*John Caldwell*

# TI Designs – Precision: Verified Design
# Single-Supply, Electret Microphone Pre-Amplifier Reference Design

![Texas Instruments]

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## Circuit Description

This single-supply microphone pre-amplifier amplifies the output signal of an electret capsule microphone to audio line levels. An op amp is used as a transimpedance amplifier to convert the output current from the microphone into a signal level voltage. The circuit is designed to be operated from a single 9V supply so it is appropriate for battery operated systems.
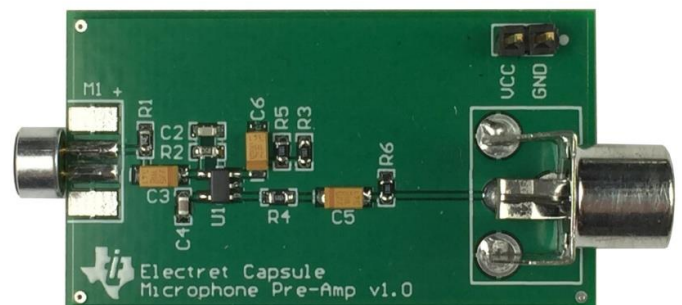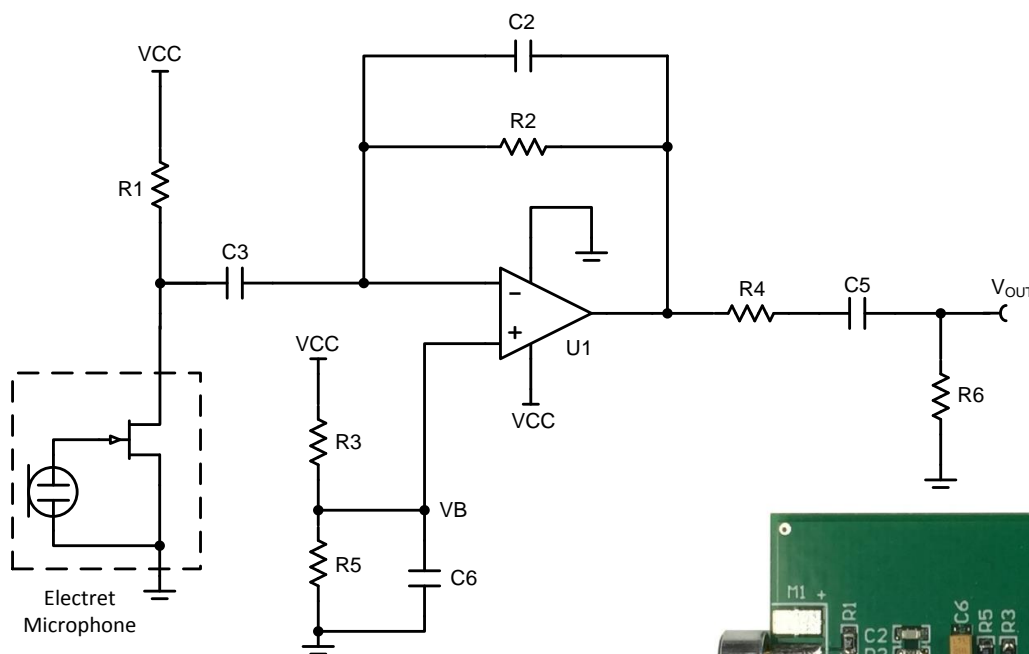
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# 1    Design Summary

The design requirements are as follows:

Power Supply Voltage: 9V

Power Supply Current: <3mA

The design goals and performance are summarized in Table 1.

**Table 1. Comparison of Design Goals, Simulation, and Measured Performance**

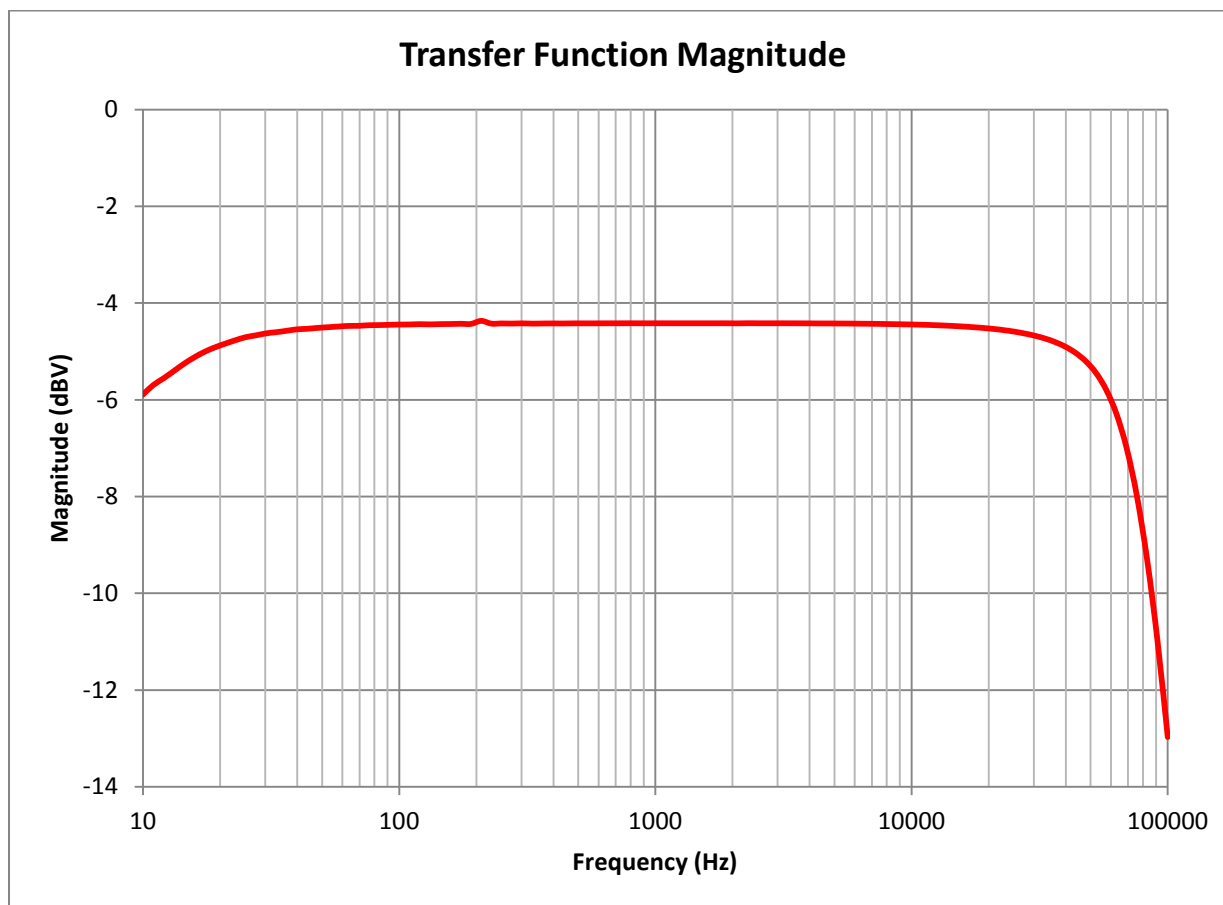|  | Goal | Simulated | Measured |
|---|---|---|---|
| **SNR (94 dB SPL)** | 68dB | 68.09dB | 67.94dB |
| **Output Level (94 dB SPL)** | 0.606Vrms | 0.6058Vrms | 0.6065Vrms |
| **Gain Deviation (20Hz – 20kHz)** | -0.5dB | -0.3dB | -.423dB(20Hz) |



**Figure 1: Measured transfer function of the microphone pre-amplifier circuit**

## 2 Theory of Operation

Electret microphones are very common in personal electronics due to their small size, excellent frequency response, and reasonable cost [1]. An "electret" is a thin, Teflon-like material with a fixed charge bonded to its surface [1]. The electret is housed between two electrodes, and the structure forms a capacitor which contains a fixed charge. Air pressure variations (sound waves) move one of the electrodes of the capacitor back and forth, changing the distance between the two electrodes, and modulating the capacitance of the structure. Because the charge on the microphone is fixed, varying the capacitance causes the voltage on the capacitor to also change, satisfying the equation:

$$Q = C * V \qquad (1)$$

Where Q is charge, C is capacitance, and V is voltage. Therefore the microphone capacitor acts as an ac-coupled voltage source. Because the charge on the microphone capacitor must be fixed, the amplifier circuitry directly in contact with it must have extremely high input impedance. Most electret microphones have an internal JFET which buffers the microphone capacitor. The voltage signal produced by sound modulates the gate voltage of the JFET, labeled $V_G$ in Figure 2 causing a change in the current flowing between the drain and source of the JFET ($I_{MIC}$). An extremely high resistance, $R_G$, may be included to bias the gate of the JFET.
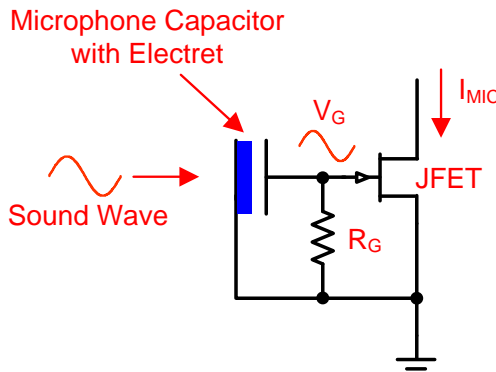


