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TI Designs – Precision: Verified Design

Tone Stack for Guitar Amplifier Reference Design

TEXAS INSTRUMENTS

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Circuit Description

This split-supply, high-performance guitar tone circuit provides control of the bass, mid, and treble frequencies of an electric guitar signal, while also providing gain with minimal distortion and noise. Buffered inputs and outputs preserve the behavior of the system independent of the source and load impedances, and a radio frequency (RF) filter on the circuit front end attenuates noise from outside the audio band.

Design Resources

Design Archive	All Design files
TINA-TI™	SPICE Simulator
OPA1642	Product Folder

TI E2E™ Community

Ask the Analog Experts
WEBENCH® Design Center
TI Designs – Precision Library
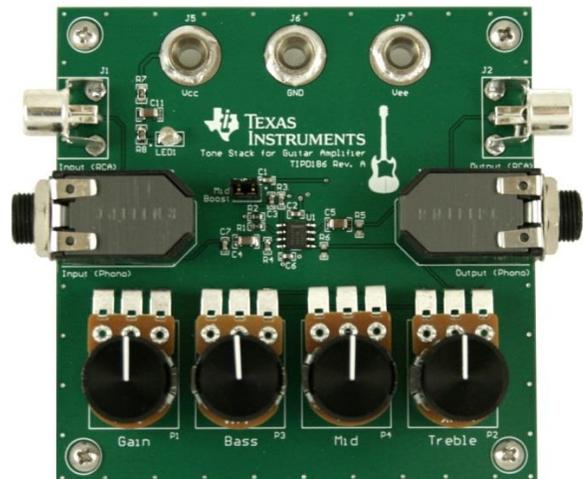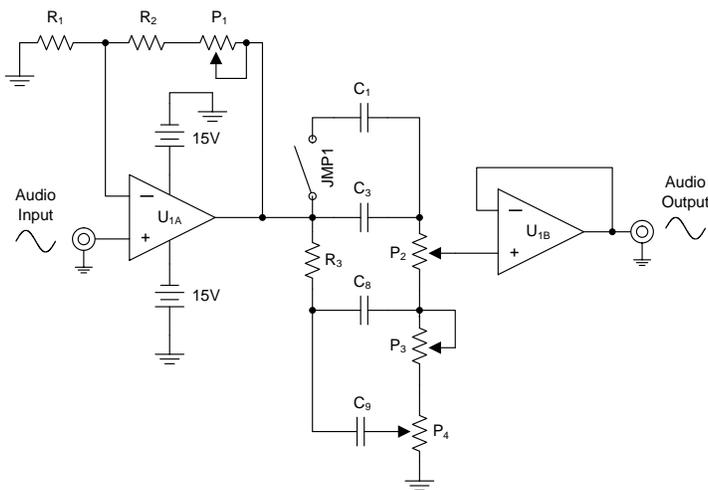
1   Design Summary

The design requirements are as follows:

- Supply voltage: ± 15 V

- Input voltage: 1 V$_{RMS}$

- Source impedance: 6 kΩ

- Input stage signal gain: 6 dB – 28 dB

- Total harmonic distortion + noise (THD+N) level at 1 kHz: -100 dB (0.001%)

- Treble adjustment range: 10 dB

- Mid adjustment range: 6 dB

- Bass adjustment range: 15 dB

The design goals and performance are summarized in Table 1. Figure 1 depicts the measured transfer function of the design.

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

	Goal	Simulated	Measured
THD+N level at 1 kHz	-100 dB (0.001%)	-102.6 dB (0.00074%)	-105.4 dB (0.00054%)
Treble adjustment range	10 dB	10.4 dB	10.6 dB
Mid adjustment range	6 dB	6.1 dB	8.9 dB
Bass adjustment range	15 dB	18 dB	19.2 dB
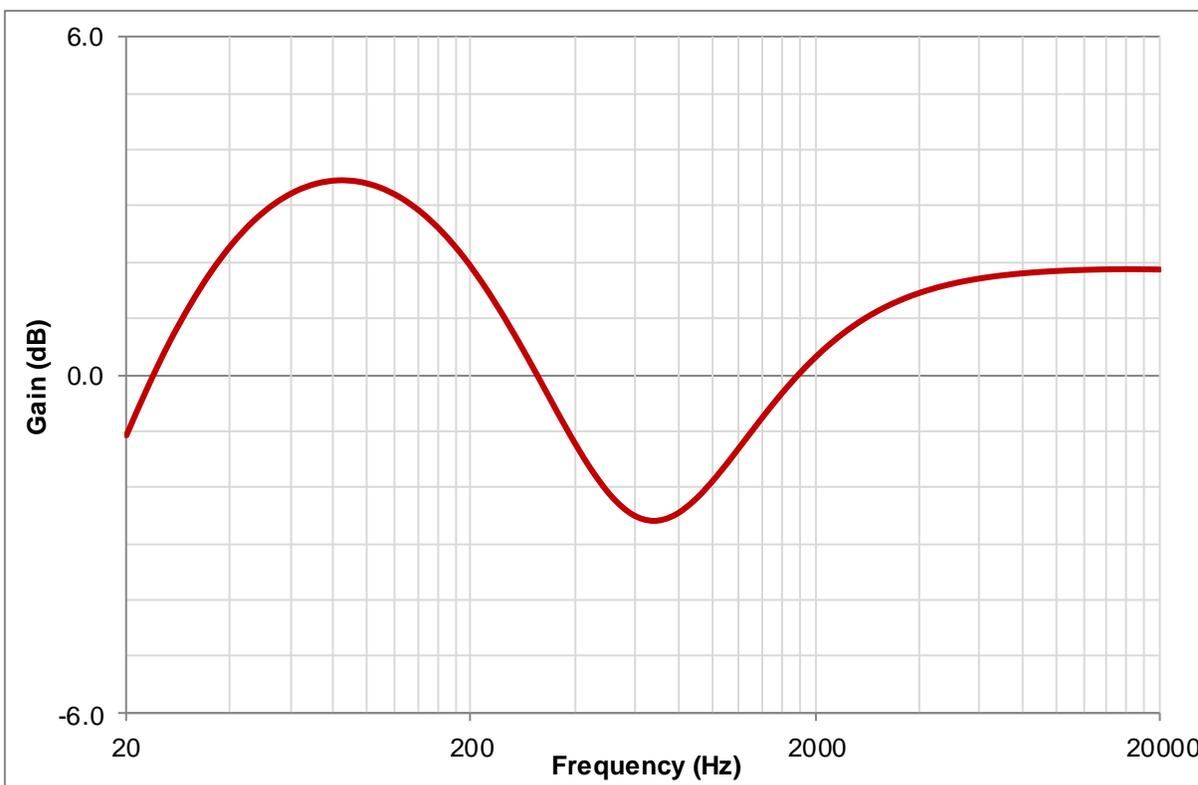
Figure 1: Measured Transfer Function – Bass, Mid, Treble at 50%

2 Theory of Operation

A more complete schematic for this design is shown in Figure 2. The three primary functional blocks of the circuit are the input filter and gain stage, tone stack, and output buffer.
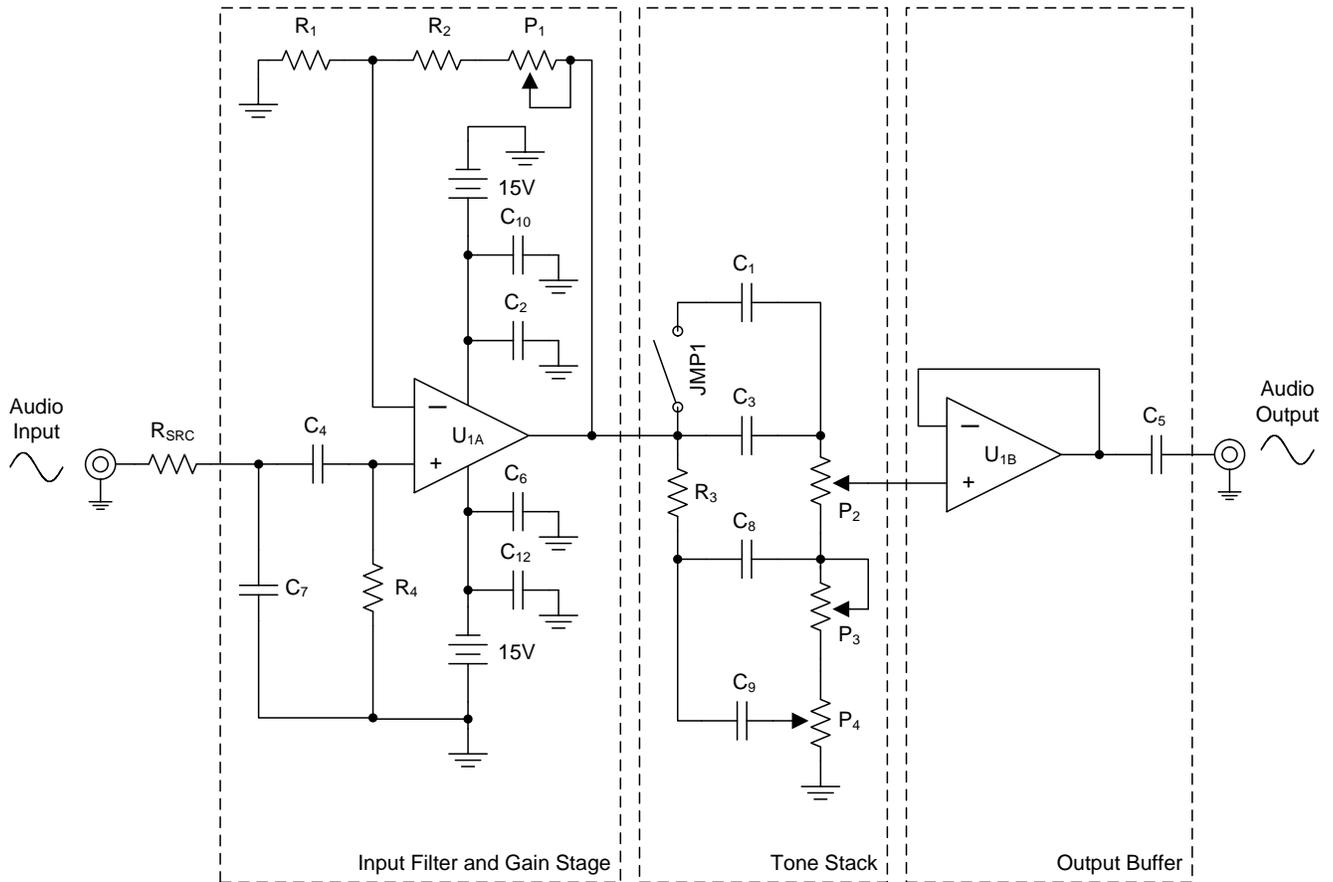
Figure 2: Complete Circuit Schematic

