SGA-SOA-1 Documentation

A Discrete Operational Amplifier For Audio Use

Samuel Groner, November 27, 2008

1 Introduction

This document introduces a discrete operational amplifier specifically designed for audio applications. It meets the following design criterions:

- good audio performance, low distortion
- low voltage and current noise, optimum source impedance 2.5 kΩ
- class A output into 600 Ω up to clipping
- unity gain stable
- offset below 20 mV
- reasonably high CMRR and PSRR
- running on ±18 V supply rails and easy to adapt for other supply voltages up to ±20 V
- output short-circuit protected
- low parts count
- no matching and selecting of parts needed

A double-sided PCB board with ground planes which follows the standard API 2520-style footprint is supplied.

2 Circuit Description

Figure 1 shows the circuit diagram of the novel operational amplifier. In the following section I will briefly discuss the design.

Q1 and Q2 form a differential input pair. Its quiescent current is set by the current source formed by Q3 and the resistive collector load R3. The component values are chosen such that the collector currents of Q1 and Q2 are equal—which results in lowest distortion and offset—and about 0.5 mA each. At this collector current the chosen transistor type shows an excellent noise figure for source impedances around 2.5 kΩ. D1 and D2 protect Q1 and Q2 from revers biasing which could otherwise result in increased noise or even failure.

R1, R2, L1 and L2 shape the transconductance (i.e. the gain) of the input stage such that open-loop gain within the audio band and stability is improved. The resulting two-pole compensation gives lower distortion and higher slew-rate than the standard single-pole compensation.

The resistive collector load of the input pair has the advantage of lower noise and simplicity over active collector loads (i.e. a current mirror). However it shows potentially decreased DC precision and slew-rate performance.

The second stage—often called *voltage amplifier stage*—is formed by Q4 and its active collector load Q5. It is run at a higher quiescent current than the input pair for best open-loop gain, stability and transient response. The gain of this stage is rolled off by Miller compensation capacitor C2 such that the unity gain crossover frequency is about 15 MHz which is a save figure for excellent stability. R8 shapes the open-loop gain response to improve phase margin. R6 is needed to limit the current through Q4 in case of a shorted output. In addition to this, it improves slew-rate symmetry. For better phase margin it is bypassed with a capacitor (C1).

The bias for the two transistors used for current sources (Q3 and Q5) is set by D3, D4 and R5. This arrangement shows sufficient PSRR and drift performance yet is simple and has low parts count.

The complementary emitter follower Q6 and Q7 forms the output stage. D5 and D6 as well as R9 and R10 set the quiescent current to about 15 mA which allows for a good class A drive of a 600 Ω load. D7 and D8 limit the output current to about 300 mA which protects the output stage transistors from overheating and failure.

C3 and C4 decouple the supply rails to prevent oscillation. The low-pass filter formed by R11 and C5 improves the negative PSRR at high frequencies in order to decrease the susceptibility to supply-related instability.

3 Parts List And Replacements

Table 1 lists the recommended parts and some possible replacements. Obviously the resistors can have higher precision and wattage, although this will not improve performance in any way. The same applies to the precision of the capacitors. Other inductors than the listed part can be used. For best noise performance and stability, they should have low DC resistance (below

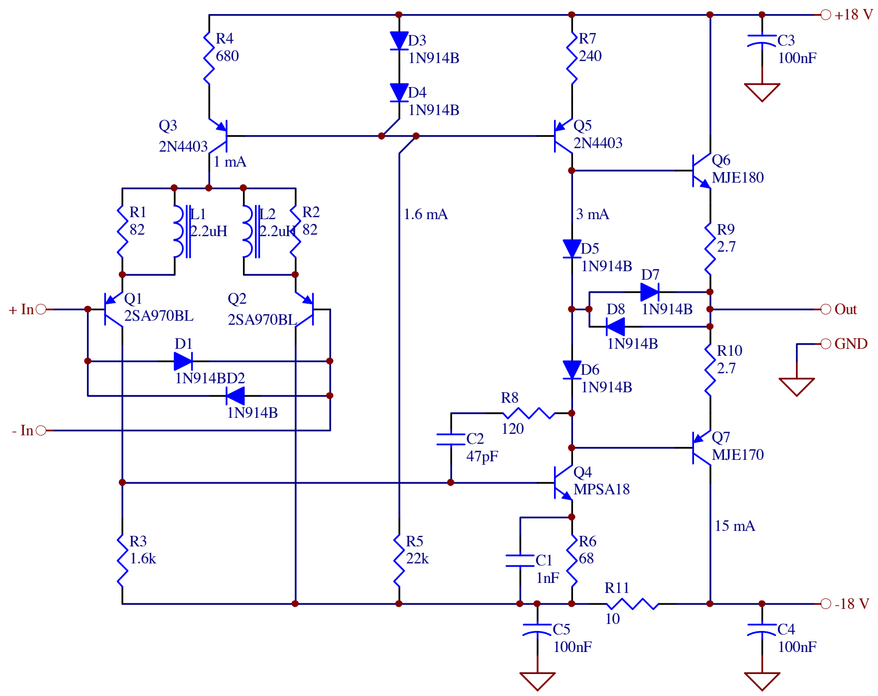


Figure 1: Schematic for the SGA-SOA-1.