**Figure 2: A simplified circuit schematic of an electret microphone**

An example construction of an electret microphone is shown in Figure 3. One electrode of the capacitor is formed by a metallization layer on the charged polymer film. The metallization layer on the film is connected to the microphone case by a metal washer, and the microphone case is typically connected to the source terminal of the internal JFET. The other plate of the capacitor is formed by a metal back plate, separated from the film by a plastic washer, and connected to the gate of the JFET. Sound waves deform the metalized film, effectively changing the distance between the two capacitor plates and producing a voltage.
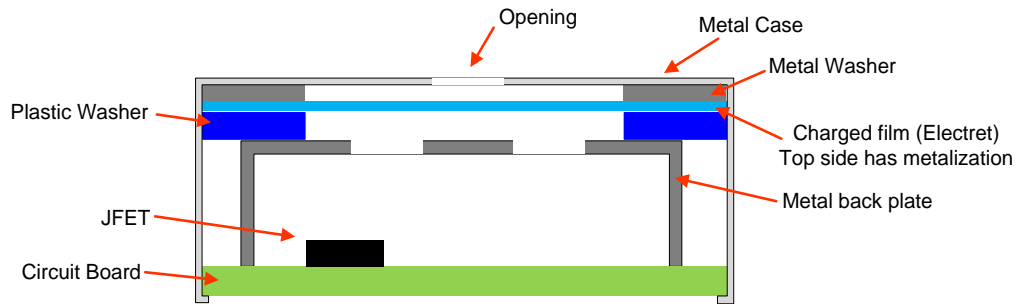


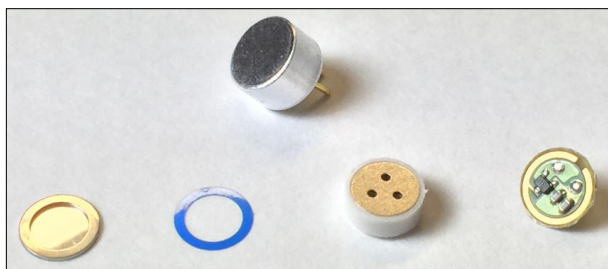**Figure 3: Cutaway diagram of an example electret microphone**

**Figure 4: The case of an electret microphone (top) was removed to show internal components. From left to right: metalized film and metal washer, plastic washer, metal back plate, and PCB with JFET.**

The basic schematic of the pre-amplifier is shown in Figure 5. To understand its operation consider that the current in the microphone ($I_{MIC}$) has a dc component ($I_{dc}$) necessary to bias the internal JFET, and an ac component ($I_{ac}$) caused by sound waves. If the impedance of capacitor C3 is much less than R1 at audio frequencies, then $I_{ac}$ will flow through C3 and not R1. Op amp U1 acts as a transimpedance amplifier, and attempts to hold its inverting input at a constant voltage ($V_B$) by varying its output. The output voltage of the op amp ($V_{OA}$) will be:

$$V_{OA} = I_{ac}R_2 + V_B \qquad (2)$$

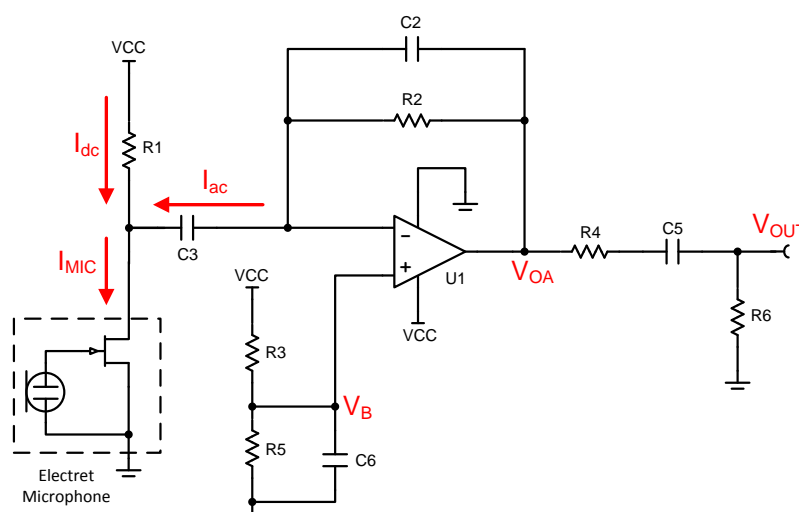Finally, the dc component of the output signal is removed by capacitor C5.



**Figure 5: A simplified schematic of the microphone pre-amplifiers with selected current pathways and voltages labeled.**

This topology was selected for a few reasons. First, it allows for single-supply operation to be easily accommodated by biasing the non-inverting input of the op amp to the mid-supply point. Second, the gain of the pre-amp is determined by R2 but the noise gain of the op amp is determined by the ratio of R2 to R1. Therefore it is possible to achieve lower noise with this topology than with a non-inverting amplifier. Finally, because capacitor C3 is chosen to have a very low impedance at audio frequencies, the voltage at the drain of the microphone JFET varies very little, potentially reducing distortion caused by channel length modulation in the JFET.

## 2.1 Gain Calculation

The parameters of the POM-3535P-R microphone selected for this design are shown in Table 2. The typical microphone sensitivity will be used to calculate the gain of the amplifier. First, the dB value must be converted to a linear value for these calculations. Microphone sensitivity is given as a dB value relative to 1V, measured at 94dB SPL (1 Pascal). Therefore the sensitivity of the microphone in volts per Pascal of air pressure is:

$$10^{\frac{-35dB}{20}} = 17.78mV/Pa \tag{3}$$

**Table 2: Selected parameters from microphone datasheet[2]**

| Parameter | Value | Units |
|---|---|---|
| Sensitivity | -35 +/- 4 | dBV |
| Standard Operating Voltage | 2 | Vdc |
| Current Consumption (Max) | 0.5 | mA |
| Impedance | 2.2 | kOhm |
| Signal to Noise Ratio (Min) | 68 | dB |

However, because the pre-amplifier used in this reference design is a transimpedance type, this must be converted to a value of current per Pascal of air pressure. Most likely, the microphone sensitivity was measured using a 2.2k ohm impedance as indicated in the microphone specification table. The output current per Pascal of air pressure will be:

$$\frac{17.78mV/Pa}{2.2k\Omega} = 8.083\mu A/Pa \tag{4}$$

The gain calculation depends on the maximum sound pressure level expected at the microphone input. For this design, we will use 100dB SPL as the maximum sound pressure level expected, and map this sound pressure level to typical line level audio levels (1.228Vrms). 100dB SPL is an air pressure of 2 Pa, giving a microphone output current of

$$\frac{8.083\mu A}{Pa} * 2Pa = 16.166\mu A \tag{5}$$

The gain calculation for the transimpedance amplifier is:

$$V_{OUT} = I_{Iac} * R_2 \rightarrow R_2 = \frac{V_{OUT}}{I_{IN}} = \frac{1.228V}{16.166\mu A} = 75961.1\Omega \rightarrow 75k\Omega \tag{6}$$

The feedback capacitor C2 compensates for parasitic capacitance at the op amp inverting input which can cause instability. Capacitor C2 also forms a pole with resistor R2 in the response of the pre-amplifier. The frequency of this pole must be high enough to not affect the microphone transfer function within the audible bandwidth. For this design, a response deviation of -0.1dB at 20kHz is acceptable. The location of the pole can be calculated using the relative gain at 20kHz:

$$f_p = \frac{f}{\sqrt{\left(\frac{G_0}{G_f}\right)^2 - 1}} = \frac{20kHz}{\sqrt{\left(\frac{1}{0.989}\right)^2 - 1}} = 133725Hz \tag{7}$$

In the above equation, $G_0$ and $G_f$ are the gains at low frequency and the gain at frequency "f" respectively. Inserting 20kHz for "f", and 0.989 (-0.1dB) for $G_f$, gives a pole frequency of 133725Hz. The feedback capacitor value can then be calculated:

$$C_2 = \frac{1}{2\pi f_P R_2} = \frac{1}{2\pi (133725Hz)(75k\Omega)} = 15.87pF \rightarrow 15pF \tag{8}$$

## 2.2 Microphone Bias Resistor and Coupling Capacitor

The internal JFET of the electret microphone is biased by resistor R1. The value of this resistor can be calculated from the desired supply voltage ($V_{CC}$), and the microphone operating voltage ($V_{MIC}$) and current consumption ($I_S$) given in Table 2:

$$R_1 = \frac{V_{CC} - V_{MIC}}{I_S} = \frac{9V - 2V}{0.5mA} = 14k\Omega \rightarrow 13.7k\Omega \qquad (9)$$

A value for R1 slightly less than the calculated value (13.7kΩ as opposed to 14kΩ) is used to accommodate variation in the supply voltage.