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2.1 Input Filter

A passive filter at the input of the circuit serves two purposes: provide significant attenuation at frequencies outside the audio band, and remove any dc voltage from the input signal. The filter is made up of R_{SRC}, the guitar pickup output impedance, capacitors C_7 and C_4, and resistor R_4, as shown in Figure 3.
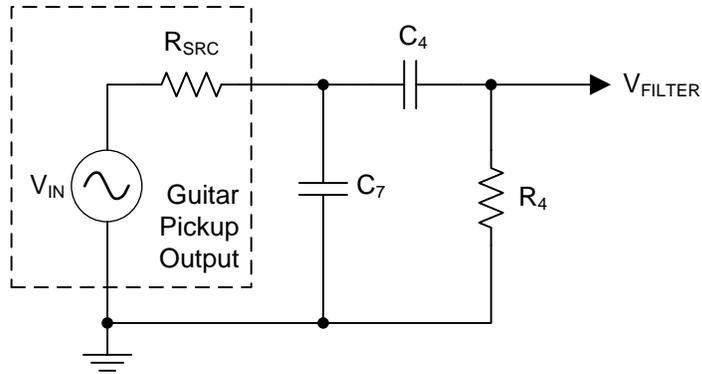
Figure 3: Input Filter Schematic

2.1.1 Low Pass Filter

R_{SRC} and C_7 create a first-order low pass filter. The -3 dB cutoff frequency of the filter is calculated using Equation 1.

$$f_{C\_LPF} = \frac{1}{2\pi * R_{SRC} * C_7}$$

(1)

400 kHz is selected as the -3 dB cutoff frequency for the filter. This will effectively attenuate RF noise while preserving the gain and phase behavior at 20 kHz. Since R_{SRC} is specified at 6 kΩ, simply rearrange terms and solve for C_7 in order to achieve the desired cutoff frequency, as shown in Equation 2.

$$C_7 = \frac{1}{2\pi * R_{SRC} * f_{C\_LPF}} = \frac{1}{2\pi * 6k\Omega * 400kHz} = 66.3 \text{ pF}$$

(2)

The required value for C_7 is calculated to be 66.3 pF. The nearest standard capacitor value of 68 pF is selected as the actual value. The actual cutoff frequency of the filter is calculated using Equation 3.

$$f_{C\_LPF} = \frac{1}{2\pi * R_{SRC} * C_7} = \frac{1}{2\pi * 6k\Omega * 68pF} = 390 \text{ kHz}$$

(3)

2.1.2 High Pass Filter

C_4 and R_4 create a first-order high pass filter. The -3 dB cutoff frequency of the filter is calculated using Equation 4.

$$f_{C\_HPF} = \frac{1}{2\pi * R_4 * C_4} \tag{4}$$

A value of 10 µF is selected for C_4, as it is a common value already used in the circuit for power supply decoupling. R_4 must be significantly higher resistance than R_{SRC} in order to prevent unwanted attenuation from the voltage divider formed by these two resistances. Therefore, an initial value of 499 kΩ is selected for R_4 and the high pass filter cutoff frequency is calculated using Equation 5.

$$f_{C\_HPF} = \frac{1}{2\pi * R_4 * C_4} = \frac{1}{2\pi * 499k\Omega * 10\mu F} = 0.03 \text{ Hz} \tag{5}$$

The -3 dB cutoff frequency of the filter is calculated to be 0.03 Hz. This will effectively ac couple the input signal while preserving the gain and phase behavior at 20 Hz.

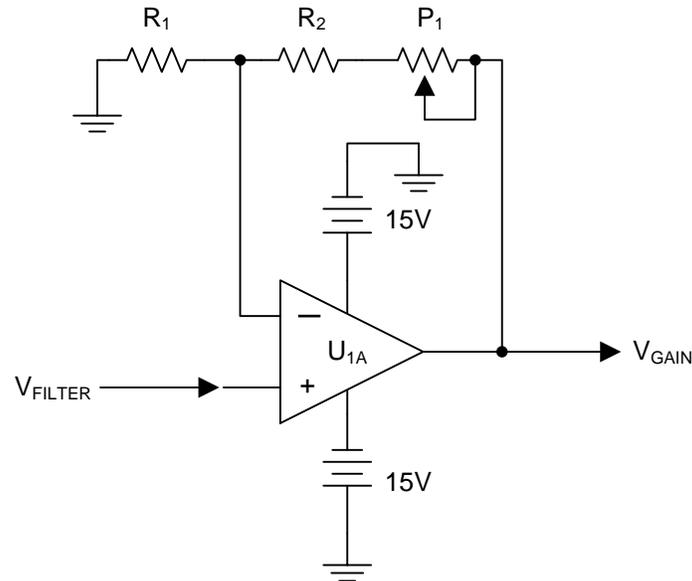### 2.1.3 Input Filter Transfer Function

The complete transfer function of the input filter is shown in Figure 4.
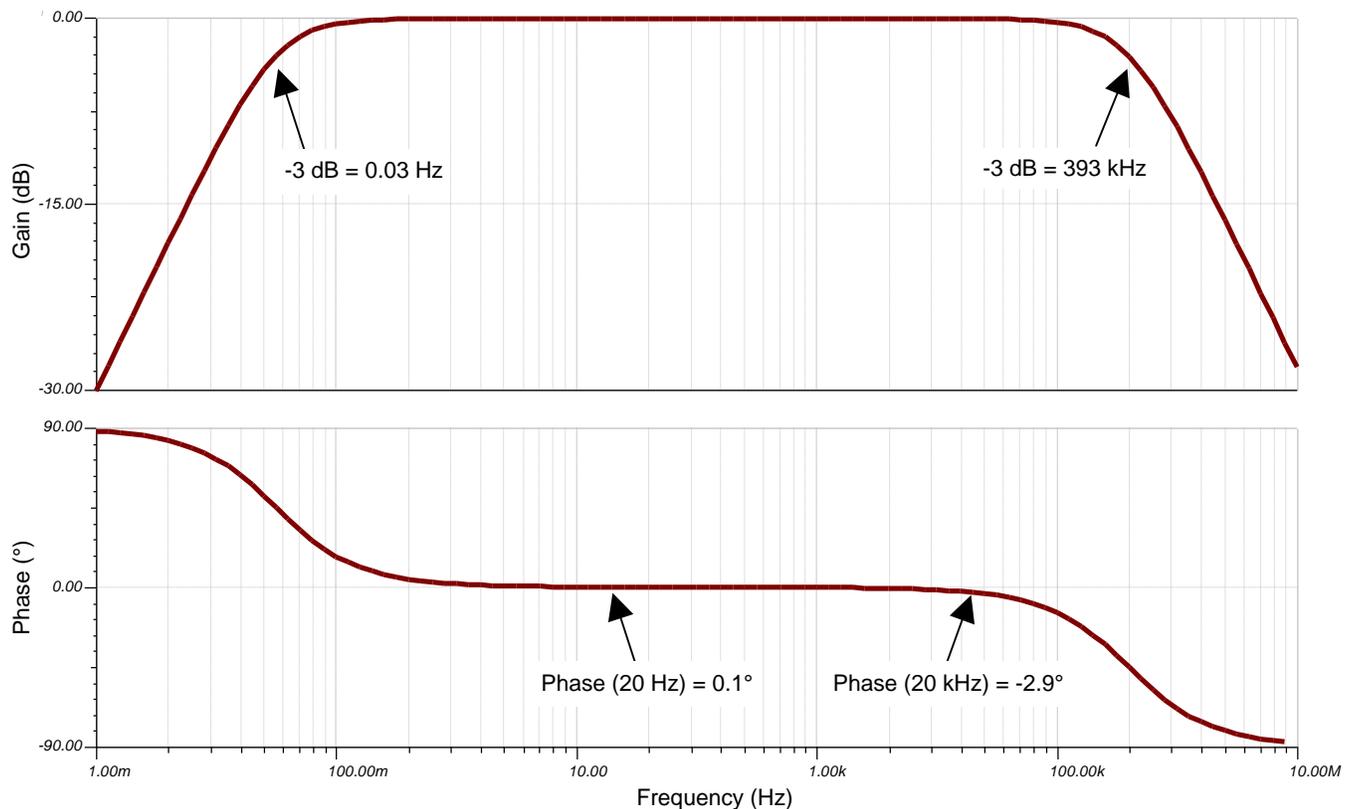
Figure 4: Transfer Function - Input Filter

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2.2   Input Gain Stage

The input signal in this design is specified at 1 V_{RMS}, or 1.414 V_{PK}. Since the OPA1642 used in this design can swing its output voltage within 200 mV from each rail and ±15 V power supply rails are provided, an input gain stage is used to amplify the input signal as needed. The gain stage is made up of amplifier U_{1A}, potentiometer P_1, and resistors R_1 and R_2, as shown in Figure 5.

Figure 5: Input Gain Stage Schematic

This straightforward non-inverting gain stage has a transfer function as defined as in Equation 6, where R_{P1} is the equivalent series resistance of potentiometer P_1.

$$V_{GAIN} = \left(1 + \frac{R_2 + R_{P1}}{R_1} \right) * V_{FILTER} \tag{6}$$

At 0% rotation of potentiometer P_1, its equivalent series resistance is 0 Ω. Therefore the transfer function of the gain stage simplifies to Equation 7.

$$V_{GAIN} = \left(1 + \frac{R_2}{R_1} \right) * V_{FILTER} \tag{7}$$

The minimum gain of the circuit is specified at 6 dB, or 2 V/V. To achieve this, the resistances of R_1 and R_2 must be equal. A value of 1 kΩ is selected for R_1 and R_2 in order to ensure low thermal noise. The maximum gain of the circuit is specified at 28 dB, or approximately 25 V/V. This gain occurs at 100% rotation of potentiometer P_1. The required value of R_{P1} is calculated by rearranging the terms of Equation 6 and solving for R_{P1}, as shown in Equation 8.

$$R_{P1} = R_1 * \left(\frac{V_{GAIN}}{V_{FILTER}} - 1 \right) - R_2 = 1k\Omega * (25 - 1) - 1k\Omega = 23k\Omega \tag{8}$$

The required value for R_{P1} is calculated to be 23 kΩ. The nearest standard value of 25 kΩ is selected as the actual value.

2.3 Tone Stack

The tone stack is a passive filter network which allows a guitarist to control the frequency response of the amplifier [1]. Many different tone stack implementations exist, but this design uses what is known as the FMV tone stack. Introduced by Fender in 1957 in the 5F6 Bassman, it was later copied by Marshall, Vox, and many other guitar amplifier manufacturers [2]. Because of its ubiquity, the circuit is very well understood with extensive analysis and documentation widely available.