Quantity	Circuit #	Identification	Description	Manufacturer	Part #	Case
2	Q1, Q2	2SA970BL	PNP transistor	Toshiba		TO-92
2	Q3, Q5	2N4403	PNP transistor	ON Semiconductor		TO-92
1	Q4	MPSA18	NPN transistor	ON Semiconductor		TO-92
1	Q6	MJE180	NPN transistor	ON Semiconductor		TO-126
		BD135	*NPN transistor*			
1	Q7	MJE170	PNP transistor	ON Semiconductor		TO-126
		BD136	*PNP transistor*			
8	D1–D8	1N914B	Diode	Fairchild		DO-35
		1N4148	*Diode*			
2	R1, R2	82 Ω	Resistor metal film, 1%, $\frac{1}{4}$ W			axial
1	R3	1.6 kΩ	Resistor metal film, 1%, $\frac{1}{4}$ W			axial
1	R4	680 Ω	Resistor metal film, 1%, $\frac{1}{4}$ W			axial
1	R5	22 kΩ	Resistor metal film, 1%, $\frac{1}{4}$ W			axial
1	R6	68 Ω	Resistor metal film, 1%, $\frac{1}{4}$ W			axial
1	R7	240 Ω	Resistor metal film, 1%, $\frac{1}{4}$ W			axial
1	R8	120 Ω	Resistor metal film, 1%, $\frac{1}{4}$ W			axial
2	R9, R10	2.7 Ω	Resistor metal film, 1%, $\frac{1}{2}$ W			axial
1	R11	10 Ω	Resistor metal film, 5%, $\frac{1}{4}$ W			axial
1	C1	1 nF	Capacitor ceramic C0G/NP0, 5%			radial, 0.1"/2.5 mm
1	C2	47 pF	Capacitor ceramic C0G/NP0, 5%			radial, 0.1"/2.5 mm
3	C3–C5	100 nF	Capacitor film metallised polyester, 10%			radial, 0.2"/5 mm
			Capacitor ceramic, X7R, 10%			
2	L1, L2	2.2 μH	Inductor, 10%	EPCOS	B78108S1222K	axial
6			Pin	Mill-Max	3103-2-00-15-00-00-08-0	
6			Socket	Mill-Max	0344-2-19-15-34-27-10-0	

Table 1: Parts list with recommended parts and manufacturers. Replacements are set in *italic*.

5 Ω), a sufficient high resonance frequency (above 10 MHz) and a precision of 10%.

The suggested transistors for Q1 and Q2 have an unique combination of very low voltage noise and high β; no replacement can be given which would be easier to source than the recommended part. The same applies to Q4 which needs very high β. If the output transistors are replaced with BD135/BD136 the highest β rating (with the suffix -16) is preferred.

I have not tested different manufacturers for the semiconductors, but I don't expect large differences in performance with other brands.

4 Adaptation For Different Supply Voltages

The presented circuit is easily adapted to different supply voltages by changing the value of R5. It should be chosen according to the following formula:

$$\mathrm{R5} = \frac{V - 1.49}{0.001569}$$

with V the total supply voltage in volts (e.g. 30 for a ±15 V supply).

Supply voltages over ±20 V are not recommended as some of the suggested transistors have rather low breakdown voltages.

5 Selecting And Matching Of Parts

Matching of the input transistor pair for β will reduce offset current and for V_{be} offset. However before the reader starts to spend endless hours with matching he should check section 10.4 and note that:

- matching will not directly improve audio performance (e.g. distortion), it increases DC precision only
- many audio circuits (e.g. about any with AC coupling) don't need better DC precision than what this circuit offers without matching
- low offset current reduces offset only with matched DC resistances at both inputs or with additional bias current compensation circuitry, which is not the case in many audio applications
- the presented circuit has pretty bad drift performance and a rather large unit-to-unit deviation in DC precision, i.e. there are other hard limits to DC precision
- the suggest Toshiba transistors are very consistent and show better inherent matching than standard transistors
- matching of β and especially V_{be} is very temperature sensitive—unless great care is given to the measurement setup and procedure the results are doubtful

No selecting or matching of other parts is necessary.

6 Heatsinking And Thermal Coupling

The circuit for the presented operational amplifier has been designed such that no heatsinking or thermal coupling of any part is necessary.

7 PCB Layout

Figure 2 shows the provided PCB design. A dual ground plane has been chosen to ensure best shielding and stability of the amplifier. The traces which carry large currents have been made much wider than the other traces. The decoupling caps C3 and C4 and the output bias diodes D5 and D6 have been placed in close proximity to the output transistors Q6 and Q7 for lowest sensitivity to parasitic oscillation and good thermal tracking respectively.

8 Stuffing Boards

The following order is recommended for stuffing and soldering of the boards: R8, C2, Q4, C1, R6, R5, D4, D3, R7, D8, R9, D6, R10, D7, Q5, D5, Q1, L1, R1, Q6, Q7, R4, Q2, R3, C5, R2, L2, pins, R11, C4, C3, D2, D1, Q3

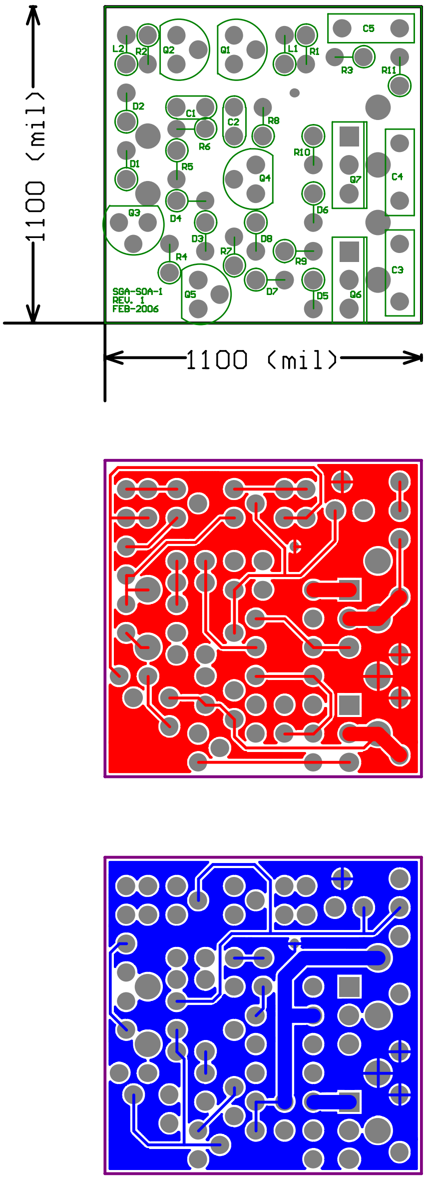


Figure 2: Overlay print (top), top layer (middle) and bottom layer (bottom) of the PCB layout. Scaled to 150%.

The pins are solder from the component side. To fix them during the soldering process it is recommended to place them in sockets soldered to a PCB board with the correct spacing (i.e. the API 2520-style footprint) or to use a jig.

The cathode of the diodes (black line on the case) is marked with a circle on the overlay print. For best space economy it is recommended to place the body of the vertically mounted resistors, inductors and diodes above the circle on the overlay print.

Work carefully—desoldering components or removing solder bridges is difficult on the crowded layout.

9 Performance

Table 2 lists some preliminary specifications. Figures 3–10 give simulated curves for open-loop gain, phase response, CMRR, PSRR and pulse response.