For this pre-amplifier design, it is beneficial to have the largest value for R1 possible for two reasons. First, the noise gain of op amp U1 is:

$$A_N = 1 + \frac{R_2}{R_1} \qquad (10)$$

But the signal gain is directly determined by R2. Therefore, increasing the value of R1 decreases the noise gain of the op amp. Second, capacitor C3 must be large enough that its impedance is much less than resistor R1 at audio frequencies. Increasing the value of R1 allows for smaller capacitances to be used for C3.

Resistor R1 and capacitor C3 form a high-pass filter. The corner frequency of this filter must be low enough to not attenuate low-frequency sound waves. A 5Hz corner frequency is used to calculate the value of C3:

$$C_3 = \frac{1}{2\pi R_1 f_C} = \frac{1}{2\pi(13.7k\Omega)(5Hz)} = 2.32\mu F \rightarrow 2.2\mu F \qquad (11)$$

## 2.3 Op Amp Bias Network

Resistors R3 and R5 center the op amp input and output at the midpoint between the power supplies to allow for the widest possible output signal swing. Therefore $R_3 = R_5$ for $V_B = V_{CC} / 2$. 100kΩ resistors were used in order to limit the power supply current drawn by this voltage divider. The current in the voltage divider will be:

$$i_D = \frac{V_{CC}}{R_3 + R_5} = \frac{9V}{200k\Omega} = 45\mu A \qquad (12)$$

Larger resistors may be used if this current needs to be limited further. Capacitor C6 is included to filter thermal noise created by the resistors and any noise which may be present on the power supply. The corner frequency of the low pass filter formed by R3, R5, and C6 is:

$$f_C = \frac{1}{2\pi(R_3 || R_5)C_6} \qquad (13)$$

This corner frequency should be well below the audible range (<20Hz) in order to prevent noise from affecting the audio performance of the design. For simplicity, a 2.2uF capacitor was selected for C6 as this value was used elsewhere in the design. The resulting corner frequency is:

$$f_C = \frac{1}{2\pi(100k\Omega || 100k\Omega)2.2\mu F} = 1.447Hz \qquad (14)$$

## 2.4 Ac Coupling Network

Audio systems are commonly ac-coupled so that only the audio signal is passed between them, and any dc voltages are removed. Ac-coupling is accomplished by capacitor C5, but resistors R4 and R6 should also be included. Resistor R6 provides a discharge pathway to prevent the build-up of charge on capacitor C5. If charge was allowed to build-up on the capacitor, it could potentially be rapidly discharged when connecting equipment to the output of the pre-amplifier. This rapid discharge can cause audible "thumps" when connecting equipment, and can also damage sensitive circuitry. A 100kΩ resistor is a typically used for this discharge pathway.

Resistor R4 limits the current through the capacitor in case a circuit at a different dc voltage is connected to the pre-amplifier output. This resistor may be 10's of ohms to several hundred ohms. A value of 49.9Ω is selected to also allow the circuit to also be properly interfaced to 50Ω test equipment.

The ac coupling capacitor, C5, forms a high-pass filter with resistor R6. However, R6 will be in parallel with the input impedance of any circuitry attached to the microphone pre-amplifier. For this design we will assume that the combined impedance of R6 and any attached circuitry will be 10kΩ:

$$R_6 || R_{IN(EXT)} = 10k\Omega \tag{15}$$

The corner frequency should be selected to avoid excessive attenuation at low frequencies. -0.5dB is considered acceptable in this design. Equation (7) can be modified in order to determine the corner frequency of a high-pass filter:

$$f_C = f\sqrt{\left(\frac{G_0}{G_f}\right)^2 - 1} = 20\sqrt{\left(\frac{1}{0.944}\right)^2 - 1} = 6.986 Hz \tag{16}$$

C5 can then be calculated to produce the desired corner frequency:

$$C_5 = \frac{1}{2\pi f_C (R_6 || R_{IN(EXT)})} = \frac{1}{2\pi(6.986Hz)(10k\Omega)} = 2.278\mu F \rightarrow 2.2\mu F \tag{17}$$

# 3 Component Selection

## 3.1 Passive Components

1% Thick film resistors in 0603 surface mount packages were used for this design. It should be noted that no additional distortion was observed due to these resistors. Distortion due to resistor non-linearity is typically seen at higher signal levels. Capacitor C2 must be a NP0/C0G type ceramic capacitor or film capacitor. High-K ceramic capacitors (X7R, Y5V, etc.) will produce distortion if installed at C2. Capacitors C3 and C5 are tantalum capacitors selected to avoid microphonic behavior associated with high-k ceramics[3]. Because these capacitors are polarized, correct polarity must be observed when they are installed in the circuit. Furthermore, there is minimal ac voltage drop across these capacitors and therefore they do not contribute distortion to the signal path.

## 3.2 Amplifier

Some basic amplifier selection criteria are given in Table 3. The maximum supply voltage, and power supply current are determined by the design requirements. A SOT23-5 package is preferred in this application to keep the overall solution size small.

**Table 3: Basic op amp selection criteria**

| Criteria | Value |
|---|---|
| Max. Power Supply Voltage | >9V |
| Power Supply Current | < 2.5mA |
| Package | SOT23-5 |
| Amplifiers in Package | 1 |

The required slew rate of the amplifier can be determined by calculating the maximum rate of change for a sine wave at 20kHz and line level audio voltages (1.228Vrms→1.736Vp):

$$R_S = 2\pi f A = 2\pi (20kHz)(1.736Vp) = 0.2182 V/\mu s \tag{18}$$

As a conservative rule, an op amp should be selected with 10 times this slew rate to eliminate any possibility of slew-induced distortion, making the required slew rate: 2.182V/µs.

The op amp selected should not degrade the signal to noise ratio of the microphone itself. To meet this criterion, the output noise spectral densities of the microphone and the amplifier circuit will be compared. Only broadband noise of the op amp and microphone is considered in this analysis. The signal to noise ratio in Table 2 is used to determine the current noise of the microphone. Several assumptions are made about the measurement conditions of this value. It is assumed to be an A-weighted value, measured at 1 Pa / 94 dB SPL, with a 2.2 kOhm resistor in the JFET drain. In this measurement condition, the signal current of the microphone will be 8.083uA$_{RMS}$, and the noise current will be:

$$68dB = 20 * \log\left(\frac{I_S}{I_N}\right) \rightarrow I_N = \frac{8.083\mu A}{10^{\frac{68dB}{20}}} = 3.218 nA_{RMS} \tag{19}$$

This RMS noise current can be converted to a current noise spectral density by dividing by the square root of bandwidth of integration. An A-weighting curve can be approximated using a 13.5 kHz noise bandwidth [4]:

$$\frac{3.218 nA_{RMS}}{\sqrt{13.5kHz}} = 27.7 pA/\sqrt{Hz} \tag{20}$$

This current noise spectral density includes contributions from both the microphone and the 2.2 kΩ resistor used for the measurement. The current noise contribution of the resistor is:

$$i_{nR} = \sqrt{\frac{4K_B T}{R}} = \sqrt{\frac{4(1.381 \times 10^{-23})(298)}{2.2k\Omega}} = 2.735 pA/\sqrt{Hz} \tag{21}$$

Where T is the absolute temperature in degrees Kelvin and $K_B$ is Boltzmann's constant. Extracting this noise from the total noise value gives the current noise spectral density of the microphone:

$$i_{nMIC} = \sqrt{i_{nT}^2 - i_{nR}^2} = \sqrt{\left(\frac{27.7pA}{\sqrt{Hz}}\right)^2 - \left(\frac{2.735pA}{\sqrt{Hz}}\right)^2} = 27.56pA/\sqrt{Hz} \tag{22}$$

The voltage noise spectral density from the microphone at the amplifier output is:

$$E_{NO(MIC)} = i_{nMIC}R_2 = \frac{27.56pA}{\sqrt{Hz}} * 75k\Omega = 2.067\mu V/\sqrt{Hz} \tag{23}$$