The tone stack is made up of capacitors C_1, C_3, C_8, and C_9, jumper JMP1, resistor R_3, and potentiometers P_2, P_3, and P_4, as shown in Figure 6.
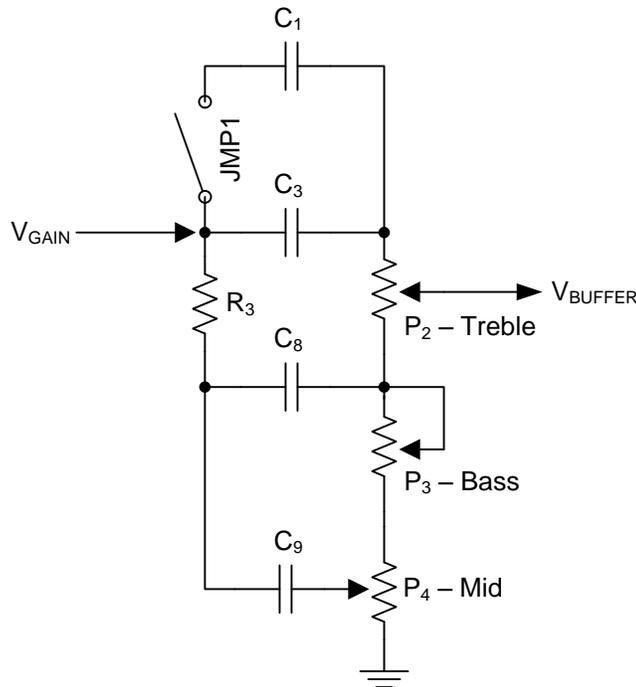
Figure 6: Tone Stack Schematic

Since the tone stack contains multiple filters with many possible states and interactive impedances, it is not trivial to analyze. Rather than perform a complete analysis here, the approach of this document will be to summarize the effect of each potentiometer on the circuit behavior, provide equations which allow the user to customize component values, and then refer to material where the full analysis is available.

2.3.1 Potentiometer Effects

The treble potentiometer P_2 acts as a balance control between the output of a high-pass filter formed by C_3 (in parallel with C_1 if JMP1 is closed) and the series combination of all three potentiometers, and the output of the complex filter created by R_3, C_8, C_9, P_3, and P_4.

The bass potentiometer P_3 sets the lower -3 dB cutoff frequency of a band-pass filter formed by R_3, C_8, P_3, and P_4. It also affects the -3 dB cutoff frequency of the treble control circuit.

The mid potentiometer P_4 controls the attenuation of the band-pass filter formed by R_3, C_9, and P_4. It also acts as a variable attenuator for the tone stack output [3].

2.3.2 Tone Stack Component Values

Calculating the values of R_3 and P_4 first allows the circuit designer to set the attenuation of all frequencies when the bass and treble controls are at 0% and the mid control is at 100%. An initial value of 25 kΩ is selected for P_4 as it is already used for the gain potentiometer P_1. The value of R_3 can then be calculated using Equation 9, where R_{P4} is the maximum resistance of P_4 and A is a positive number representing the desired amount of attenuation in dB. This design targets the frequency response of the Marshall JMP50 amplifier, where a value of 7.3 is used for A.

$$R_3 = R_{P4} * \left(10^{\frac{A}{20}} - 1\right) = 25 \text{ k}\Omega * \left(10^{\frac{7.3}{20}} - 1\right) = 33 \text{ k}\Omega \tag{9}$$

The required value for R_3 is calculated to be 33 kΩ, which is a standard value.

Once the values of R_3 and P_4 are set, the value of C_9 can be determined. This capacitor defines the upper cutoff frequency off the bass passband, which is a function of C_9 and R_3. C_3 is calculated as shown in Equation 10, where f_1 is the upper cutoff frequency of the bass passband. The Marshall JMP50 uses a cutoff frequency of 219 Hz.

$$C_9 = \frac{1}{2\pi * f_1 * R_3} = \frac{1}{2\pi * 219 \text{ Hz} * 33\text{k}\Omega} = 22 \text{ nF} \tag{10}$$

The required value for C_9 is calculated to be 22 nF, which is a standard value.

Next, the value of C_8 can be determined in order to complete the bass passband design. C_8 controls the amount of bass attenuation when the bass potentiometer is at 0% (in a short-circuit condition), so the resistance of P_3 is not included in the calculation. A value of 1 MΩ is selected for P_3, consistent with the values used in the Marshall JMP50. This ensures that the lower end of the bass passband is well below the lowest frequencies output by a guitar. C_8 is calculated as shown in Equation 11, where f_2 is the lower cutoff frequency of the bass passband. The Marshall JMP50 uses a cutoff frequency of 62 Hz.

$$C_8 = \frac{1}{2\pi * f_2 * (R_3 + R_{P4})} - C_9 = \frac{1}{2\pi * 62 \text{ Hz} * (33 \text{ k}\Omega + 25 \text{ k}\Omega)} - 22 \text{ nF} = 22 \text{ nF} \tag{11}$$

The required value for C_8 is also calculated to be 22 nF.

Finally, the values of C_3 and P_2 can be selected, which set the cutoff frequency of the treble high pass filter. A value of 250 kΩ is selected for P_2, and C_3 is calculated as shown in Equation 12, where f_3 is the cutoff frequency of the treble high pass filter and R_{P2} is the maximum resistance of P_2. The Marshall JMP50 uses a cutoff frequency of 1.4 kHz.

$$C_3 = \frac{1}{2\pi * f_3 * R_{P2}} = \frac{1}{2\pi * 1.4 \text{ kHz} * 250 \text{ k}\Omega} = 455 \text{ pF} \tag{12}$$

The required value for C_3 is calculated to be 455 pF. The nearest standard value of 470 pF is selected as the actual value.

Closing switch JMP1 connects C_1 in parallel with C_3, adding the two capacitances to the cutoff frequency calculation. If C_1 also has a value of 470 pF, the cutoff frequency will be reduced by a factor of two to approximately 700 Hz [4].

The calculated component values and associated tone stack characteristics are summarized in Table 2.

Table 2. Summary of Tone Stack Component Values and Performance Characteristics

	Components	Values	Behavior
Overall attenuation	P_4	25 kΩ	-7.3 dB
	R_3	33 kΩ	
Bass passband upper cutoff frequency	C_9	22 nF	219 Hz
Bass passband lower cutoff frequency	C_8	22 nF	62 Hz
	P_3	1 MΩ	
Treble high pass cutoff frequency	C_3	470 pF	1.4 kHz
	P_2	250 kΩ	
Mid boost high pass cutoff frequency	C_1	470 pF	700 Hz

2.3.3 Further Reading

If the reader wishes to expand their understanding of the FMV tone stack, a more thorough analysis is available in *Circuit Analysis of a Legendary Tube Amplifier: The Fender Bassman 5F6-A* by Richard Kuehnel. Another useful resource is *Designing Tube Preamps for Guitar and Bass* by Merlin Blencowe, which discusses the FMV tone stack as well as several other topologies along with their advantages and disadvantages.

3 Component Selection

3.1 Amplifier

This tone stack circuit must provide gain and accurate control over the frequency response of the input audio signal while introducing as little distortion or noise as possible. Therefore, the amplifier selected must have very low distortion and noise performance in the audio frequency range, even when high source impedances are present [5]. A wide supply voltage range is also required, as most professional audio circuits use large split supplies in order to avoid output clipping. Low quiescent current and relatively low cost are also desirable qualities which help to maintain an efficient design.

The OPA1642 is an excellent choice for this high-performance audio application, with total harmonic distortion + noise (THD+N) of only -126 dB (0.00005%) and input voltage noise density of 5.1 nV/√Hz. The amplifier can utilize power supply voltages up to ±18 V while consuming only 1.8 mA of quiescent current per channel, and its reasonable price point ensures that the total solution cost remains competitive.

3.2 Passive Component Selection

3.2.1 Resistor Selection

The type of resistors used in an ultra-low distortion audio circuit can have a significant impact on the circuit's overall performance. Real resistors have a certain amount of nonlinearity, which results in unwanted contributions to distortion and noise [6]. The most common sources of resistor nonlinearity are temperature coefficient of resistance (TC_R), which describes how the resistance changes as a function of temperature, and voltage coefficient of resistance (VC_R), which describes how the resistance changes as a function of applied voltage. Both VC_R and TC_R are related to the resistor's self-heating – as the voltage across the resistor increases, the current through the resistor increases and its temperature rises.

Two of the most common types of surface mount resistors are thick film and thin film. Thin film resistors typically perform better than thick film resistors, but thin film resistors also typically cost several times as much. When high-level audio signals are involved, the lower VC_R and TC_R of thin film resistors can become critical to achieving ultra-low distortion performance.

This design involves audio signals with maximum amplitude of 15 V_{PK}, or 10.6 V_{RMS}. Despite the significant voltage, current through the signal path resistors remains low and meaningful self-heating does not occur. Therefore, all signal path resistors on the board are thick film, ±1% tolerance, 0.1 watt devices in a 0603 package.

3.2.2 Capacitor Selection

Like resistors, capacitors also have a voltage coefficient (VC_C), which describes how the capacitance changes as a function of applied voltage. This change in capacitance results in unwanted distortion [7], so any capacitors in the audio signal path which can be subjected to significant voltages should have low VC_C.
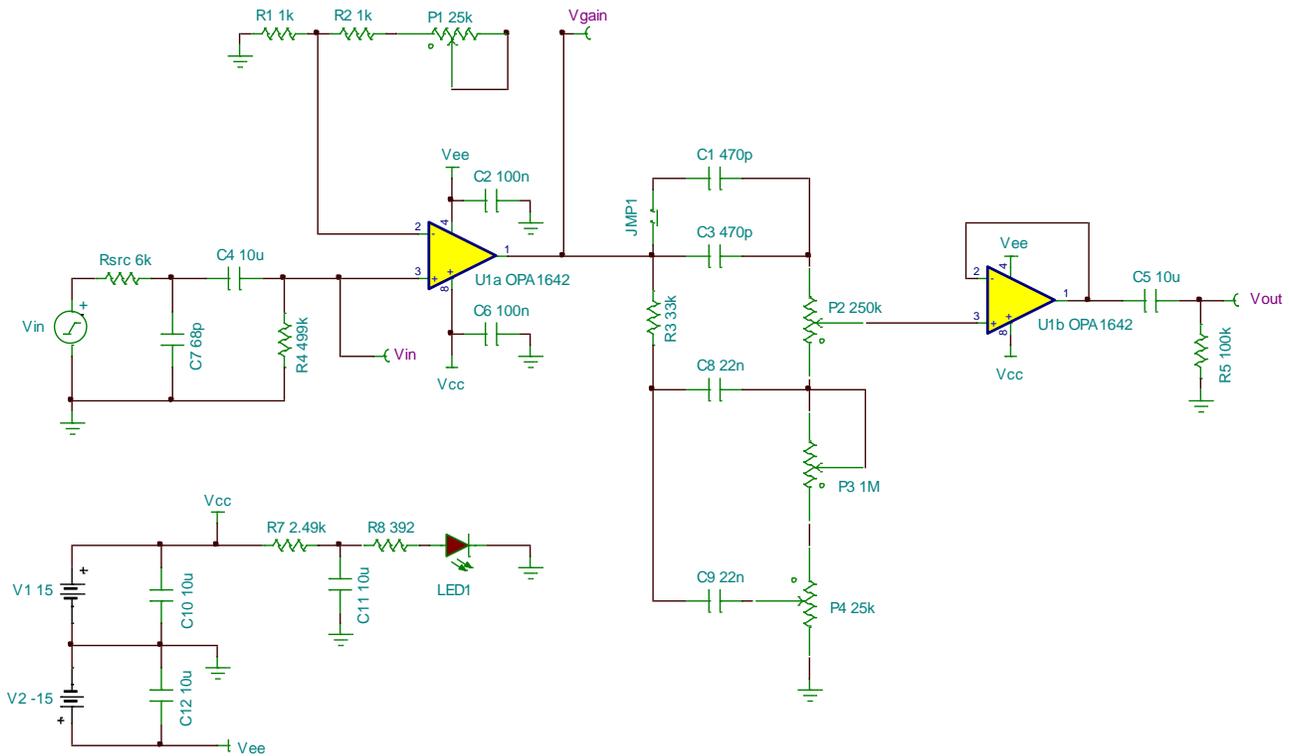
The critical signal path capacitors in this design are C_7 in the input filter and C_1, C_3, C_8, and C_9 in the tone stack. For these components NP0-type capacitors are used. All other capacitors for ac coupling and power supply bypass are of type X7R.