Voltage noise: Current noise:	1.5 $\mathrm{nV}/\sqrt{\mathrm{Hz}}$ 0.6 $\mathrm{pA}/\sqrt{\mathrm{Hz}}$
Slew rate:	10 V/μS
Output swing: Maximum output current:	± 17 V 300 mA
Quiescent current:	21 mA
Input bias current:	-1.1 μA

Table 2: Preliminary typical specifications. ± 18 V supply rails and 600 Ω load unless otherwise noted.

10 Application Notes

The SGA-SOA-1 is a voltage feedback design and can be used like standard IC operational amplifiers such as NE5532 and similar. The following sections provide additional information on how to implement audio circuits for best performance with the presented amplifier.

10.1 Decoupling

It is recommended to decouple each amplifier with a pair of low-ESR electrolytic capacitors to ground. Suitable values are between 47 μF and 220 μF. High quality types rated for 25 V and 105 °C such as Panasonic FC are rec-

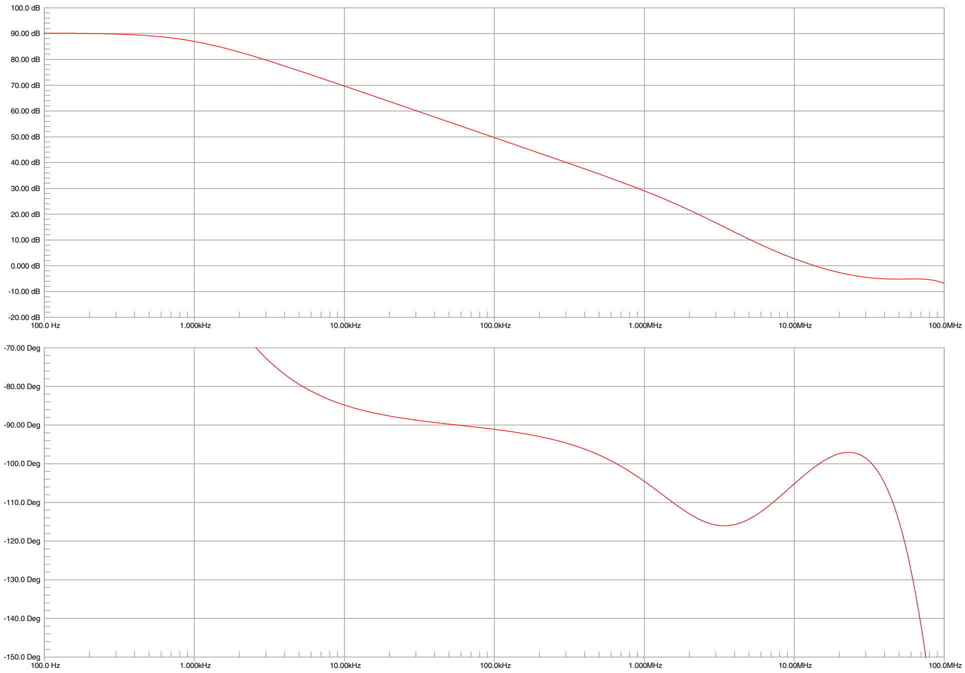


Figure 3: Simulated open-loop gain (top) and phase response (bottom).

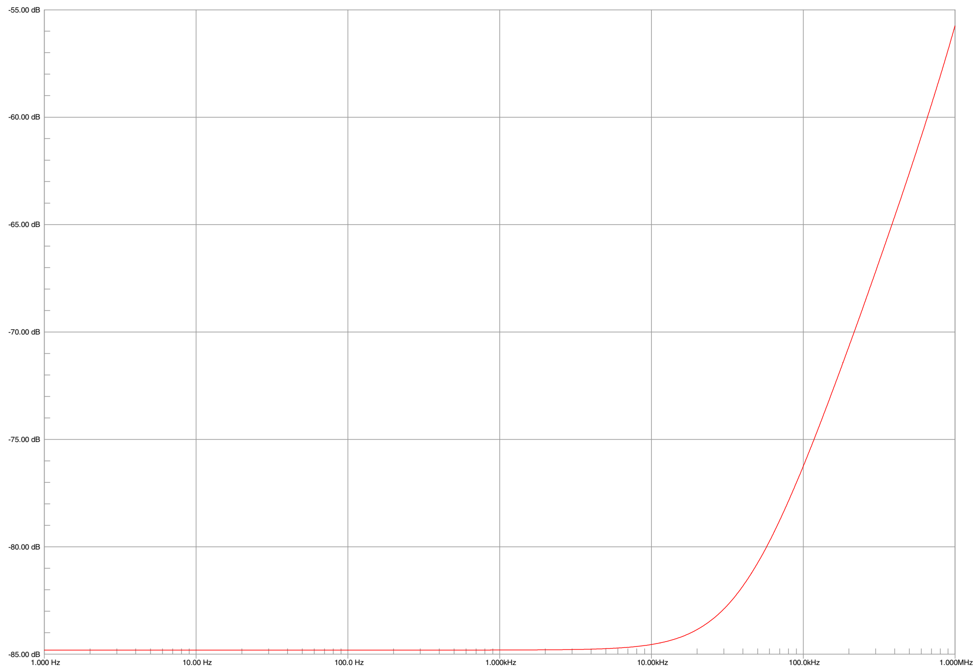


Figure 4: Simulated CMRR.