To ensure that the signal-to-noise ratio of the microphone is not significantly degraded, we will specify that the output voltage noise of the amplifier should be less than 1/10th of the microphone:

$$E_{NO(AMP)} < \frac{E_{NO(MIC)}}{10} \tag{24}$$

The output noise spectral density of the amplifier circuit is:

$$E_{NO(AMP)} = A_N\sqrt{E_{NI}^2 + E_{NV}^2 + E_{NR}^2} \tag{25}$$

Where $A_N$ is the noise gain, $E_{NI}$ is the op amp current noise contribution, $E_{NV}$ is the op amp voltage noise contribution, and $E_{NR}$ is the thermal noise contribution from the feedback network. Substituting the microphone noise into the equation, the terms can be rearranged to give some selection criteria for the op amp:

$$\left(\frac{E_{NO(MIC)}}{10A_N}\right)^2 - E_{NR}^2 > E_{NI}^2 + E_{NV}^2 \tag{26}$$

$$\left(\frac{E_{NO(MIC)}}{10A_N}\right)^2 - E_{NR}^2 > (I_N * R_2||R_1)^2 + E_{NV}^2 \tag{27}$$

The terms to the left of the inequality are already set by the microphone selection and circuit design. The terms to the right of the inequality, op amp input current noise ($I_N$) and input voltage noise ($E_{NV}$), are determined by the op amp selection. Solving for the terms to the left of the inequality first will give the op amp requirements for noise:

$$A_N = 1 + \frac{R_2}{R_1} = 1 + \frac{75k\Omega}{13.7k\Omega} = 6.474 \tag{28}$$

$$E_{NR} = \sqrt{4K_BT(R_2||R_1)} = \sqrt{4(1.381 \times 10^{-23})(298)(11584)} \\ = 13.8nV/\sqrt{Hz} \tag{29}$$

$$\left(\frac{2.067\mu V/\sqrt{Hz}}{10 * 6.474}\right)^2 - (\frac{13.8nV}{\sqrt{Hz}})^2 > (I_N * 11584)^2 + E_{NV}^2 \\ 8.2928 \times 10^{-16} > (I_N * 11584)^2 + E_{NV}^2 \tag{30}$$

Due to the high source impedance at the inverting input (11584Ω), the op amp selection will be narrowed to devices with JFET or CMOS input devices which feature extremely low current noise. For this reason the input current noise term can be ignored, allowing a limit of input voltage noise spectral density to be calculated:

$$8.2928 \times 10^{-16} > E_{NV}^2 \\ \frac{28.8nV}{\sqrt{Hz}} > E_{NV} \tag{31}$$

The OPA172 was selected for this design because of its small package size, low power consumption, and excellent noise and slew rate as shown in Table 4.

**Table 4: Comparison of the OPA172's specifications to those required by the design.**

| Criteria | Required | OPA172 |
|---|---|---|
| Max. Power Supply Voltage | >9V | 36V |
| Power Supply Current | < 2.5mA | 1.5mA |
| Package | SOT23-5 | SOT23-5 |
| Amplifiers in Package | 1 | 1 |
| Slew Rate | 2.2V/μs | 10V/μs |
| Input Voltage Noise | 28.8nV/√Hz | 7nV/√Hz |
| Input Current Noise | N/A | 1.6fA/√Hz |

## 4    Simulation

A basic model of an electret microphone was used for simulation. The model uses a voltage controlled current source (VCCS) which mimics the function of the internal JFET of the microphone. A voltage generator (VS) is used to represent sound pressure level in Pascals (1V = 1Pa), and a dc voltage source (VE) produces a bias current through the VCCS. A current noise source, In, is used to simulate noise of the microphone capsule itself.
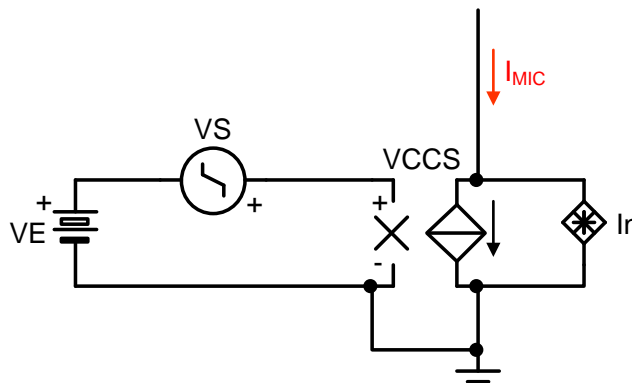


**Figure 6: TINA-TI™ schematic of an electret microphone simulation model.**

The transfer function of the VCCS is taken directly from equation 3, replacing Pascals with volts: 8.083μA/V. DC voltage source VE is then calculated to establish a bias current using the "current consumption" value given in Table 2.

$$I_{dc} = 0.5mA = V_E * \frac{8.083\mu A}{V} = 61.858V \tag{32}$$

A generic current noise source is available for download at: http://www.ti.com/tool/tina-ti. The net list of the current noise source will need to be modified to include the broadband current noise spectral density previously calculated. Double clicking on the icon of the current noise source brings up the parameters dialog. Selecting "Enter Macro" allows the user to modify the net list of the current noise source.
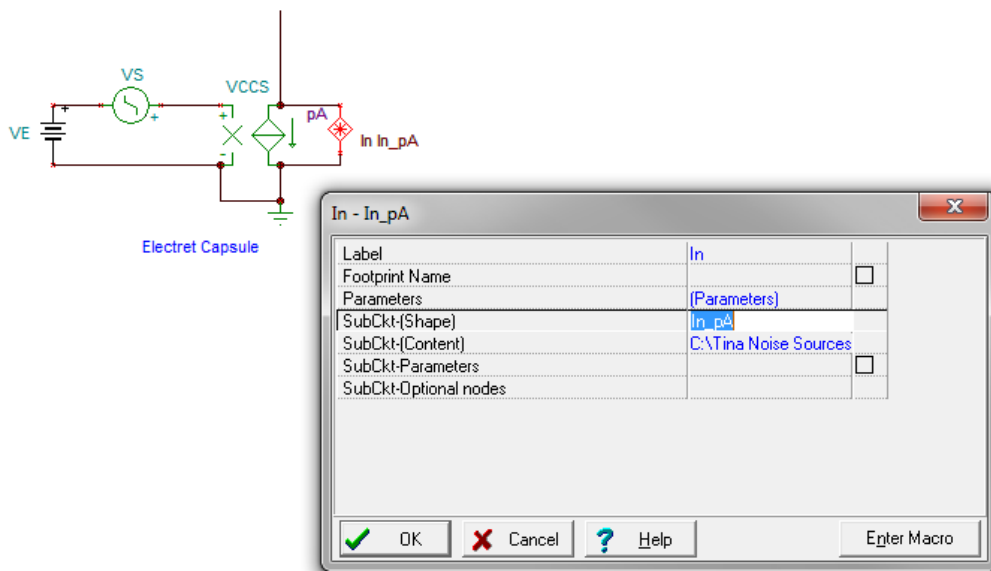


**Figure 7: Double-clicking on the noise source icon opens the parameters dialog. The "Enter Macro" button allows the user to modify the model netlist.**

Within the net list, the NLFP and NVRP values are the 1/f and broadband noise values in pA/√Hz respectively. Both of these are changed to 27.7 to model the noise of the microphone capsule as shown in Figure 8.

```
* BEGIN PROG NSE PICO AMP/RT-HZ
.SUBCKT PICO 1 2
* BEGIN SETUP OF NOISE GEN - PICOOAMPS/RT-HZ
* INPUT THREE VARIABLES
* SET UP INSE 1/F
* PA/RHZ AT 1/F FREQ
.PARAM NLFP=26.49          <--
* FREQ FOR 1/F VAL
.PARAM FLWP=1
* SET UP INSE FB
* PA/RHZ FLATBAND
.PARAM NVRP=26.49          <--
* END USER INPUT
* START CALC VALS
.PARAM GLFP={PWR(FLWP,0.25)*NLFP/1164}
.PARAM RNVP={1.184*PWR(NVRP,2)}
.MODEL DVNP D KF={PWR(FLWP,0.5)/1E11} IS=1.0E-16
* END CALC VALS
I1 0 7 10E-3
I2 0 8 10E-3
D1 7 0 DVNP
D2 8 0 DVNP
E1 3 6 7 8 {GLFP}
R1 3 0 1E9
R2 3 0 1E9
R3 3 6 1E9
E2 6 4 5 0 10
R4 5 0 {RNVP}
R5 5 0 {RNVP}
R6 3 4 1E9
R7 4 0 1E9
G1 1 2 3 4 1E-3
C1 1 0 1E-15
C2 2 0 1E-15
C3 1 2 1E-15
.ENDS
* END PROG NSE PICO AMP/RT-HZ
```

