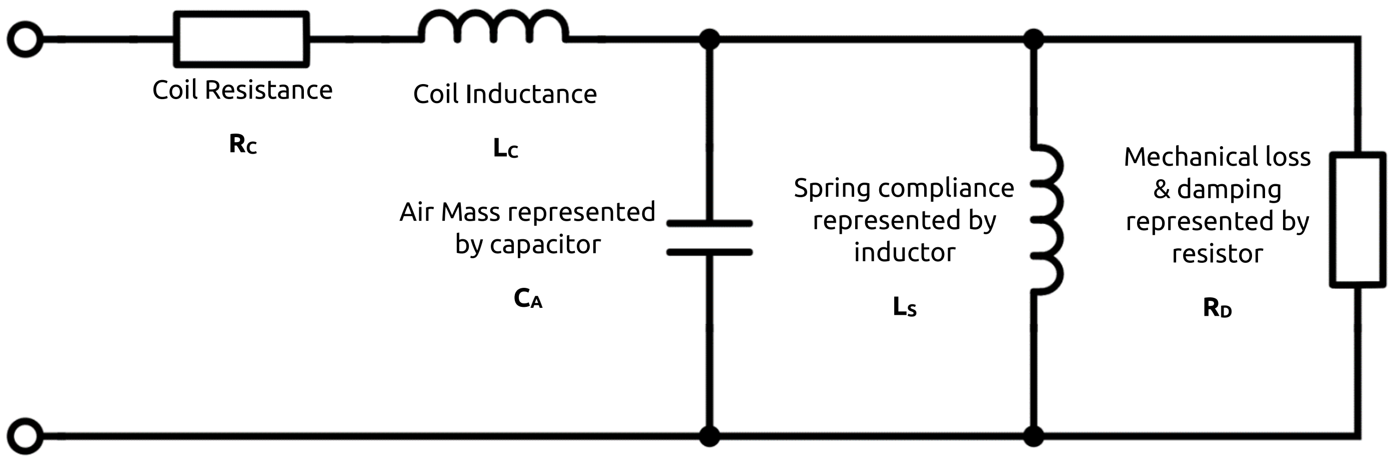
As seen, the amplifier is able to **drive the load** at **+-28V**, meaning it is able to deliver the **specified 100W of power.**

The amplifier is connected to a low frequency loudspeaker, it's specifications are shown below.

Specification	
Coil Diameter	0.05m
Flux Density in Airgap	1.2T
Spring Constant	3.55×10^{6}
Damping Coefficient	88.83
Mass of Moving System & Air	0.106kg
Number of Turns in Coil	100
Wire Diameter	0.0018m
Wire Material	Copper
Coil Inductance	10μH



The equivalent circuit for the loudspeaker is shown below:



To calculate the coil resistance (R_C), the physical resistance of the voice coil must be calculated:

$$R = \frac{\rho l}{A}$$

$$\rho_{copper} \approx 1.68 \times 10^{-8} \Omega m$$

Taking into account the wires diameter

$$l = \pi \times (0.05 + \left(\frac{0.0018}{2}\right)) \times 100 = 15.99m$$

$$A = \pi(0.0009)^2 = 2.54 \times 10^{-6} m^2$$

$$\boldsymbol{R_C = \frac{1.68 \times 10^{-8} \times 15.99}{1.96 \times 10^{-3}} = 0.1056\Omega}$$