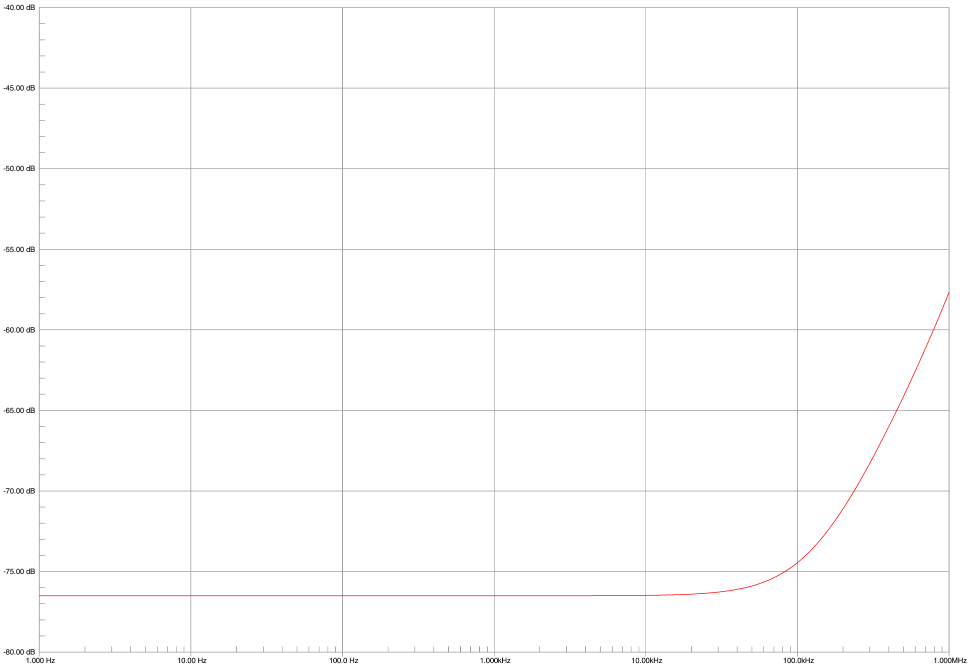


Figure 5: Simulated positive PSRR.

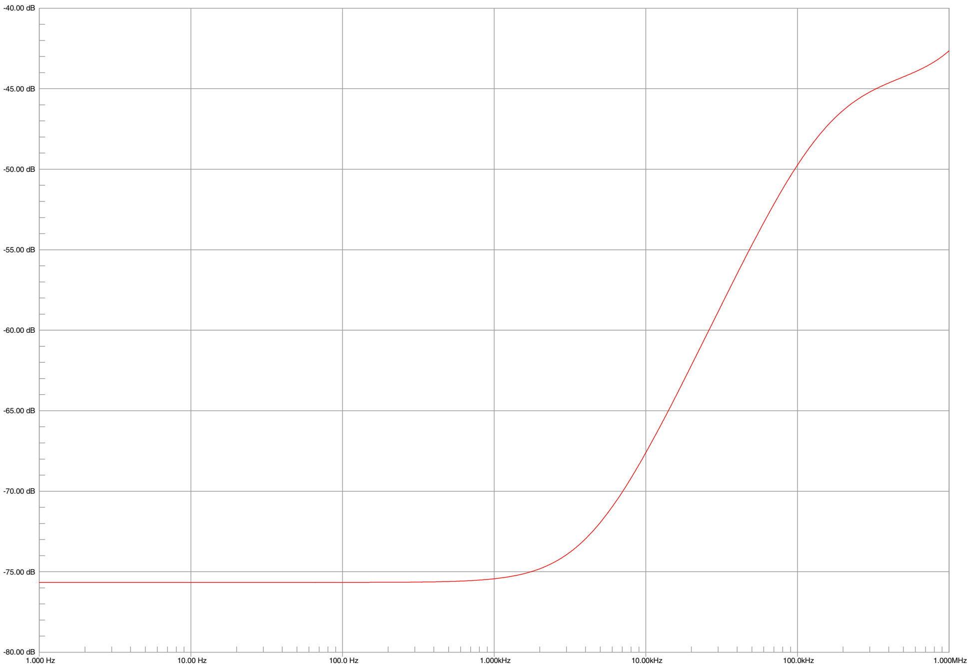


Figure 6: Simulated negative PSRR.

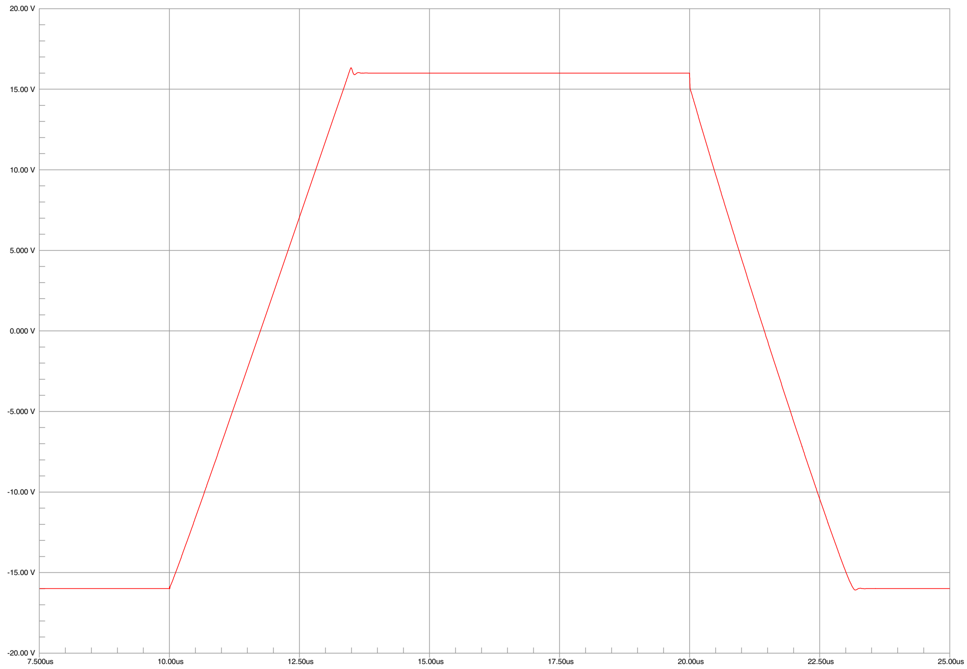


Figure 7: Simulated voltage follower large signal impulse response. R_{source} 1 kΩ.