**Figure 8: Current noise model netlist. The modified parameters are highlighted with red arrows.**

## 4.1   Transfer Function

Voltage source VG2 was configured as a 1Vrms, 1kHz sinusoid in order to simulate an input signal of 94dB SPL to the microphone. The ac nodal voltages were calculated and the result of the simulation is displayed in Figure 9. For a 94dB SPL input signal, the simulated output voltage is 605.8mVrms.
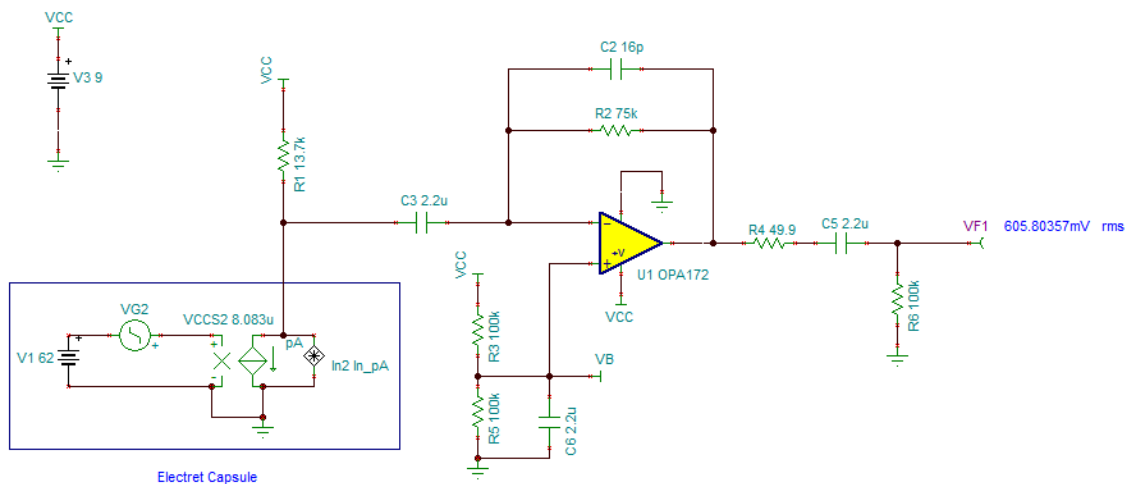


**Figure 9: An ac nodal voltage simulation for a simulated 94dB SPL input signal.**

An ac transfer characteristic simulation was also performed in TINA-TI™ and the results are displayed in Figure 10. It is important to note that the gain measurement is taken from the input voltage source (VG2) which produces a somewhat misleading gain magnitude value. The transfer ratio of the VCCS is 8.083µA/V, or effectively -101.849dB. The gain of the pre-amp circuit is 75kV/A or 97.5dB. This gives a an expected gain with respect to VG2 of 97.5dB – 101.849dB = -4.347dB.
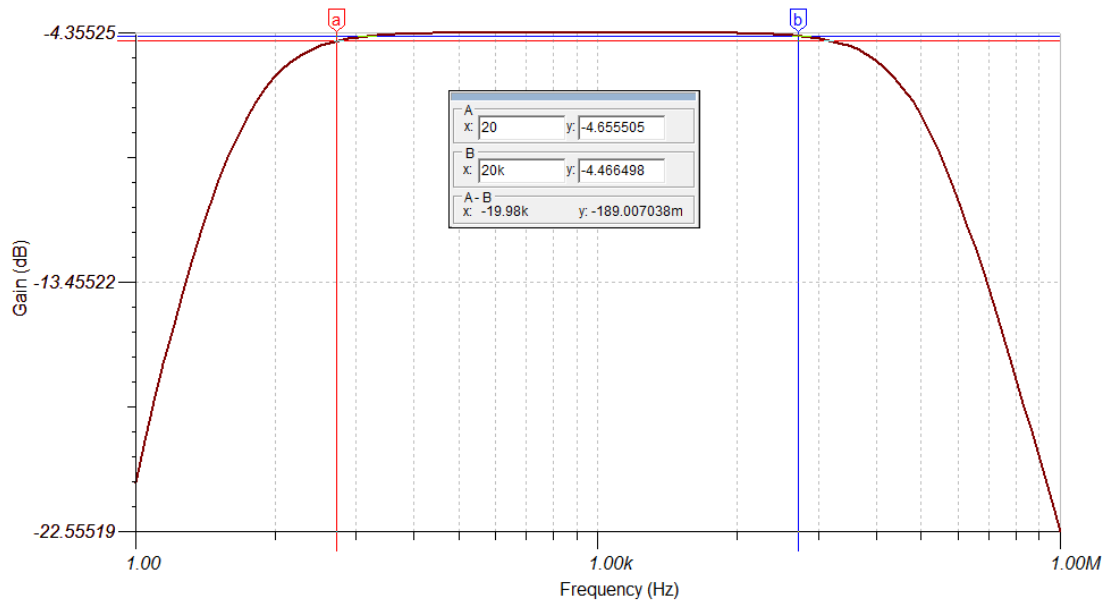


**Figure 10: An ac transfer characteristic simulation with cursor measurements at 20Hz and 20kHz.**

At 20Hz the gain has deviated by -0.3dB and at 20kHz the deviation is -0.1dB. This does not include the transfer function of the microphone itself, which is not modeled.

## 4.2    Noise

The total noise of the microphone and pre-amplifier circuit was simulated in a bandwidth of 13.5kHz in order to mimic an A weighting curve. The noise in this bandwidth was 238.78 µVrms.
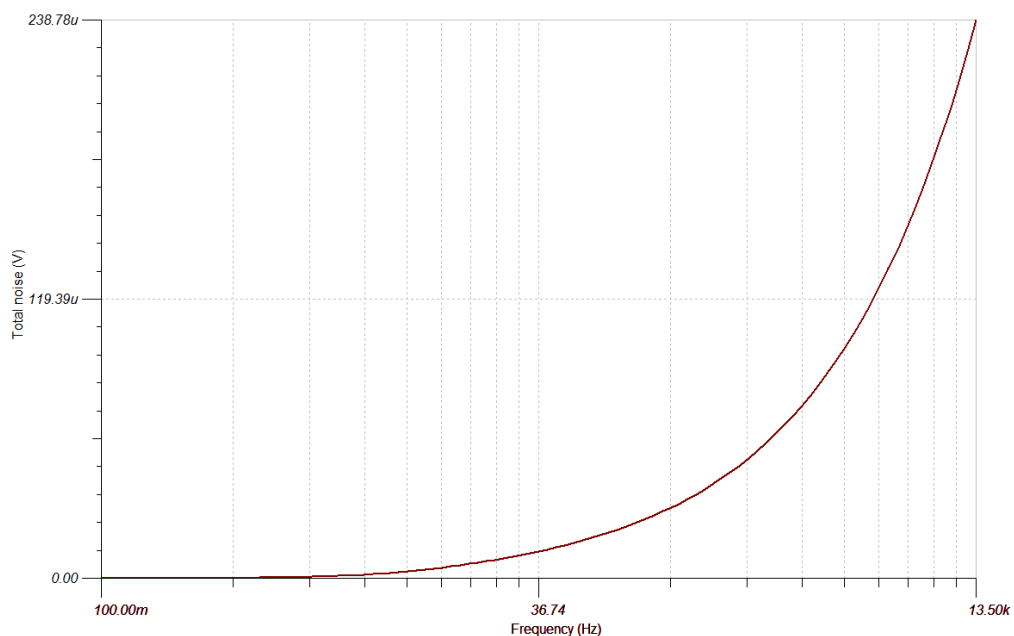


**Figure 11: A total noise integration over 13.5kHz performed in TINA-TI™**

This noise voltage can be used to calculate the expected signal-to-noise ratio (SNR) for a given sound pressure level. At 94 dB SPL, the output voltage of the amplifier will be 0.606Vrms, producing an SNR of:

$$\text{SNR} = 20\log\left(\frac{V_s}{V_n}\right) = 20\log\left(\frac{0.606\text{Vrms}}{238.78\mu\text{Vrms}}\right) = 68.09\text{dB} \tag{33}$$

This SNR is essentially identical to the microphone capsule itself, which shows that the pre-amplifier contributes minimal noise to the signal chain. The calculated value is slightly higher than the intrinsic SNR of the microphone due to slight inaccuracies in the noise model.