3.2.3 Potentiometer Selection

Other than the maximum resistance and taper characteristics, the potentiometers selected for this design are not critical. Single-turn rotary potentiometers with ±20% tolerance and right-angle PCB mount termination were selected for this application.

4 Simulation

The TINA-TI™ schematic shown in Figure 7 includes the circuit values obtained in the design process. The source impedance of V_{IN}, the input signal, is included as a discrete resistance R_{SRC}. A load resistance of 100 kΩ is added to simulate the input resistance of the audio analyzer which is used for real-world measurements.

Figure 7: TINA-TI™ Schematic

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4.1 Gain Characteristic

The result of the simulated gain characteristic as a function of gain potentiometer P_1 rotation is shown in Figure 8.
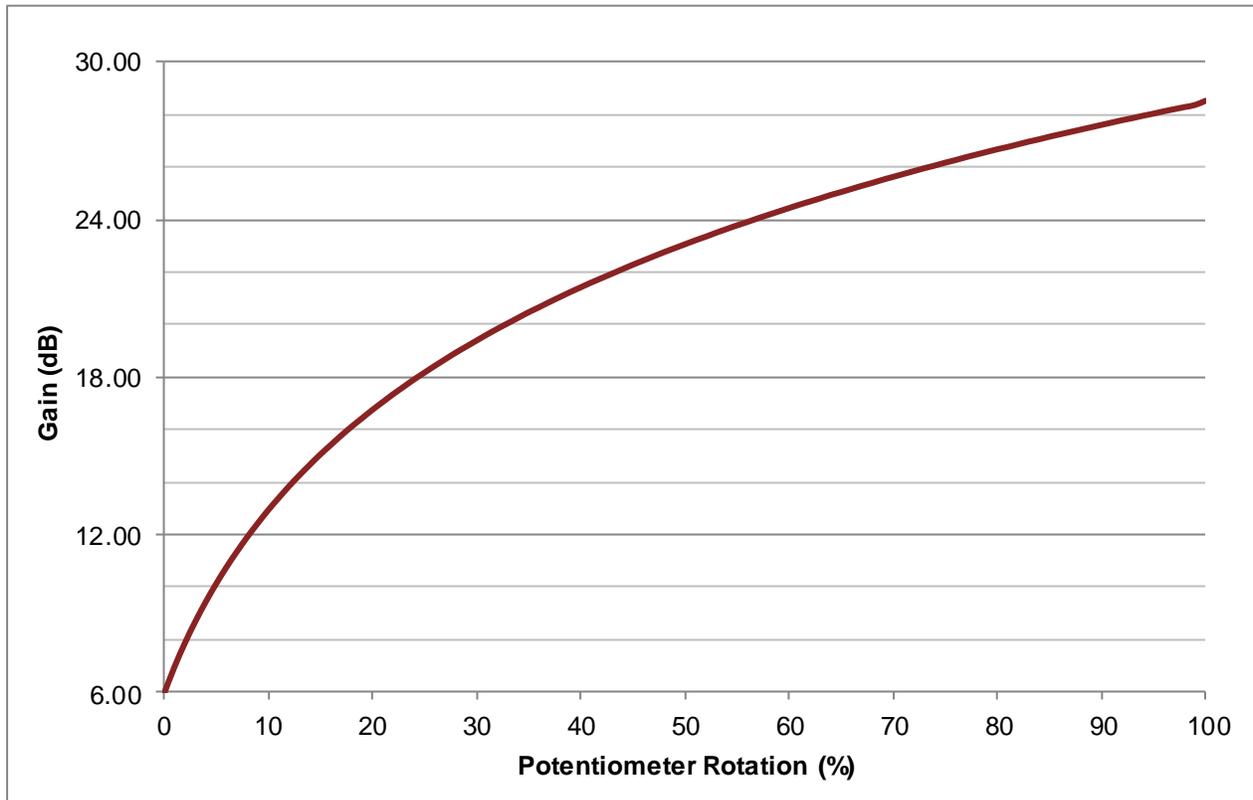
Figure 8: Simulated Gain Characteristic

4.2 Frequency Response

4.2.1 Input Filter and Gain Stage

The result of the simulated ac analysis of the input filter and gain stage when gain = 6 dB is shown in Figure 9.
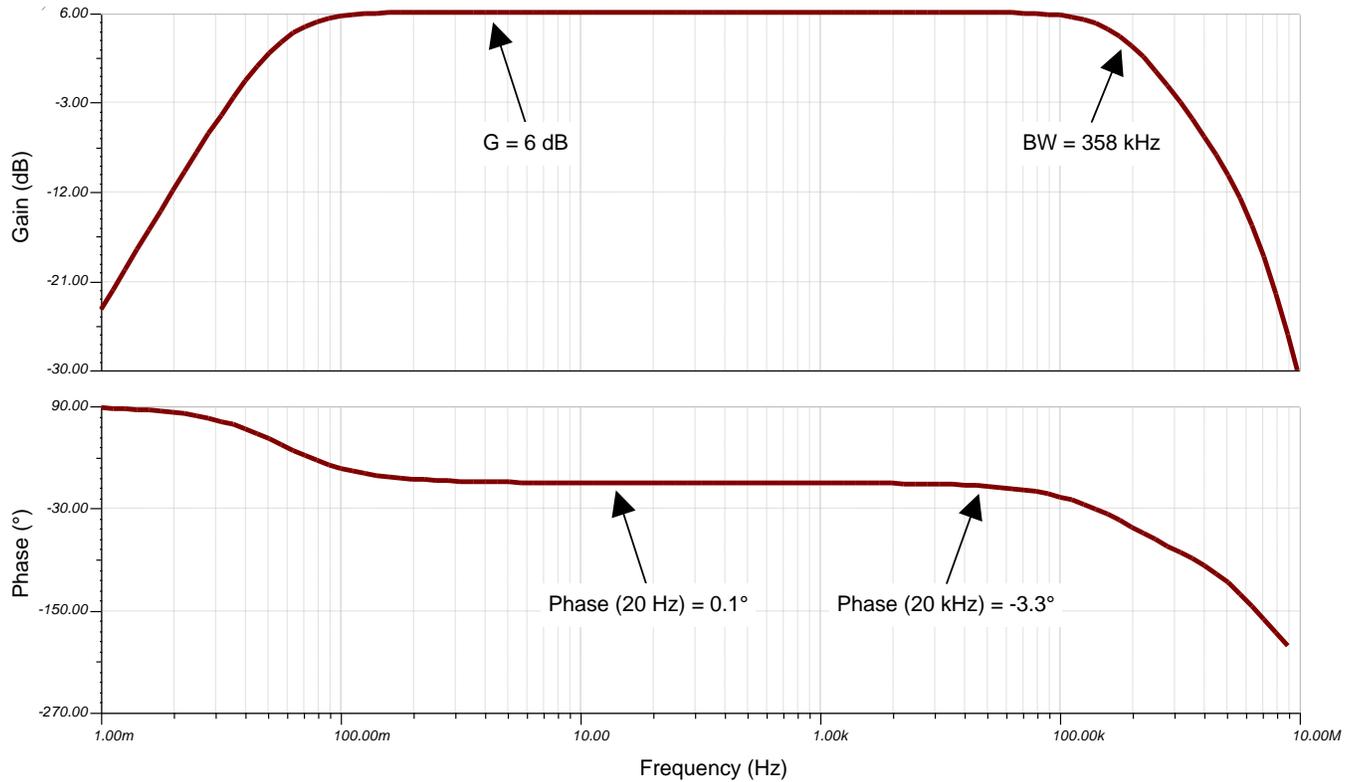
Figure 9: Simulated AC Analysis – Input Filter and Gain Stage

The gain of the simulation throughout the audio band was measured to be 6 dB. The -3 dB bandwidth was 358 kHz. The phase of the simulation was 0.1° at 20 Hz and -3.3° at 20 kHz.

4.2.2 Tone Stack

4.2.2.1 Treble Control

The result of the simulated ac analysis of the complete circuit as the treble potentiometer P_2 is rotated, when gain = 6 dB, the mid and bass potentiometers are set to 50%, and mid boost is off, is shown in Figure 10.
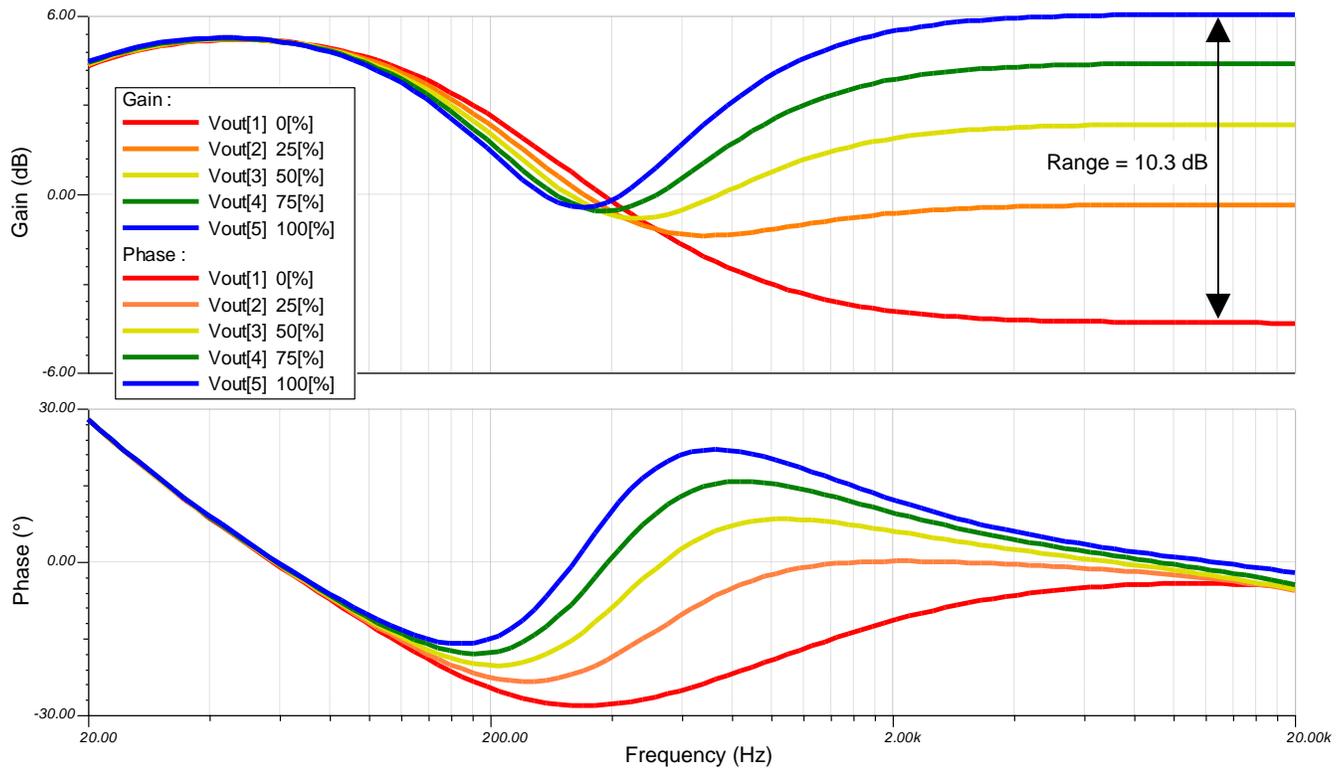
**Figure 10: Simulated AC Analysis – Treble Control**

In this condition, the treble gain varies from -4.3 dB at 0% potentiometer rotation to +6.0 dB at 100% potentiometer rotation. This gives an adjustment range of 10.3 dB, which meets the design requirement of 10 dB.