The coil inductance (L_C), is given in the specification:

$$\boldsymbol{L_C = 10\mu H}$$

Next, the capacitance that represents the air mass (C_A) can be calculated using the electrical equivalent of:

$$I = \frac{M}{{K_e}^2} \frac{\Delta E}{\Delta t}$$

Which is:

$$I = C \frac{\Delta V}{\Delta t}$$

Hence:

$$C = \frac{M}{{K_e}^2}$$

Where the electromagnetic constant is given by:

$$K_e = BN\pi D$$

$$K_e = 1.2 \times 100 \times \pi(0.05) = 18.85$$

Solving:

$$\boldsymbol{C = \frac{0.106}{18.85^2} = 298.3\mu F}$$

The spring compliance, represented by the inductor (L_S) is given by:

$$F = \sigma_s x$$

Which can be manipulated to give:

$$E = \frac{{K_e}^2}{\sigma_s}\frac{\Delta I}{\Delta t}$$

Where an electrical equivalent can be given by:

$$V = L\frac{\Delta I}{\Delta t}$$

Hence:

$$L = \frac{{K_e}^2}{\sigma_s}$$

Where the spring compliance is given in the specification:

$$\sigma_s = 3.55 \times 10^{-6}$$

Solving:

$$\boldsymbol{L_S = \frac{(18.85)^2}{3.55 \times 10^6} = 0.1mH}$$

Lastly the mechanical loss and dampening, represented by the resistor (R_D), is given by:

$$F = K_d v$$

Which can be manipulated to give:

$$E = I\frac{{K_e}^2}{K_d}$$

Where an electrical equivalent can be given by:

$$V = IR$$

Hence:

$$R = \frac{{K_e}^2}{K_d}$$

Where the damping coefficient is given in the specification:

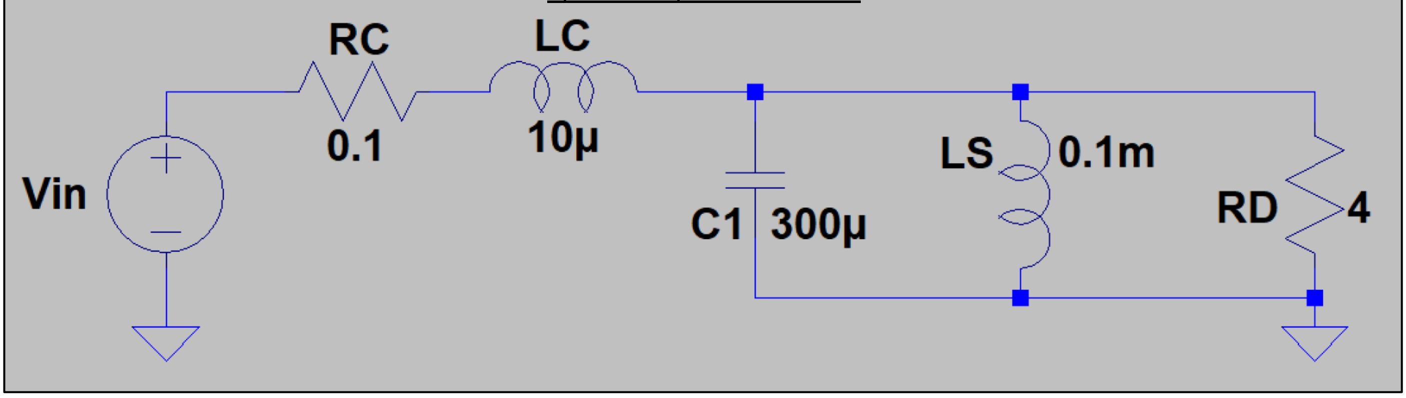
$$K_d = 88.83$$

Solving:

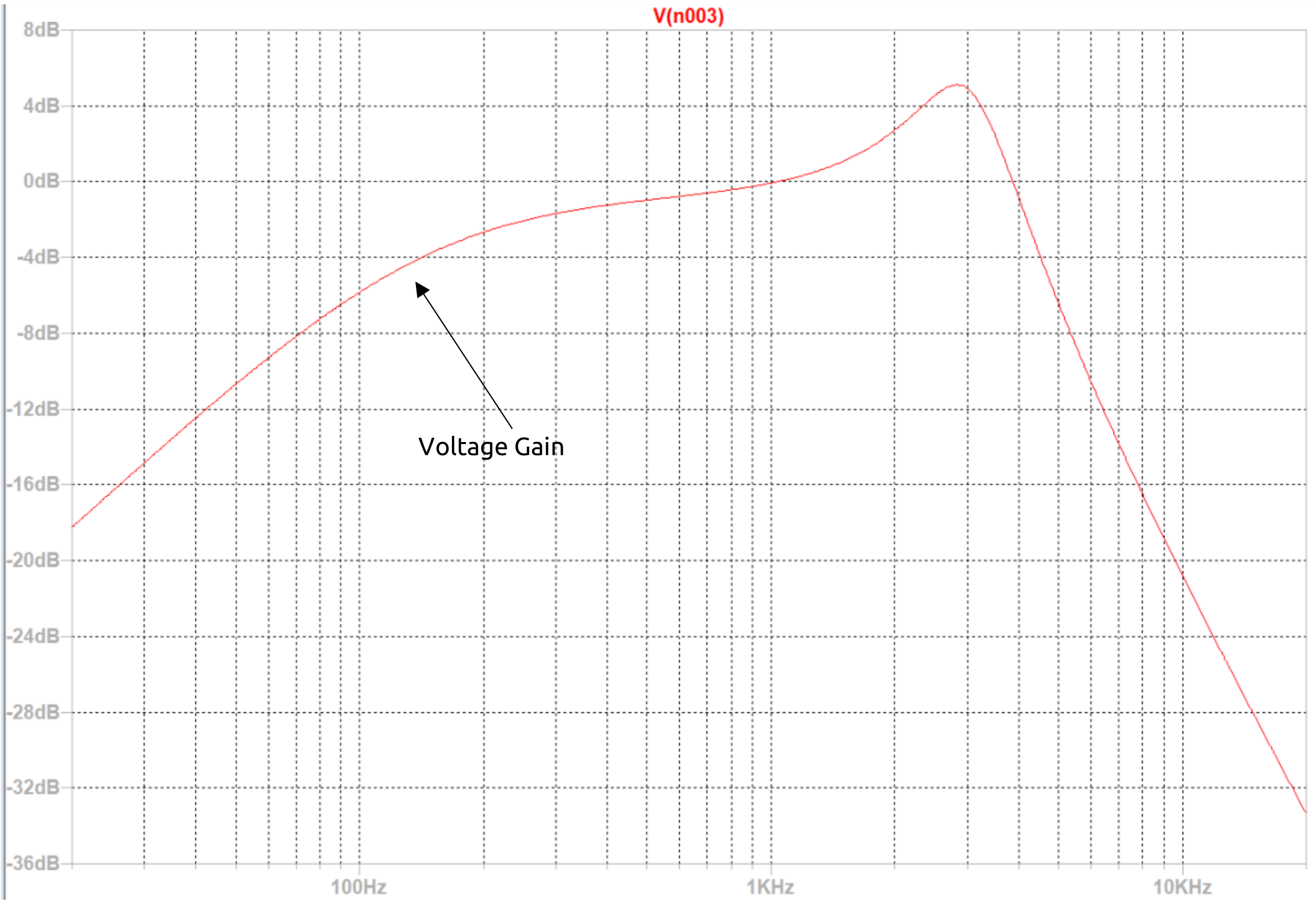
$$R = \frac{(18.85)^2}{88.83} = 4\Omega$$

The speakers equivalent circuit can now be simulated in LT Spice to find its frequency response.

Speaker equivalent circuit



Speaker frequency response



As seen from the frequency response, this speaker is **best suitable for low-mid frequencies** as the gain rapidly attenuates after 3kHz, with its most **consistent and linear frequency response being 100Hz – 1kHz.**

References

[1] TIP142 & TIP147 Data sheet: [Online]. Available: https://www.onsemi.com/pub/Collateral/TIP140-D.PDF

[2] OPA551 Data sheet: [Online]. Available: http://www.ti.com/lit/ds/symlink/opa551.pdf