## 4.3 Stability Analysis

A stability analysis was performed to ensure that the circuit would be stable with long cables attached to the output. A 1nF capacitor was added to the output of the circuit for simulation. The feedback loop was broken by inductor LT, and a test signal is injected by voltage source VG1 through capacitor CT.
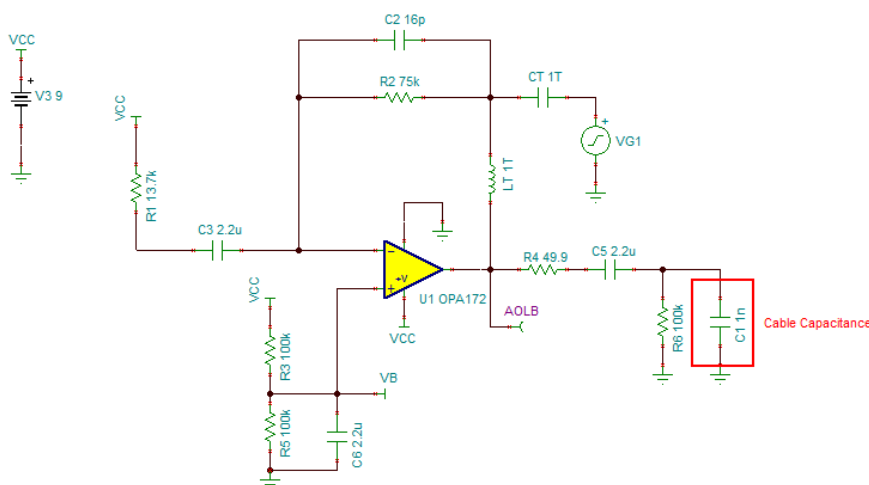


**Figure 12: TINA-TI(TM) schematic for simulating loop gain and phase.**

The loop gain is measured at the voltage probe AOLB, and phase margin is determined by measuring the phase at the frequency where the loop gain is 0 dB.
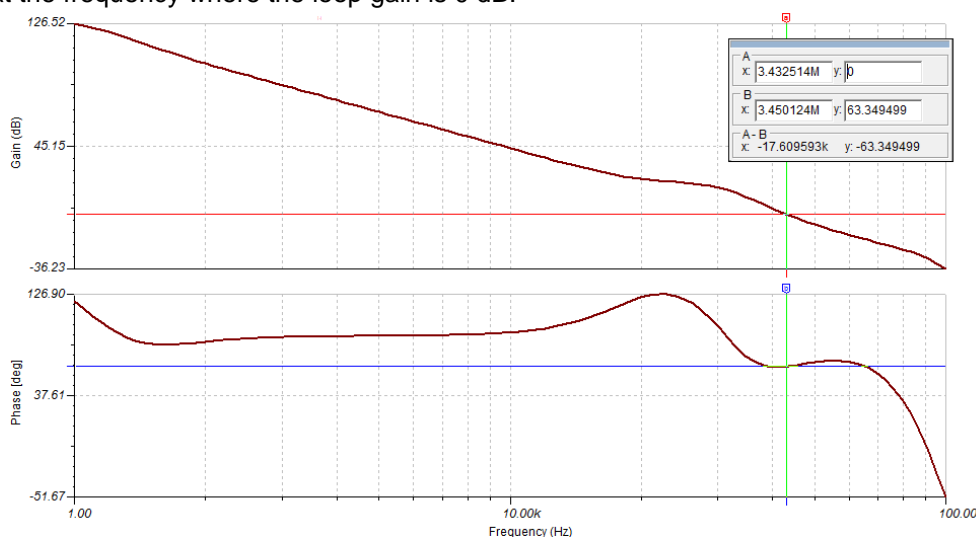


**Figure 13: A phase margin measurement taken where loop gain is 0dB.**

With 1nF of cable capacitance added to the output, the phase margin is 63.35 degrees, suggesting the design will be stable with cables connected to the output.

# 5    PCB Design

The PCB schematic and bill of materials can be found in the Appendix.

## 5.1    PCB Layout

The top and bottom layers of the PCB are shown in Figure 14. The main consideration in the PCB layout and assembly is to observe the proper polarities of the tantalum capacitors. C3 is installed with its positive terminal towards the op amp because the op amp input is biased at 4.5V, while the microphone is biased at 2V. The positive terminal of C5 also faces the op amp output because of the 4.5V dc offset on the output signal. The microphone is installed at the edge of the PCB facing away from the circuit. Two additional ground pads are given next to the microphone capsule to allow the grounding of the microphone case or other enclosures.
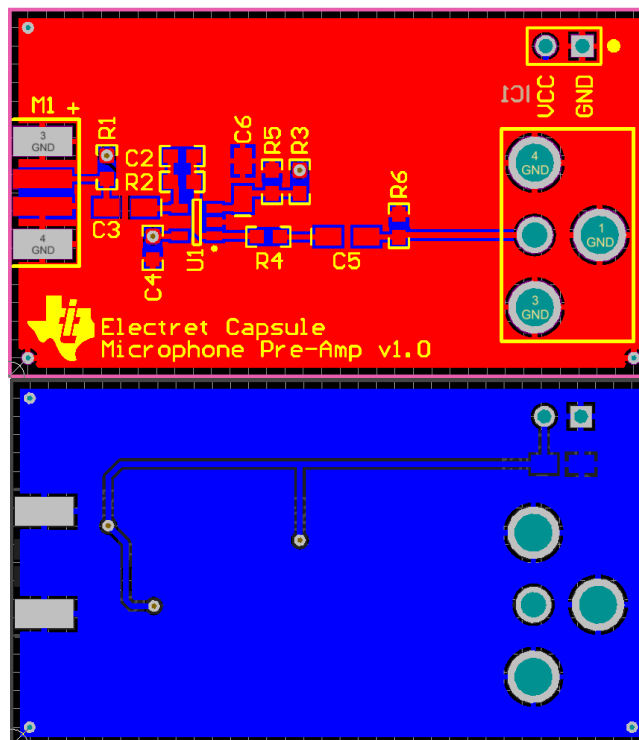


**Figure 14: The top layer (upper, red) and bottom layer (lower, blue) of the preamplifier PCB.**

# 6 Verification & Measured Performance

The test setup for measuring circuit performance is shown in Figure 15. Testing the total system performance (microphone and pre-amplifier) would require specialized acoustic facilities. An easier alternative is to remove the microphone and test only the pre-amplifier circuit. A large resistor was placed in series with the output of an audio analyzer to produce an output current similar to a microphone. The resistor was sized so that 1Vrms generator amplitude was equivalent to the 94dB SPL signal level of the microphone:

$$R = \frac{1Vrms}{8.083\mu A} = 123.716k\Omega \rightarrow 124k\Omega \tag{34}$$

The actual measured value of the resistor selected was 124.71kΩ and the signal generator has an output impedance of 20Ω. Therefore the generator voltage required to produce the proper input current level is:

$$V_{GEN} = 8.083\mu A * (124.71k\Omega + 20\Omega) = 1.008Vrms \tag{35}$$
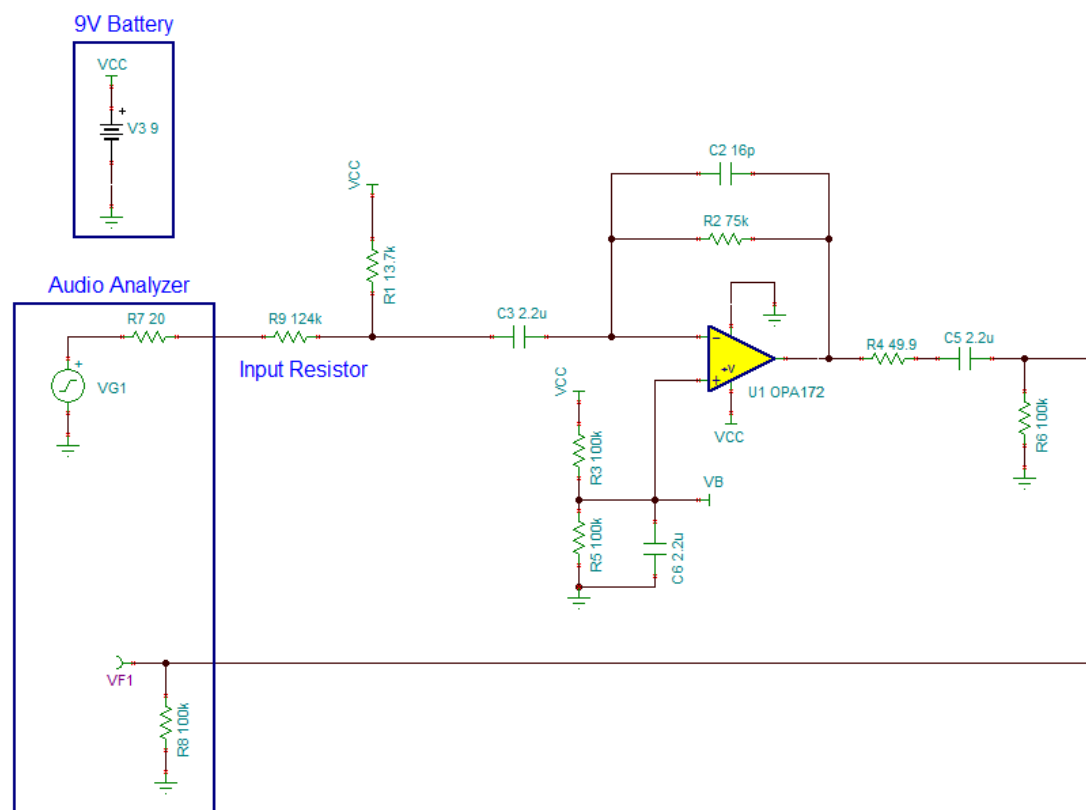


**Figure 15: The test setup used to measure the microphone pre-amplifier circuit.**

The output of the preamplifier circuit was then fed back to the input of the audio analyzer. The 100kΩ input impedance of the analyzer will only slightly affect the low frequency response of the circuit. Finally, a 9V battery was used to power the preamplifier circuit for all measurements presented here.