4.2.2.2   Mid Control

The result of the simulated ac analysis of the complete circuit as the mid potentiometer P_4 is rotated, when gain = 6 dB, the treble and bass potentiometers are set to 50%, and mid boost is off, is shown in Figure 11.
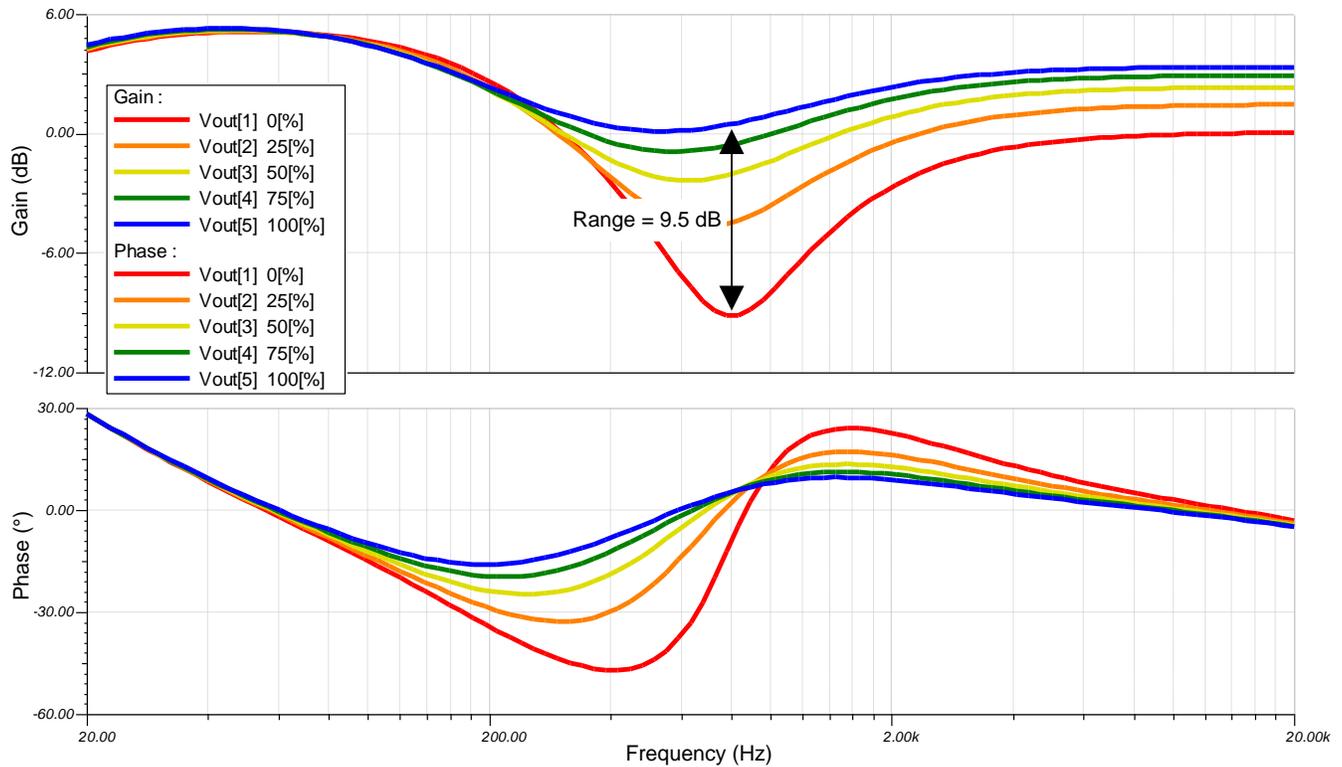
Figure 11: Simulated AC Analysis – Mid Control

In this condition, the mid gain varies from -9.1 dB at 0% potentiometer rotation to +0.4 dB at 100% potentiometer rotation. This gives an adjustment range of 9.5 dB, which meets the design requirement of 6 dB.

4.2.2.3 Mid Boost

The result of the simulated ac analysis of the complete circuit as the mid boost jumper is connected and disconnected, when gain = 6dB and all potentiometers are set to 50%, is shown in Figure 12.
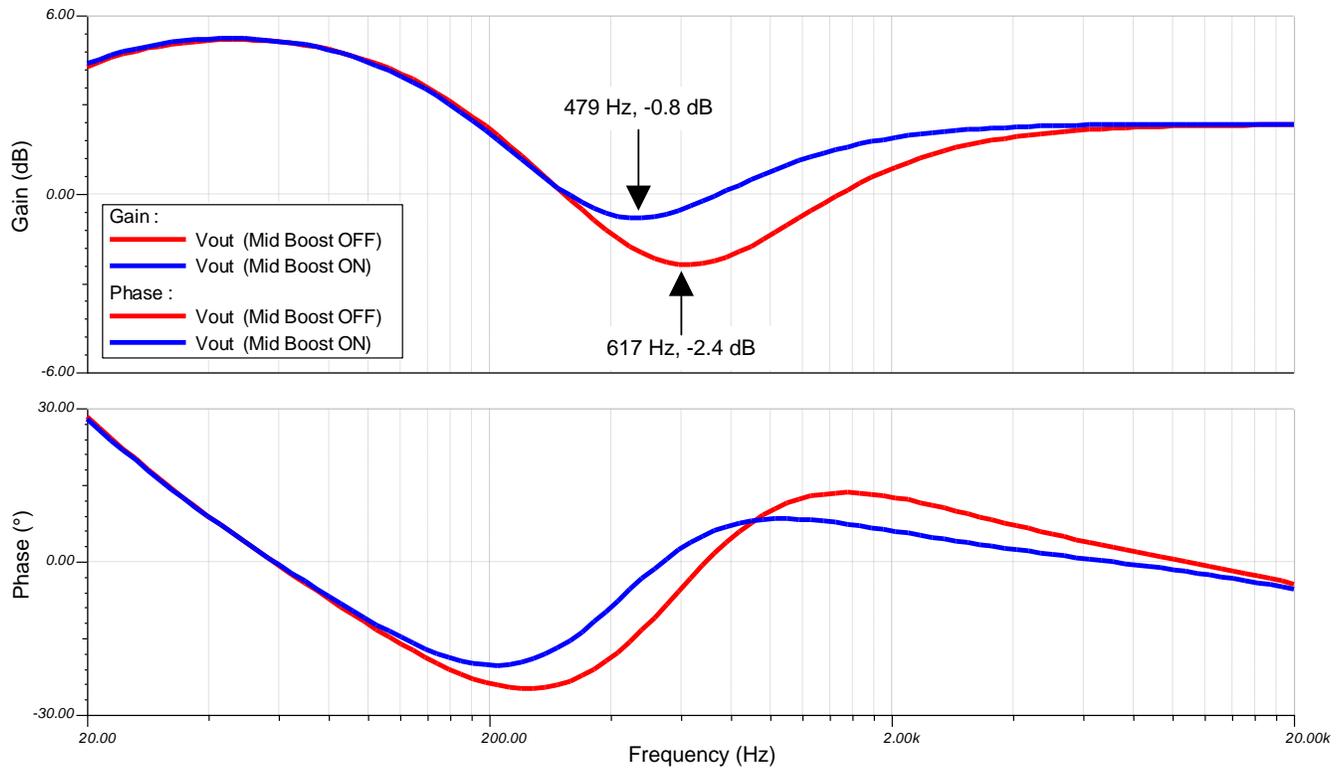
Figure 12: Simulated AC Analysis – Mid Boost

Activating the mid boost lowered the mid cut frequency from 617 Hz to 479 Hz and boosted the gain at the cut frequency from -2.4 dB to -0.8 dB.

4.2.2.4 Bass Control

The result of the simulated ac analysis of the complete circuit as the bass potentiometer P_3 is rotated, when gain = 6 dB, the treble and mid potentiometers are set to 50%, and mid boost is off, is shown in Figure 13.
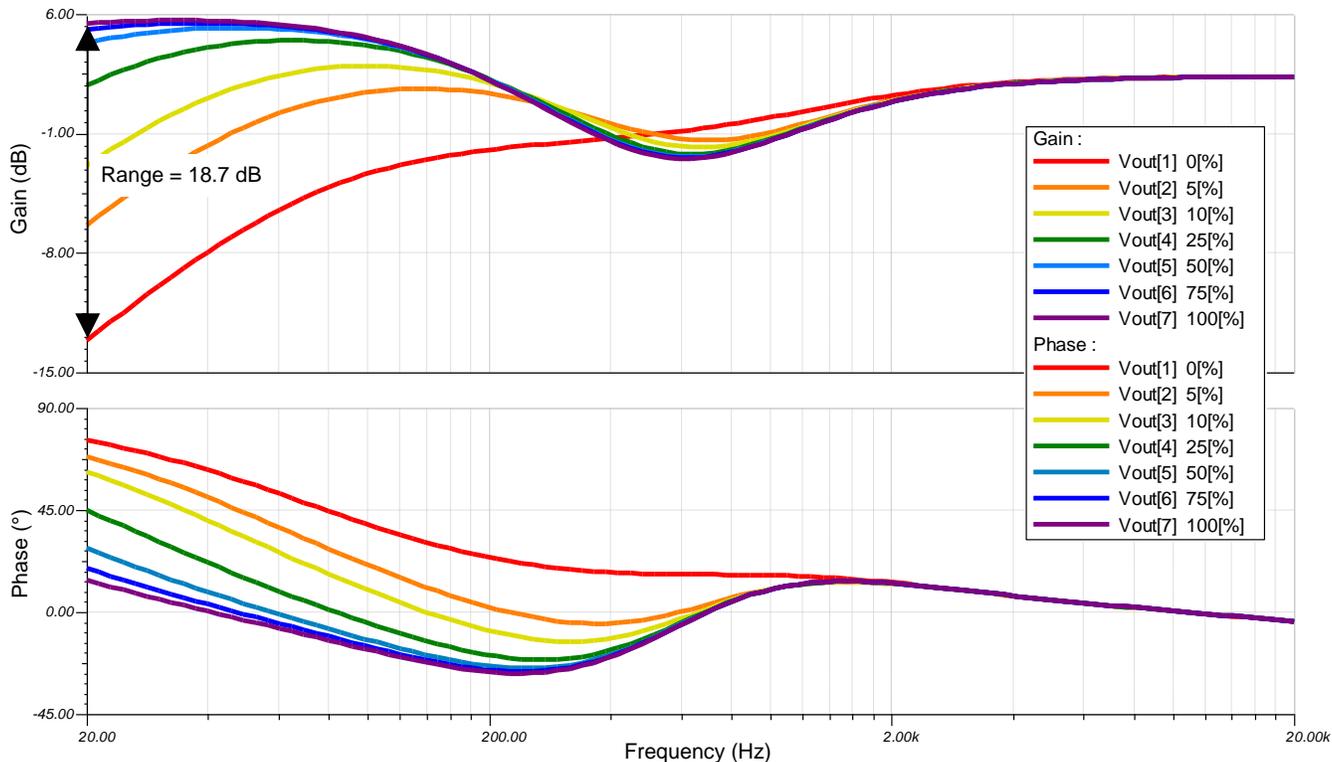
Figure 13: Simulated AC Analysis – Bass Control

In this condition, the bass gain varies from -13.2 dB at 0% potentiometer rotation to +5.5 dB at 100% potentiometer rotation. This gives an adjustment range of 18.7 dB, which meets the design requirement of 15 dB.