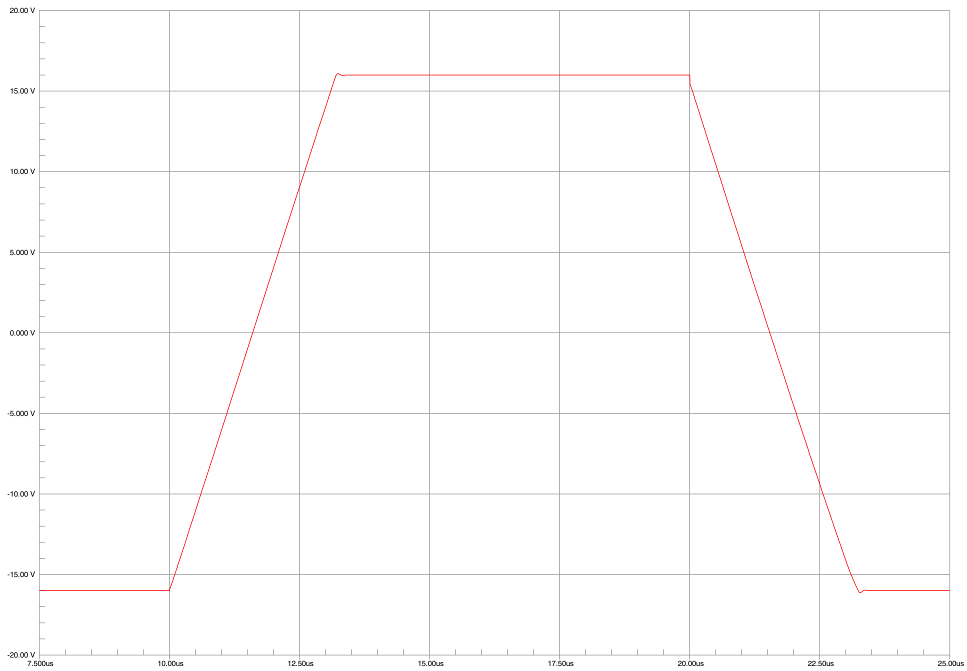


Figure 8: Simulated unity gain inverter large signal impulse response. Feedback resistors 1 kΩ, no phase lag capacitor.

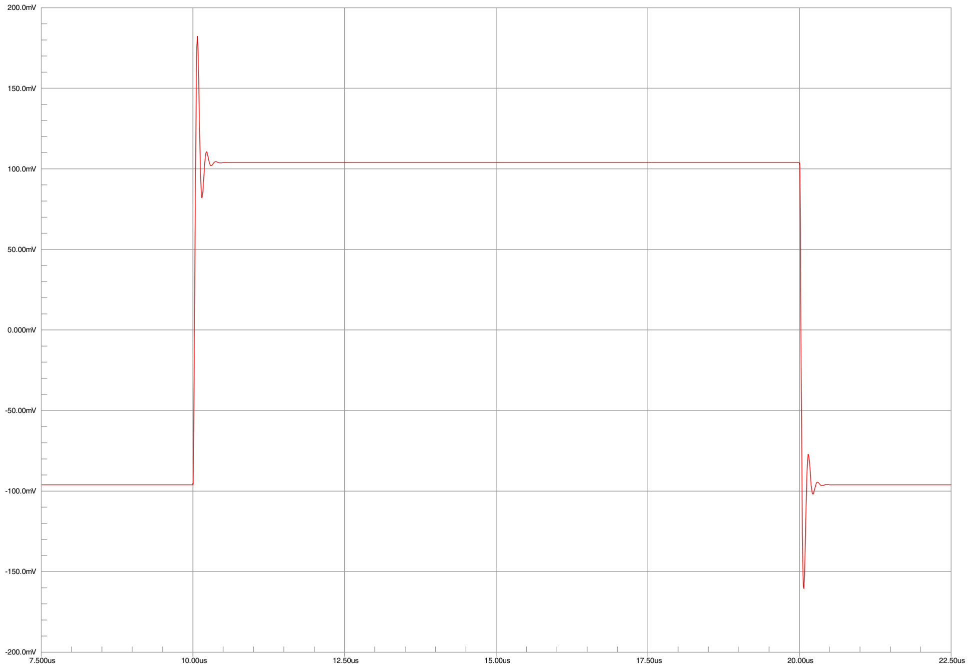


Figure 9: Simulated voltage follower small signal impulse response. R_{source} 1 kΩ.

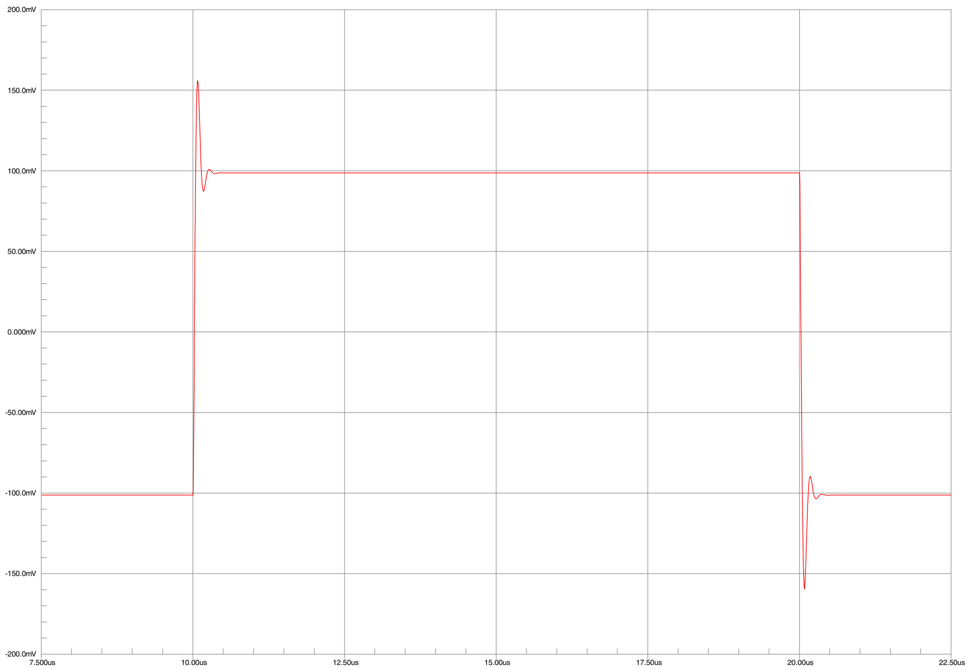


Figure 10: Simulated unity gain inverter small signal impulse response. Feedback resistors 1 kΩ, no phase lag capacitor.

ommended. No film or ceramic capacitors in parallel are necessary as they are on the amplifier board (C3 and C4).

For critical low-noise applications (e.g. first stage of a microphone preamplifier) a small series resistor of 10 Ω may be added and the value of the decoupling capacitors can be increased up to 1000 μF for best PSRR. This decoupling scheme is not recommended for applications where the load may be low (e.g. line driver) as it can increase distortion if the output stage switches to class AB.