## 6.1 Transfer Function

The pre-amplifier transfer function is shown in Figure 16. Using the gain at 1kHz as the nominal value, the response deviates by -0.423dB at 20Hz and -0.117dB at 20kHz. The output voltage for a 8.083µA input signal (representing a 49dB SPL signal into the microphone) is 0.6065Vrms.
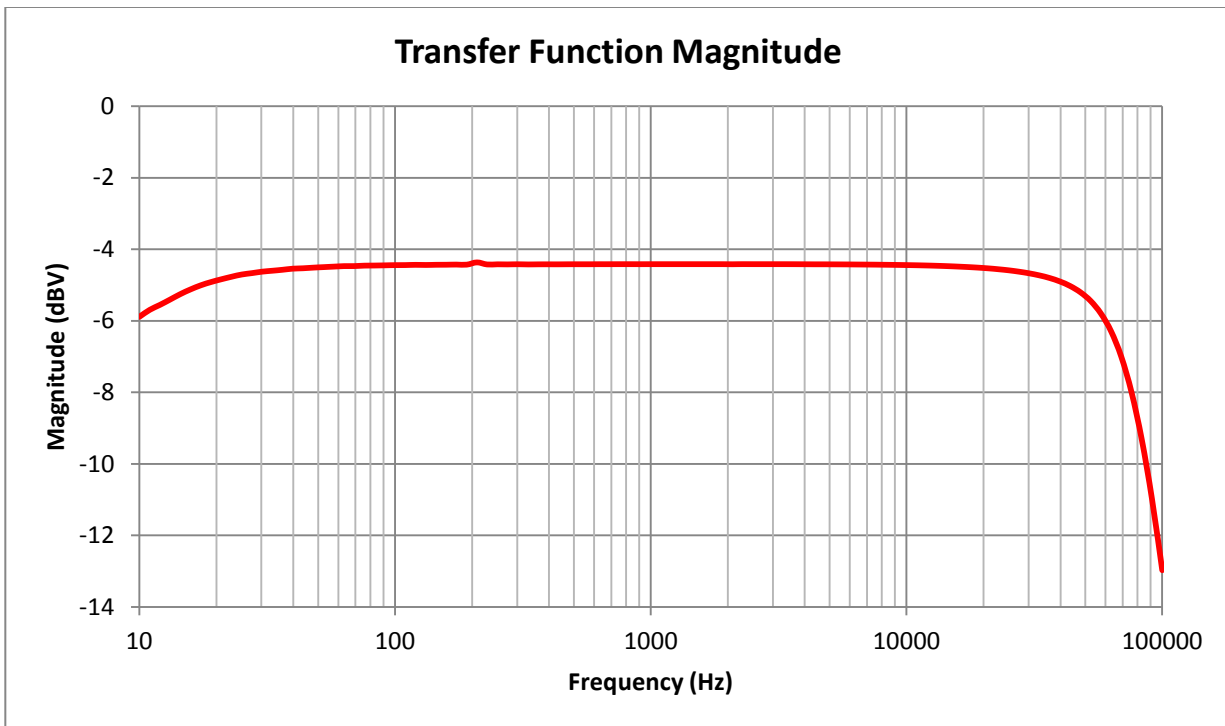
Figure 16: Measured pre-amplifier transfer function

## 6.2 Total Harmonic Distortion and Noise

An FFT was taken of a 600mVrms, 1kHz, output from the pre-amplifier to demonstrate the noise and distortion performance of the circuit without a microphone (Figure 17).
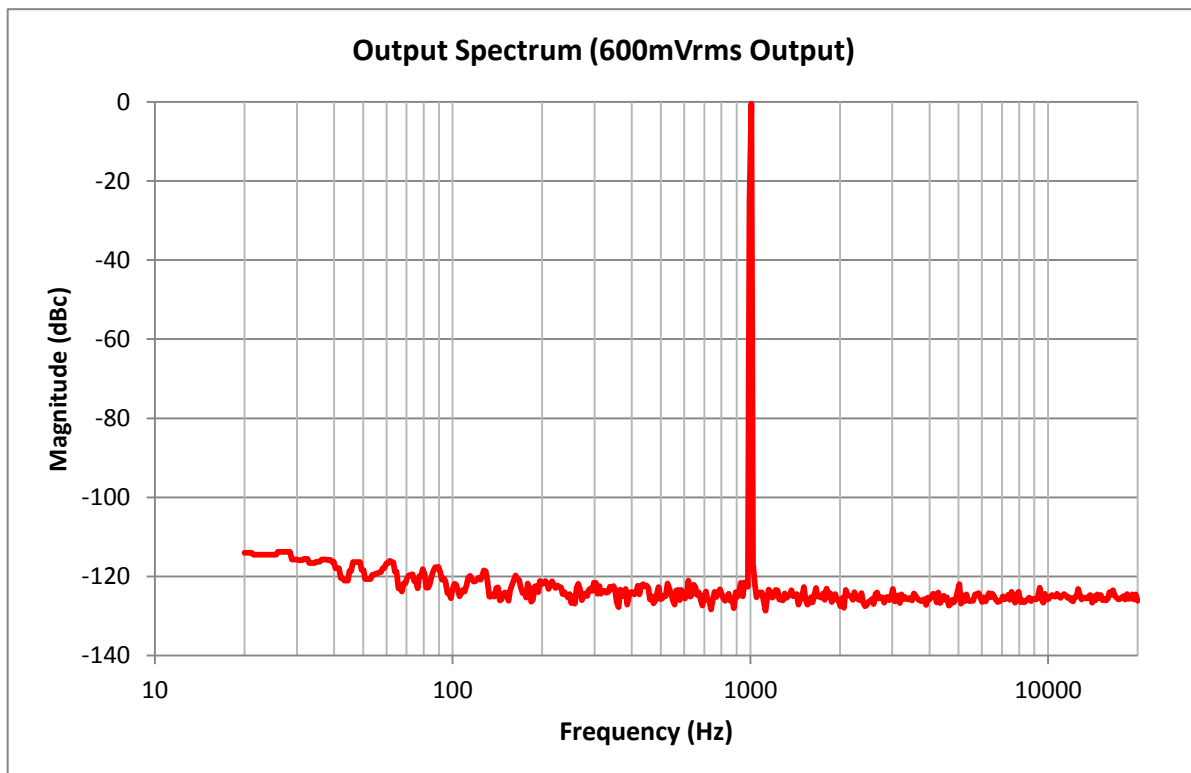


Figure 17: Output FFT for a 600mVrms, 1kHz, output signal

The A-weighted THD+N measurement was -93.5dB for a 600mVrms output signal. As the FFT in Figure 17 shows, there are no harmonics above the noise floor of the output spectrum and therefore the THD+N measurement is noise dominated. This measurement can then be used to determine the output noise of the pre-amplifier circuit without the microphone:

$$10^{\frac{THD+N(dB)}{20}} = \sqrt{\frac{V_n^2}{V_f^2}} \rightarrow \sqrt{\frac{\left(10^{\frac{THD+N(dB)}{20}}\right)^2}{V_f^2}} = V_n = 12.68 \mu Vrms \tag{36}$$

As stated in section 3.2, the output noise of the pre-amplifier circuit should be less than 1/10th the noise of the microphone to avoid degrading the overall SNR:

$$E_{NO(AMP)} < \frac{E_{NO(MIC)}}{10} \tag{37}$$

The A-weighted output RMS noise voltage from the microphone will be:

$$E_{NO(MIC)} = \frac{2.067 \mu V}{\sqrt{Hz}} \sqrt{13.5kHz} = 240.16 \mu Vrms \tag{38}$$

The total SNR of the system is:

$$SNR = 20 \times \log\left(\frac{600mVrms}{\sqrt{(240.16\mu Vrms)^2 + (12.68\mu Vrms)^2}}\right) = 67.94dB \tag{39}$$