4.3 THD+N Performance

Unfortunately, TI's op amp macromodels do not currently support proper THD+N analysis. However, the THD+N ratio of a circuit (when noise is dominant) can be predicted from the total noise analysis by using Equation 13, where V_N is the total voltage noise in V_{RMS} over a specified bandwidth and V_F is the fundamental signal amplitude in V_{RMS}.

$$\text{THD} + \text{N(\%)} = \sqrt{\frac{V_N^2}{V_F^2}} * 100 \qquad (13)$$

The result of the simulated total noise analysis at gain = 6 dB, all tone potentiometers set to 100%, and mid boost on is shown in Figure 14. The audio analyzer which is used for real-world measurements will be set to a measurement bandwidth of 90 kHz, so this simulated noise analysis is performed to 90 kHz.
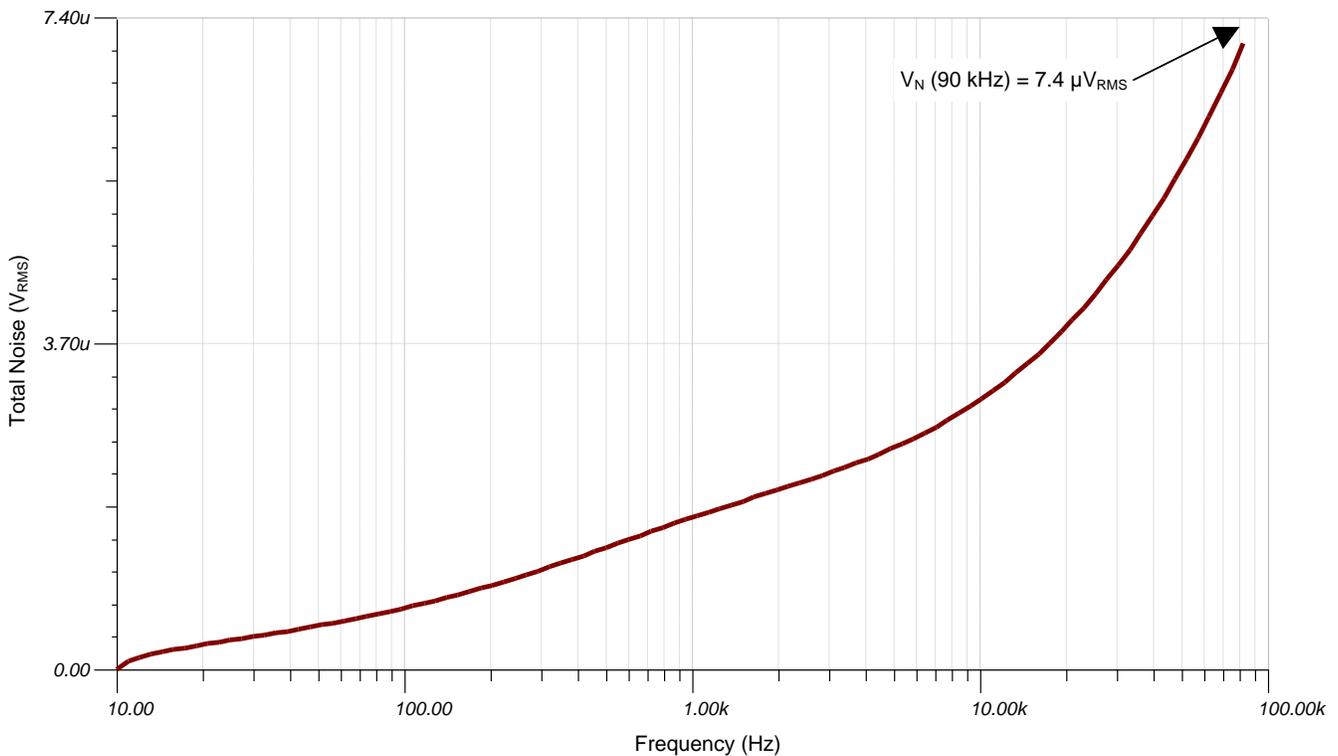
Figure 14: Simulated Total Noise Analysis

The total noise at 90 kHz was found to be 7.4 μV_{RMS}. Given our input signal amplitude of 1 V_{RMS}, the predicted THD+N ratio is calculated using Equation 14.

$$\text{THD} + \text{N(\%)} = \sqrt{\frac{V_N^2}{V_F^2}} * 100 = \sqrt{\frac{(7.4\ \mu V_{RMS})^2}{(1\ V_{RMS})^2}} * 100 = 0.00074\% = -102.6\ \text{dB} \qquad (14)$$

The simulated THD+N ratio was found to be -102.6 dB (0.00074%), which meets the design requirement of -100 dB. However, this does not account for the possibility of harmonic distortion due to output clipping which can occur at higher gain settings.

4.4 Resistor Nonlinearity

As mentioned in Section 3.2.1, resistor nonlinearity due to TC_R and VC_R can have a negative effect on distortion performance. In order to determine if thin film resistors are required, the current through feedback resistors R_1 and R_2 is simulated during the worst-case condition when V_{IN} = 20 Hz and gain = 6 dB. The result of the transient analysis is shown in Figure 15.
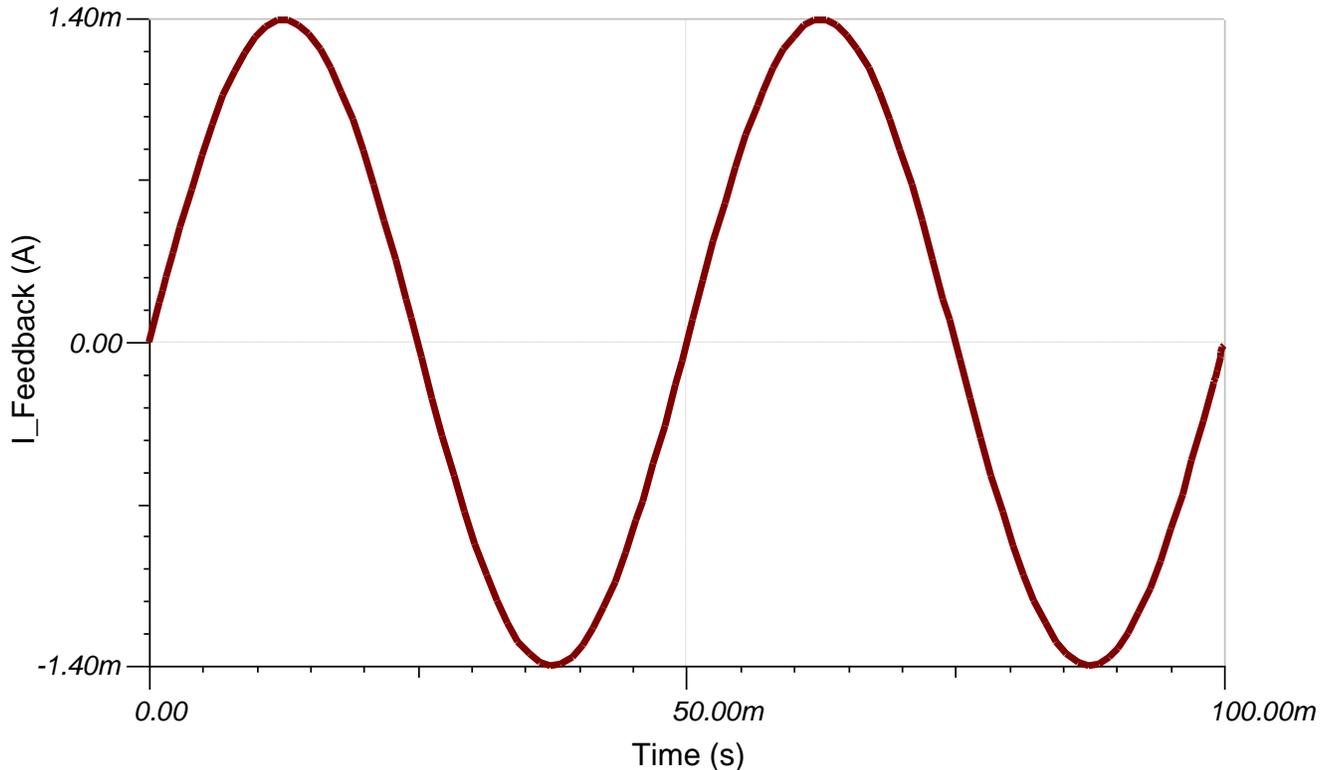
Figure 15: Simulated Transient Analysis – Feedback Resistor Current

In this condition, the maximum current through the feedback network is approximately 1.4 mA. We calculate the maximum power dissipation through the feedback network using Equation 15.

$$P_{DIS}(W) = I^2 * R = 1.4\ mA^2 * 1k\Omega = 2\ mW \tag{15}$$

The maximum power dissipation of approximately 2 mW is well below the resistors' power handling of 100 mW, so TC_R and VC_R will not be an issue and thin film resistors are not required to achieve low-distortion performance.

4.5 Simulated Results Summary

Table 3 summarizes the simulated performance of the design.

Table 3. Comparison of Design Goals and Simulated Performance

	Goal	Simulated
THD+N ratio at 1 kHz	-100 dB (0.001%)	-102.6 dB (0.00074%)
Treble adjustment range	10 dB	10.3 dB
Mid adjustment range	6 dB	9.5 dB
Bass adjustment range	15 dB	18.7 dB

5 PCB Design

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

5.1 PCB Layout

The PCB used in this design is a 3.4" by 3.4" square. This generous size allows for efficient routing of critical components and the use of larger RCA, ¼", and banana jacks, as well as the four required potentiometers. The high-level approach to this layout was to place nearly all components on the top layer, with the op amp in the center of the board, input connections on the left, output connections on the right, and gain and tone control potentiometers on the bottom. The power supply bulk capacitors were placed on the bottom layer close to the banana jacks. The two low-frequency tone control capacitors were also placed on the bottom layer close to their associated potentiometers.