10.2 Closed Loop Compensation And Stability

Due to the two-pole compensation the addition of a phase lag capacitor in parallel with the feedback resistor is mandatory in order to avoid overshoot in the transient response and peaking in the frequency response. The time constant of the resulting RC network should be set between 150 kHz and 500 kHz. Given a feedback resistor R_{fb} in Ω and a time constant f in Hz the phase lag capacitor C can be calculated as follows:

$$C = \frac{1}{2\pi f \cdot R_{fb}}$$

Capacitive loads (e.g. long cables) should be isolated from the amplifier output with a small resistor (above 15 Ω). The resistors should have a sufficient wattage in case the output is shorted. The required wattage can be calculated as follows:

$$W = 0.0625 \cdot R_{iso}$$

with R_{iso} the isolation resistor.

If a low impedance output is necessary (e.g. to drive a transformer with low distortion) the isolation resistor may be paralleled with an air coil inductor. The wattage of the isolation resistor can be reduced to 1 W in this case. A suitable assembled RL combination is available from Jensen Transformers as JT-OLI-3.

A ground plane for the main circuit board is recommended for most designs to provide a low impedance ground and additional shielding.

10.3 Noise

The SGA-SOA-1 has very low voltage noise and—considering the low voltage noise—low current noise. To avoid degradation of the noise performance the feedback network impedance should be kept as low as possible. This will have the additional benefit of improved DC precision. Usually the only lower limit to the feedback network impedance is set by the drive capability of the output—see section 10.5 for details.

According to the preliminary specifications the optimum source impedance is 2.5 kΩ. An input transformer suitable for a low-noise microphone preamplifier would thus have a turns ration between 1:3 and 1:6. Low DC resistances and good magnetic and static screening is desired as well.

The PSRR of this amplifier is about 75 dB within the audio band—rather low compared with typical IC operational amplifiers. Sufficient care must thus be given to the ripple and noise performance of the power supply. Standard adjustable regulators such as LM317/LM337 combined with the decoupling techniques described in section 10.1 will give good results.

10.4 DC Precision

The DC precision is mostly determined by the following contributions:

- balance of Q1/Q2 collector currents
- β mismatch of Q1/Q2
- V_{be} mismatch of Q1/Q2
- DC resistance and DC resistance balance seen at the inputs

The chosen topology has rather high unit-to-unit spread and thermal drift of collector current balance. In most audio applications this will dominate the other effects. Overall the input related offset is usually below 20 mV. This is not a problem if AC coupling is applied, but great care must be given to the headroom of DC servos.

Because the input transistors are operated at a high collector current for low voltage noise the input bias current is large with $-1.1\ \mu$A; clicks during gain changes with switched feedback networks or scratchy potentiometers may result if this current is allowed to flow through the feedback network path or potentiometer respectively. To reduce this effect, the bias current compensation scheme shown in figure 11 can be used. The scheme depicted in figure 13 provides additional offset compensation.

10.5 Load Impedance

The quiescent current of the output stage is about 15 mA. Thus the following load R_{load} can be driven in class A:

$$R_{load} = \frac{V_{out}}{0.03}$$

with V_{out} the peak output voltage. V_{out} is usually the maximum output swing of 17 V, but might be lower under certain circumstances (e.g. a two-stage amplifier where the first stage will not need full output swing as this would clip the second stage).

The presented amplifier will drive loads as low as 75 Ω to high levels. As the output stage now runs in class AB mode distortion will be much higher. If such low loads (or even lower loads) are frequently to be driven, an additional discrete output stage is recommended (see figure 13).

10.6 Microphone Preamplifier

Figure 11 shows a microphone preamplifier schematic with two SGA-SOA-1 discrete operational amplifiers. This preamplifier has been designed to provide excellent audio performance over a wide range of sources, gain settings and loads and meets the following design goals:

- EIN of −125 dBu or better at medium and high gain settings for a 200 Ω source
- consistent distortion characteristic over the entire gain range
- gain adjusted from 10 dB to 65 dB in 5 dB steps, 20 dB pad

The input transformer T1 provides debalancing of the signal and step-up of the low microphone source impedance to the optimum source impedance of the amplifier. A few suitable transformers are listed below:

- Cinemag CMMI-3.5C and CMMI-5C
- Jensen Transformer JT-11K8-APC, JT-13K6-C and JT-13K7-A
- Lundahl Transformers LL1528, LL1538, LL1530, LL1538XL, LL1550 and LL1571

The values for the zobel network (R8 and C2) and possibly R7 depend on the transformer type. If they are not provided by the manufacturer they should be chosen for best square wave response.

The gain is distributed between the two amplifiers such that the first amplifier has about twice the gain of the second. This distribution ensures both lowest noise and distortion. The values of the bandwith limiting capacitors in the feedback network have been chosen such that the bandwidth of the preamplifier remains essentially constant over the whole gain range.

R13 and R26 should be mounted close to the inverting input of the according amplifier to isolate the input from stray capacity which could otherwise result in instability. Ground planes and careful routing of the traces are recommended for best stability and noise performance.

RFI protection at the output has been added. Note the strict separation of chassis and audio ground for best RFI immunity.

High quality gold-plated switches or reed relays are recommended for good reliability.