## 7    Modifications

This circuit may be modified for lower power supply voltages by changing R1 to maintain a 2V bias at the microphone. Figure 18 illustrates a version of the circuit modified for a 5V power supply. Because R1 must be decreased to maintain the 2V bias of the microphone, the noise gain of the op amp is increased. In order to preserve the SNR of the microphone, the input voltage noise requirement of the op amp is now less than 10.9nV/√Hz. The OPA322 and LMV796 are excellent choices for the lower voltage version of this circuit.
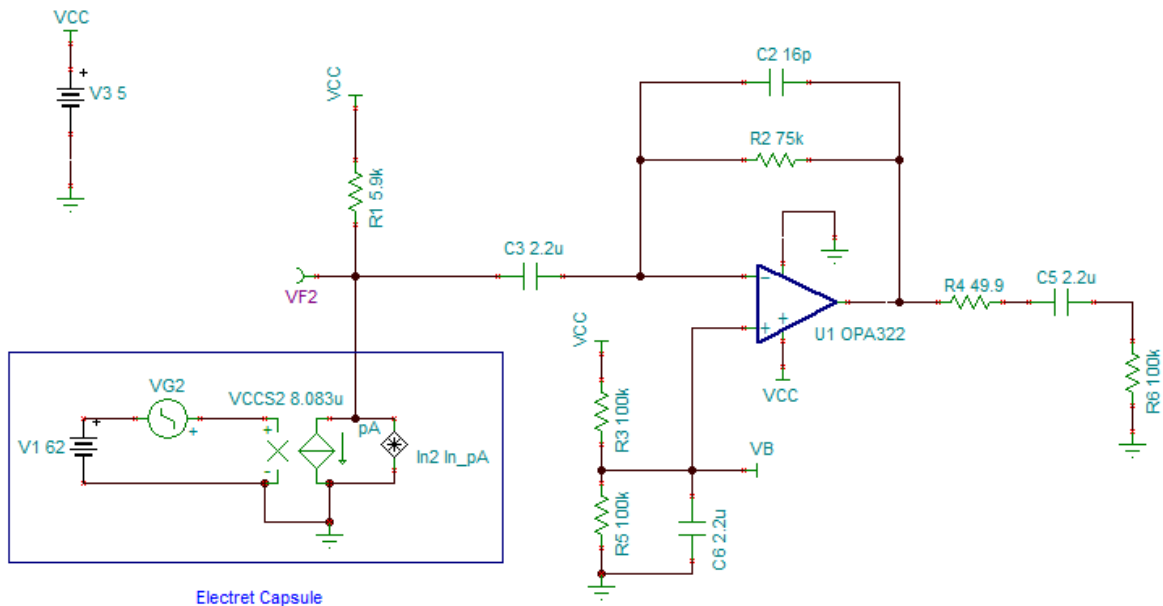


**Figure 18: A 5V version of the microphone pre-amplifier circuit.**

## 8    About the Author

John Caldwell is an applications engineer with Texas Instruments Precision Analog, supporting operational amplifiers and industrial linear devices. He specializes in precision circuit design for sensors, low-noise design and measurement, and electromagnetic interference issues.  He received his MSEE and BSEE from Virginia Tech with a research focus on biomedical electronics and instrumentation. Prior to joining TI in 2010, John worked at Danaher Motion and Ball Aerospace.

# 9 References

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2. POM-3535P-3 Datasheet, http://www.puiaudio.com/pdf/POM-3535P-3-R.pdf , PUI Audio, Dayton, OH.

3. Caldwell, J*., Stress-Induced Outbursts: Microphonics in Ceramic Capacitors (Part 2)*, http://e2e.ti.com/blogs_/b/precisionhub/archive/2014/12/23/stress-induced-outbursts-microphonics-in-ceramic-capacitors-part-2, Texas Instruments, 2014.

4. Wells, C*. Modeling Electret Microphones in TINA Spice*, Texas Instruments – Internal Presentation.

# Appendix A.

## A.1  Electrical Schematic
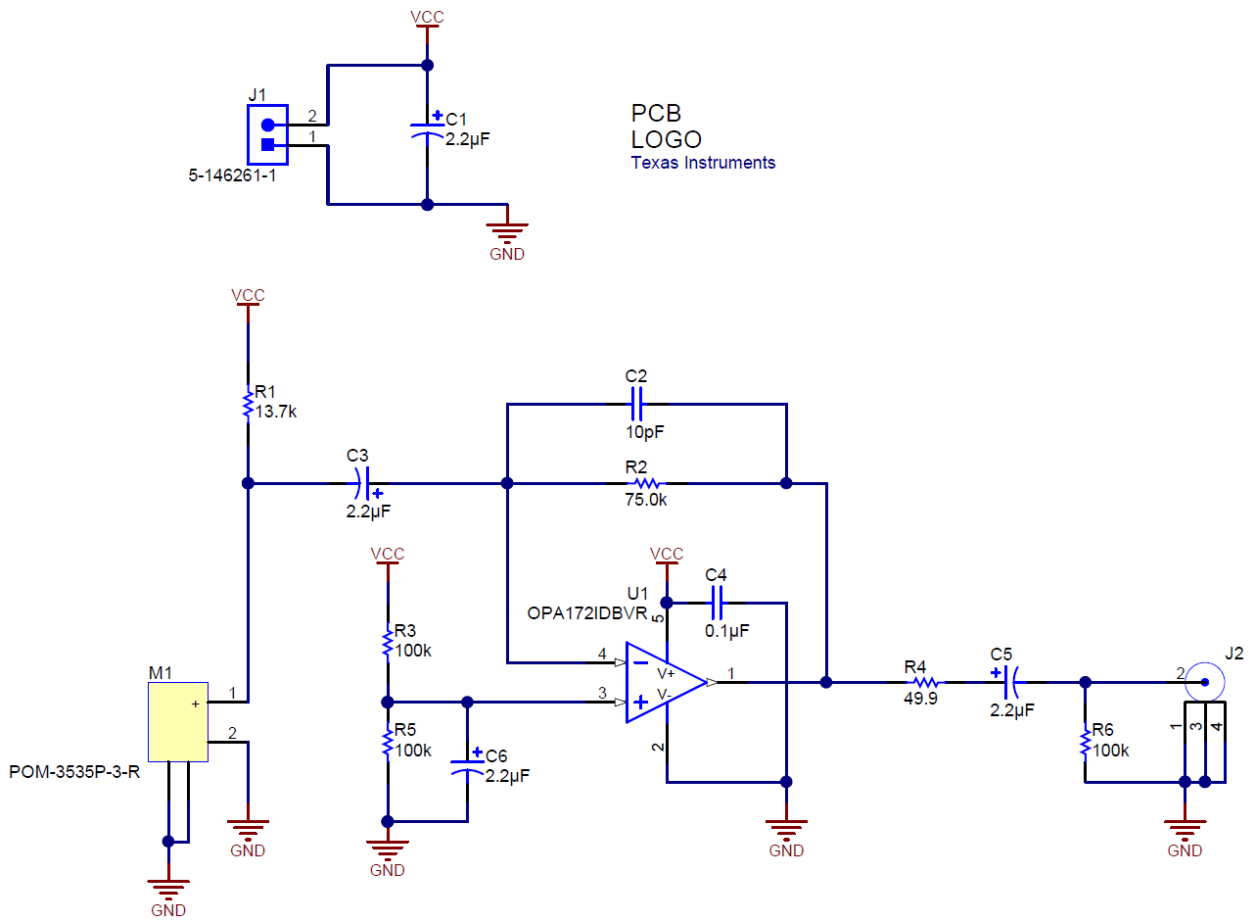


**Figure A-1: Electrical Schematic**

## A.2  Bill of Materials

| Item | Qty | Value | Designator | Description | Manufacturer | Part Number |
|------|-----|-------|------------|-------------|--------------|-------------|
| 1 | 4 | 2.2uF | C1, C3, C5, C6 | CAP, TA, 2.2uF, 16V, +/-10%, 5.9 ohm, SMD | Vishay-Sprague | 293D225X9016A2TE3 |
| 2 | 1 | 10pF | C2 | CAP, CERM, 10pF, 50V, +/-5%, C0G/NP0, 0603 | AVX | 06035A100JAT2A |
| 3 | 1 | 0.1uF | C4 | CAP, CERM, 0.1uF, 16V, +/-5%, X7R, 0603 | AVX | 0603YC104JAT2A |
| 4 | 1 | | J1 | Header, 100mil, 2x1, Gold, TH | TE Connectivity | 5-146261-1 |
| 5 | 1 | | J2 | RCA Jack, Metal, Right Angle | CUI Inc | RCJ-011 |
| 6 | 1 | | M1 | Electret capsule microphone | PUI Audio | POM-3535P-3-R |
| 7 | 1 | 13.7k | R1 | RES, 13.7k ohm, 1%, 0.1W, 0603 | Vishay-Dale | CRCW060313K7FKEA |
| 8 | 1 | 75.0k | R2 | RES, 75.0k ohm, 1%, 0.1W, 0603 | Vishay-Dale | CRCW060375K0FKEA |
| 9 | 3 | 100k | R3, R5, R6 | RES, 100k ohm, 1%, 0.1W, 0603 | Vishay-Dale | CRCW0603100KFKEA |
| 10 | 1 | 49.9 | R4 | RES, 49.9 ohm, 1%, 0.1W, 0603 | Vishay-Dale | CRCW060349R9FKEA |
| 11 | 1 | | U1 | 36V 10MHz, CMOS Operational Amplifier | Texas Instruments | OPA172IDBVR |

**Figure A-2: Bill of Materials**

# IMPORTANT NOTICE FOR TI REFERENCE DESIGNS