Standard precision analog PCB layout practices were used in order to achieve the best possible performance. All passive components in the analog signal path are placed and routed very tightly in order to minimize parasitics, and all decoupling capacitors are located very close to their associated power pins. Solid copper planes on both layers provide an excellent low-impedance path for return currents to ground, and stitching vias are used where necessary.

Connections to the split power supply are made at J_5, J_6, and J_7. Connections to the audio inputs and outputs are made at J_1, J_2, J_3, and J_4. RCA connectors J_1 and J_2 are used to easily connect to the test equipment when measuring system performance, while ¼" connectors J_3 and J_4 are used to connect standard guitar cables.

The PCB layout for both layers is shown in Figure 16.
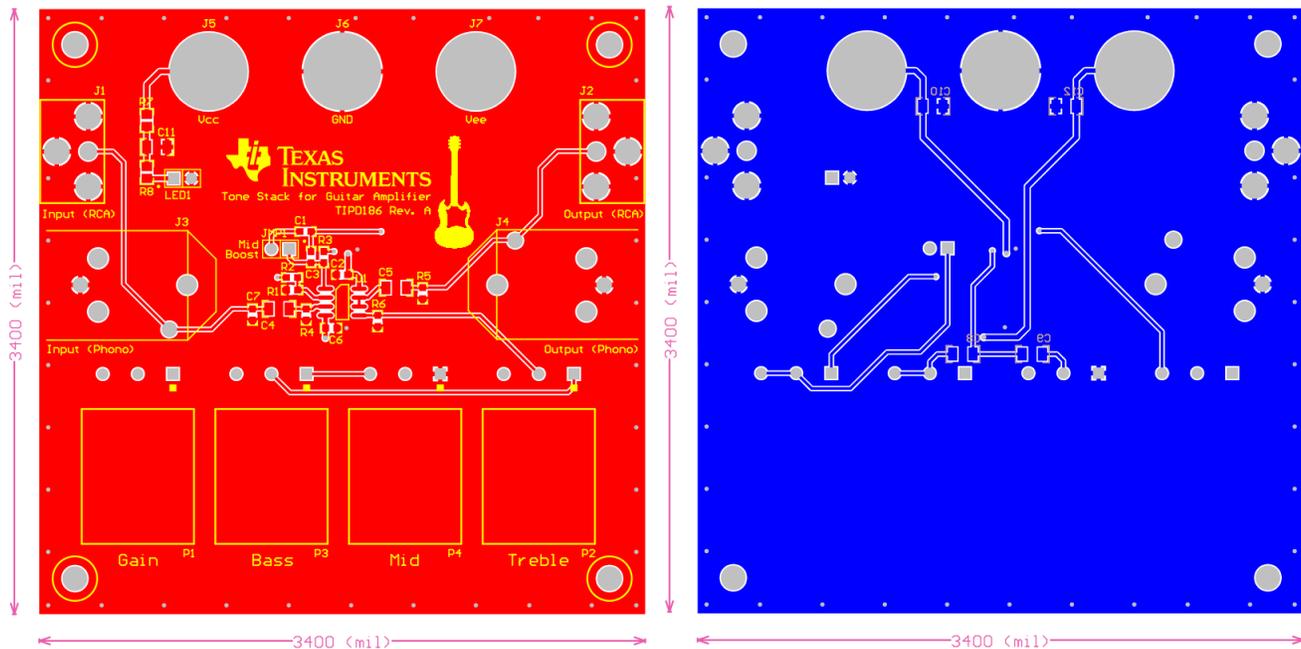
Figure 16: PCB Layout

6 Verification & Measured Performance

6.1 Bench Test Hardware Setup

The tone stack circuit defined by this reference design is intended for use within a complete guitar amplifier system. However, the circuit is also a standalone functional block whose real-world performance can be characterized. The convenient input, output and power connectors on the PCB allow the circuit to be easily tested on a bench using standard lab equipment. The test setup used consists of the components listed below. Figure 17 shows the bench test setup (computer not shown).

1. High performance audio analyzer: Provides the audio input and measures the audio output of the system.

2. Bode analyzer: Measures the gain and phase response of the system over frequency.

3. Personal computer (PC): Communicates with and controls the audio analyzer and Bode analyzer through a digital interface. Software provided by the hardware manufacturers allows the user to specify signal characteristics and perform measurements.

4. Triple output power supply: Provides ±15 V power supply rails to the system.
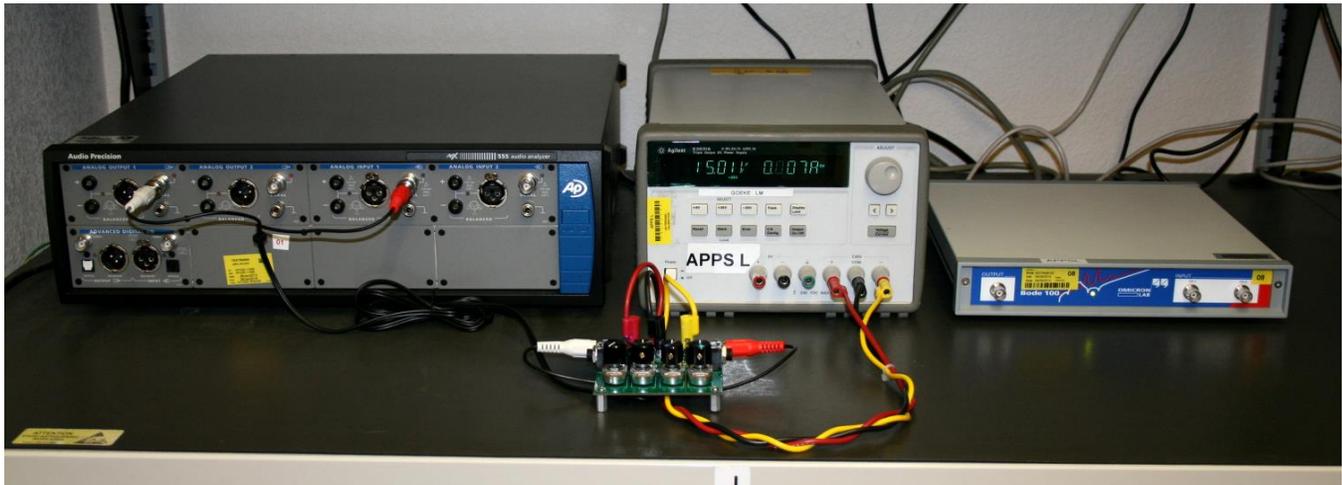
Figure 17: Bench Test Hardware Setup

6.2 Gain Characteristic

The result of the measured gain characteristic as a function of gain potentiometer P_1 rotation is shown in Figure 18. Gain was measured at 0%, 25%, 50%, 75%, and 100% rotation.
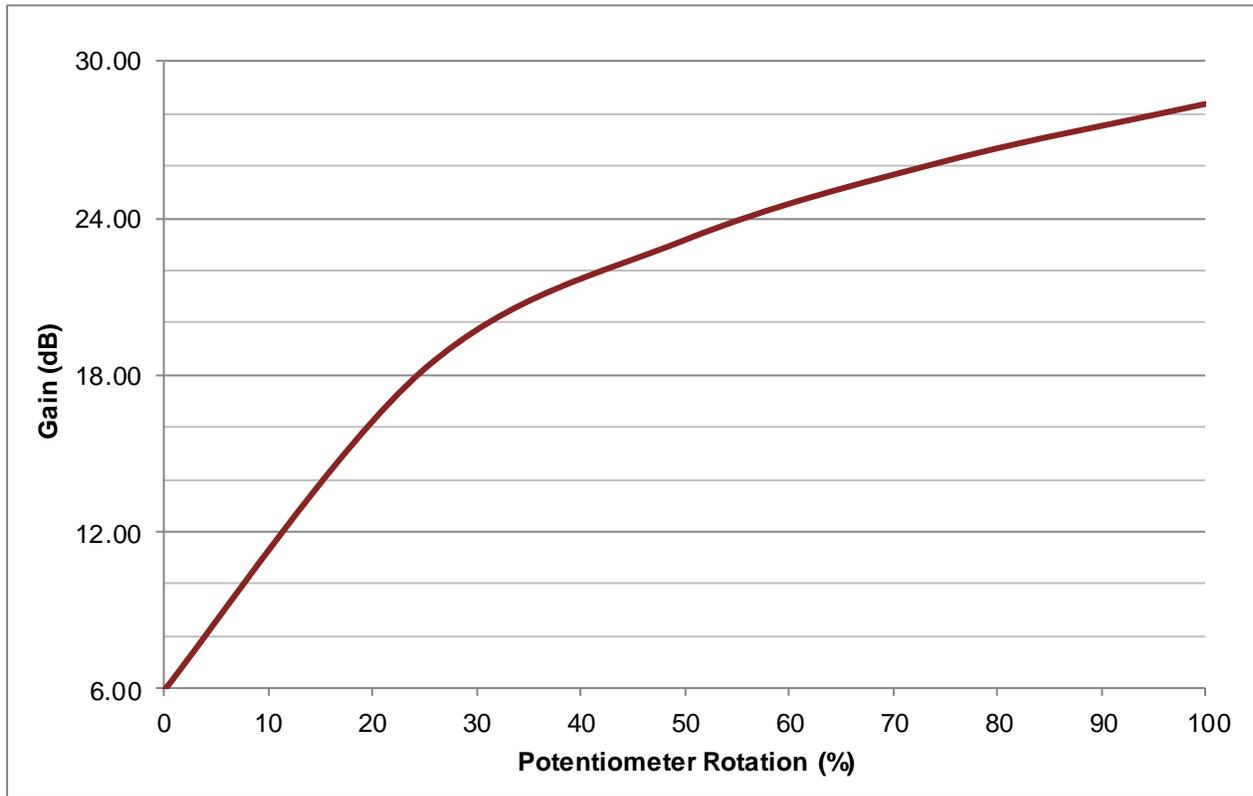
Figure 18: Measured Gain Characteristic