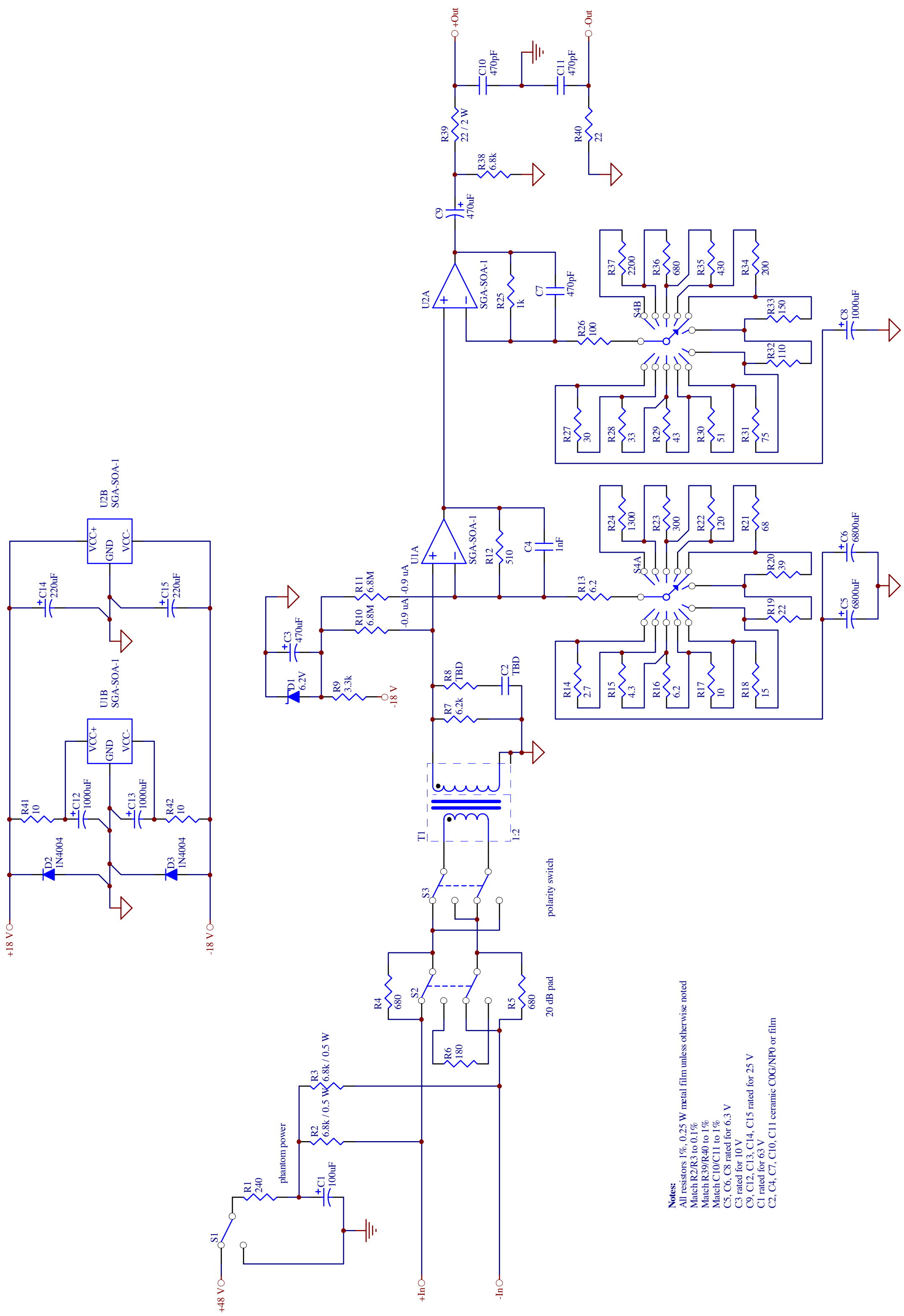


Figure 11: Microphone preamplifier schematic.

10.7 Balanced Summing Amplifier

Figure 12 depicts a schematic for a balanced summing amplifier using two SGA-SOA-1 discrete operational amplifiers. This design has been optimised to sum 24 channels with very low noise figure, wide bandwith and low distortion.

The summing is done with a passive network and a loss of about 27.6 dB. The following amplifier stage provides a make-up gain between 11.5 dB and 28 dB, covering a wide range of applications. To provide sufficient signal attenuation, unused inputs should be shorted.

To isolate the inverting inputs from stray capacity, R51 and R52 must be mounted close to the corresponding amplifier. For best CMRR the output should only be run into balanced inputs. Both input and output are well protected against RFI—note the distiction between chassis and audio ground which is necessary for effective protection.

10.8 Headphone Amplifier

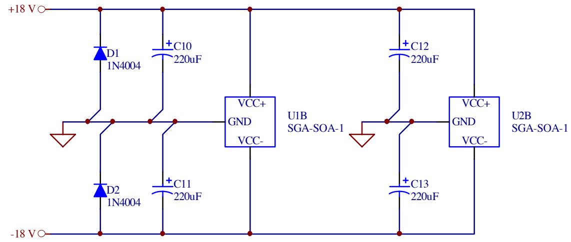
Figure 13 shows a headphone amplifier using one SGA-SOA-1 discrete operational amplifier. The circuit has been optimised to provide very low output impedance and hence excellent drive capability by the addition of a discrete output buffer.

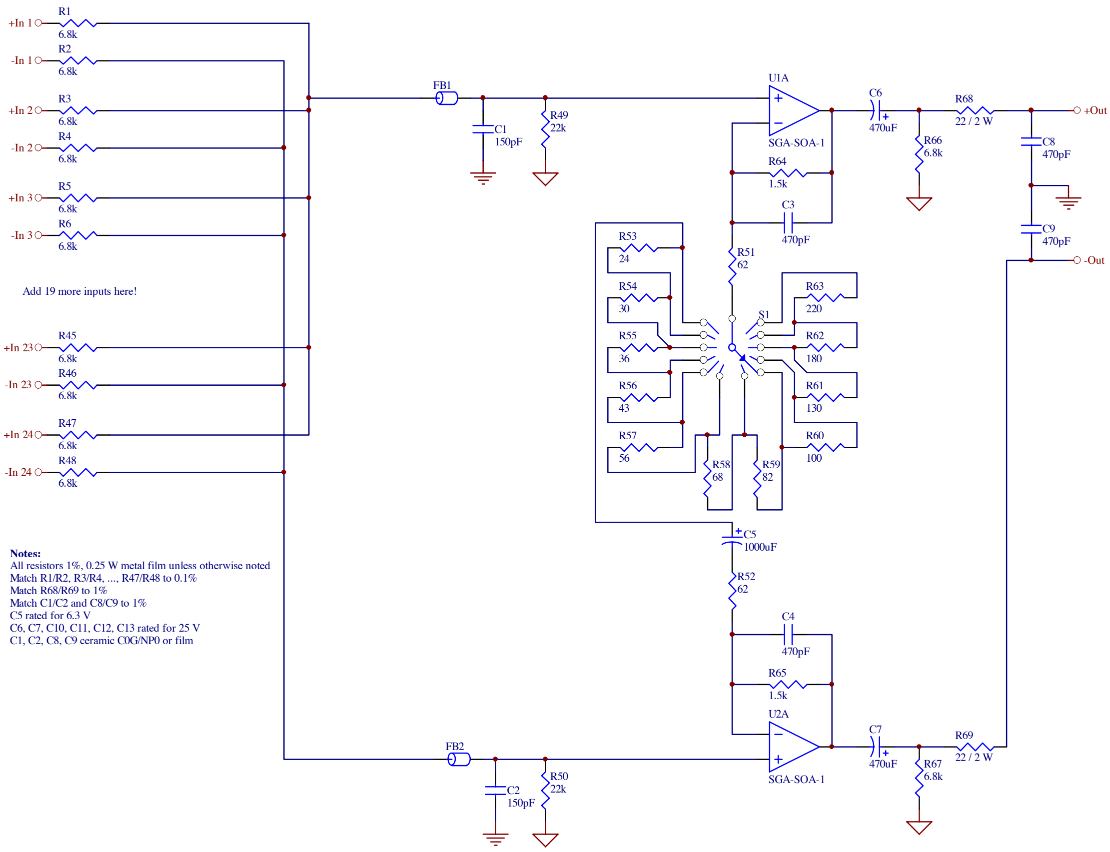
Overall gain is adjusted with a potentiometer R2 from approximately +18.5 dB down to $-\infty$ dB. To reduce current flow through R2 input bias current compensation is used. By adjusting R12 the output offset can be trimmed to zero within a range of about 30 mV (input related). A high-quality conductive plastic type is recommended for R2.

R1 and C1 protect the circuit from RFI—note that C1 should be connected to chassis rather than audio ground. If a balanced input is needed, R1 and C1 can be removed and a 1:1 line input transformer added instead. Depending on the load requirement of the transformer the value of R2 may need to be changed and/or adapted by addition of a parallel resistor. Usually a zobel network must be added as well.

The discrete output stage is biased similarly to the output stage of the operational amplifier itself. Thus the quiescent current is 15 mA as well and the short-circuit current 300 mA (which makes the output short-circuit proof without additional heatsinking). However the output impedance is now very low such that several headphones or even a small speaker can be driven with low distortion. A distribution amplifier is another recommended application.

D5 and D6 protect the circuit in case the output is forced beyond the supply rails (e.g. by inductive loads). The zobel network C5/R9 and the load isolator RL1 provide best stability into complex loads.





Notes:
All resistors 1%, 0.25 W metal film unless otherwise noted
Match R1/R2, R3/R4, ..., R47/R48 to 0.1%
Match R68/R69 to 1%
Match C1/C2 and C8/C9 to 1%
C5 rated for 6.3 V
C6, C7, C10, C11, C12, C13 rated for 25 V
C1, C2, C8, C9 ceramic C0G/NP0 or film

Figure 12: Balanced summing amplifier schematic.

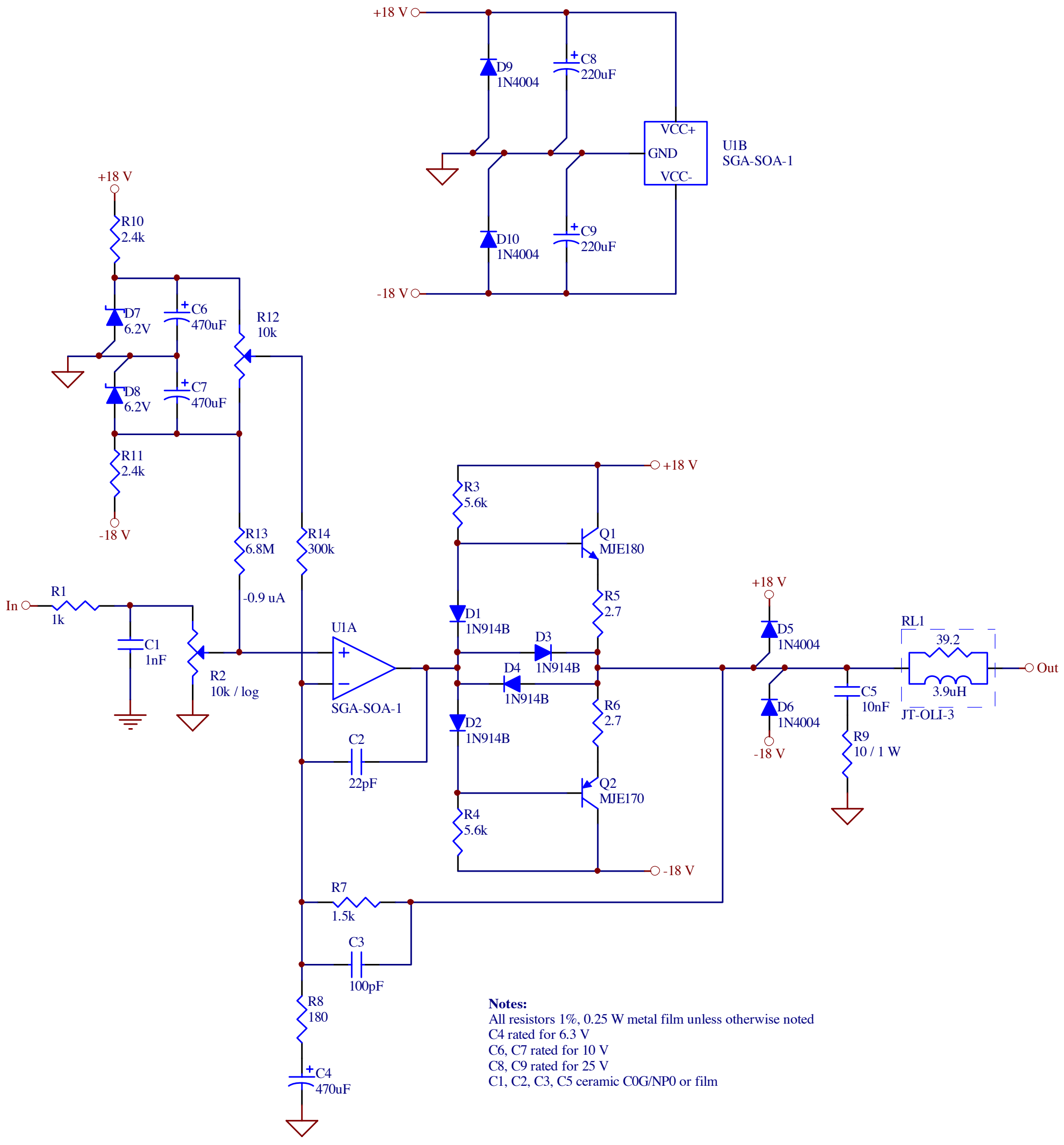


Figure 13: Headphone amplifier schematic.