6.3 Frequency Response

6.3.1 Input Filter and Gain Stage

The result of the measured ac analysis of the input filter and gain stage when gain = 6 dB is shown in Figure 19.
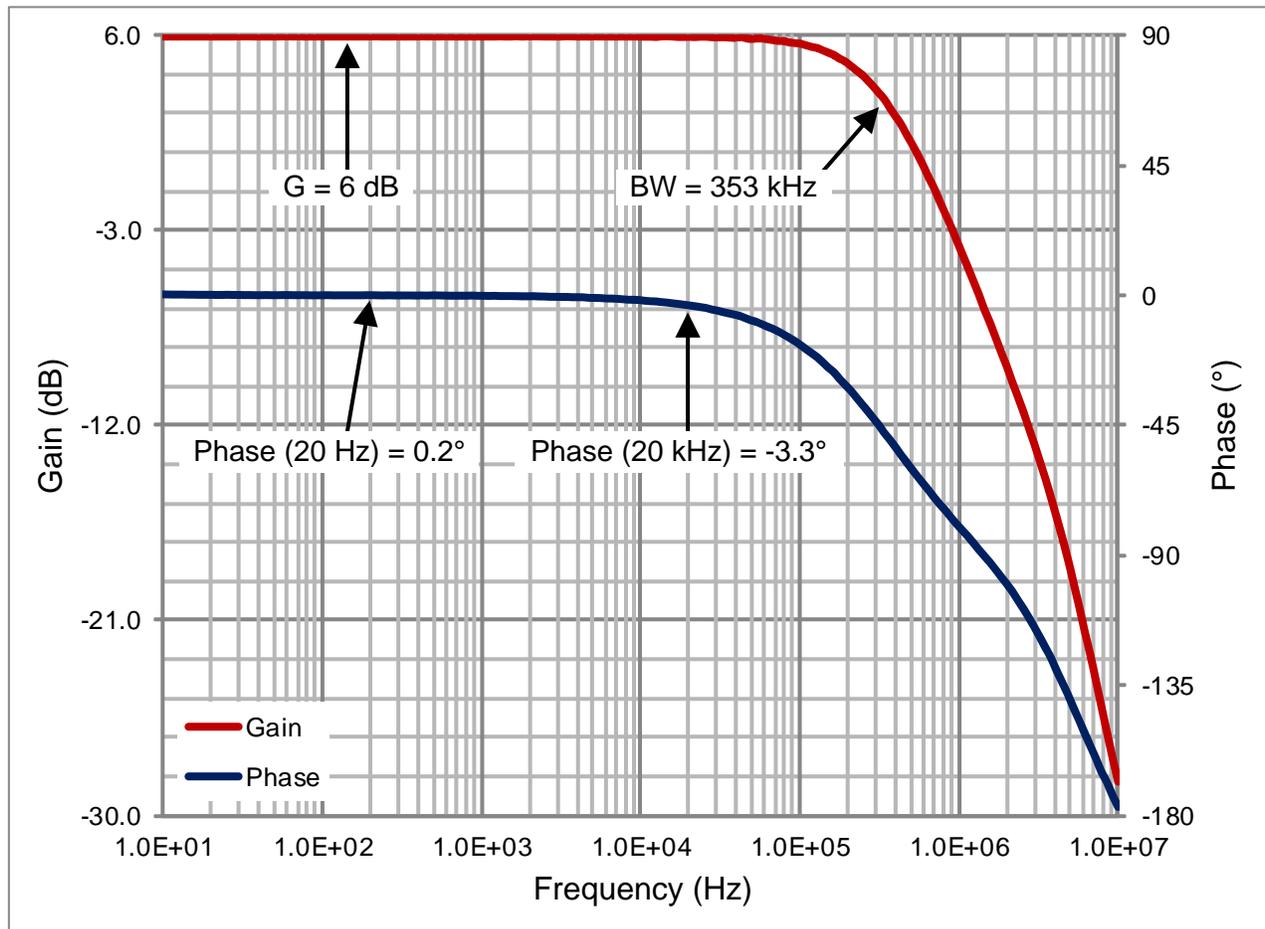
Figure 19: Measured AC Analysis – Input Filter and Gain Stage

The gain of the measurement throughout the audio band was measured to be 6 dB. The measured -3 dB bandwidth was 353 kHz, which correlates very well to the simulated -3 dB bandwidth of 358 kHz. The phase of the circuit was measured to be 0.2° at 20 Hz and -3.3° at 20 kHz, which is nearly the exact result found in simulation.

6.3.2   Tone Stack

6.3.2.1   Treble Control

The result of the measured ac gain analysis of the complete circuit as the treble potentiometer P_2 is rotated, when gain = 6 dB, the mid and bass potentiometers are set to 50%, and mid boost is on, is shown in Figure 20.
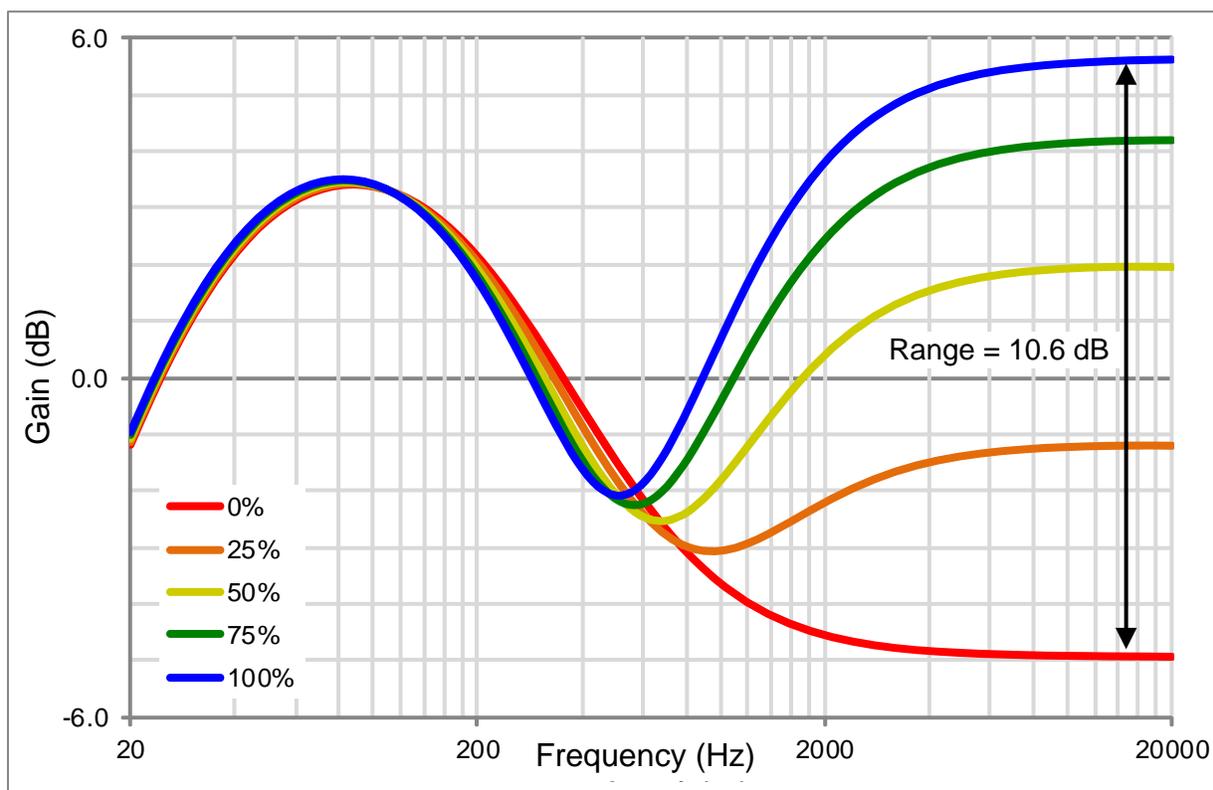
Figure 20: Measured AC Gain Analysis – Treble Control

In this condition, the treble gain varies from -5.0 dB at 0% potentiometer rotation to +5.6 dB at 100% potentiometer rotation. This gives an adjustment range of 10.6 dB, which meets the design requirement of 10 dB.

The result of the measured ac phase analysis under the same conditions is shown in section A.3.

6.3.2.2 Mid Control

The result of the measured ac gain analysis of the complete circuit as the mid potentiometer P_4 is rotated, when gain = 6 dB, the treble and bass potentiometers are set to 50%, and mid boost is on, is shown in Figure 21.
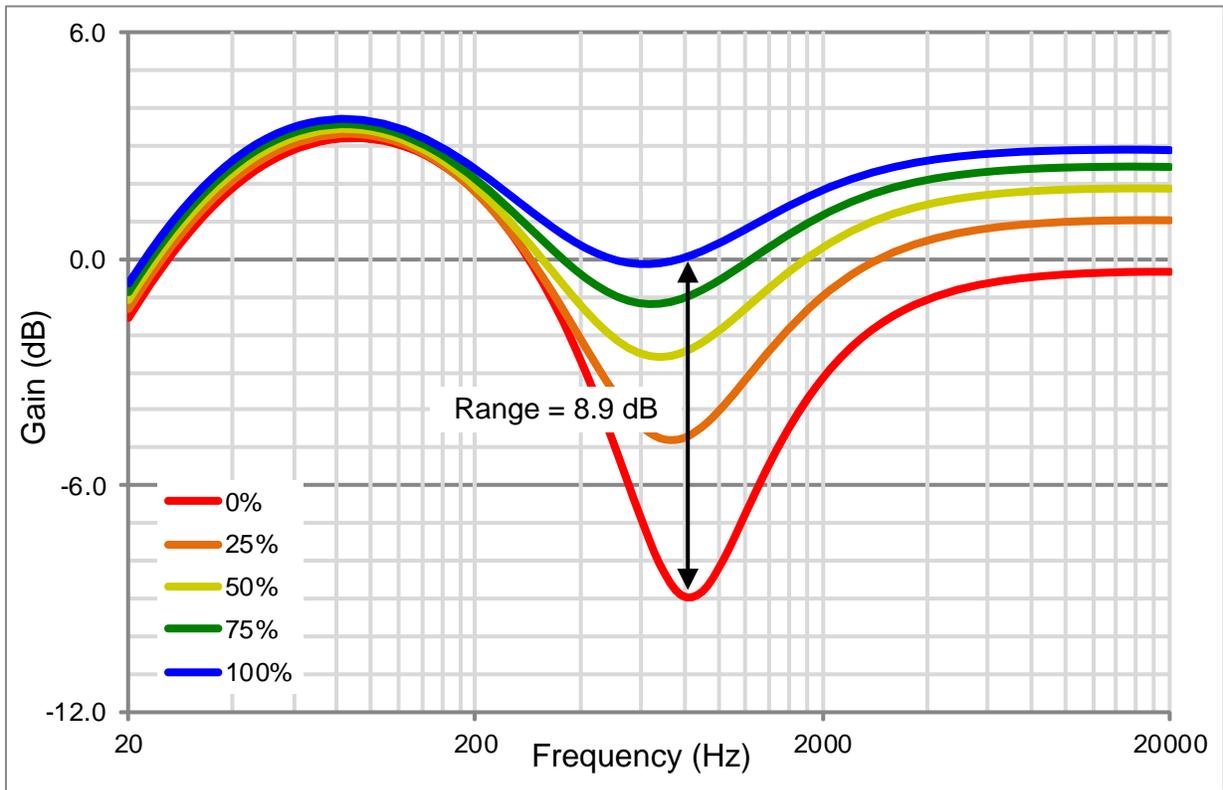
Figure 21: Measured AC Gain Analysis – Mid Control

In this condition, the mid gain varies from -8.9 dB at 0% potentiometer rotation to 0 dB at 100% potentiometer rotation. This gives an adjustment range of 8.9 dB, which meets the design requirement of 6 dB.

The result of the measured ac phase analysis under the same conditions is shown in section A.3.

6.3.2.3 Mid Boost

The result of the measured ac gain analysis of the complete circuit as the mid boost jumper is connected and disconnected, when gain = 6dB and all potentiometers are set to 50%, is shown in Figure 22.
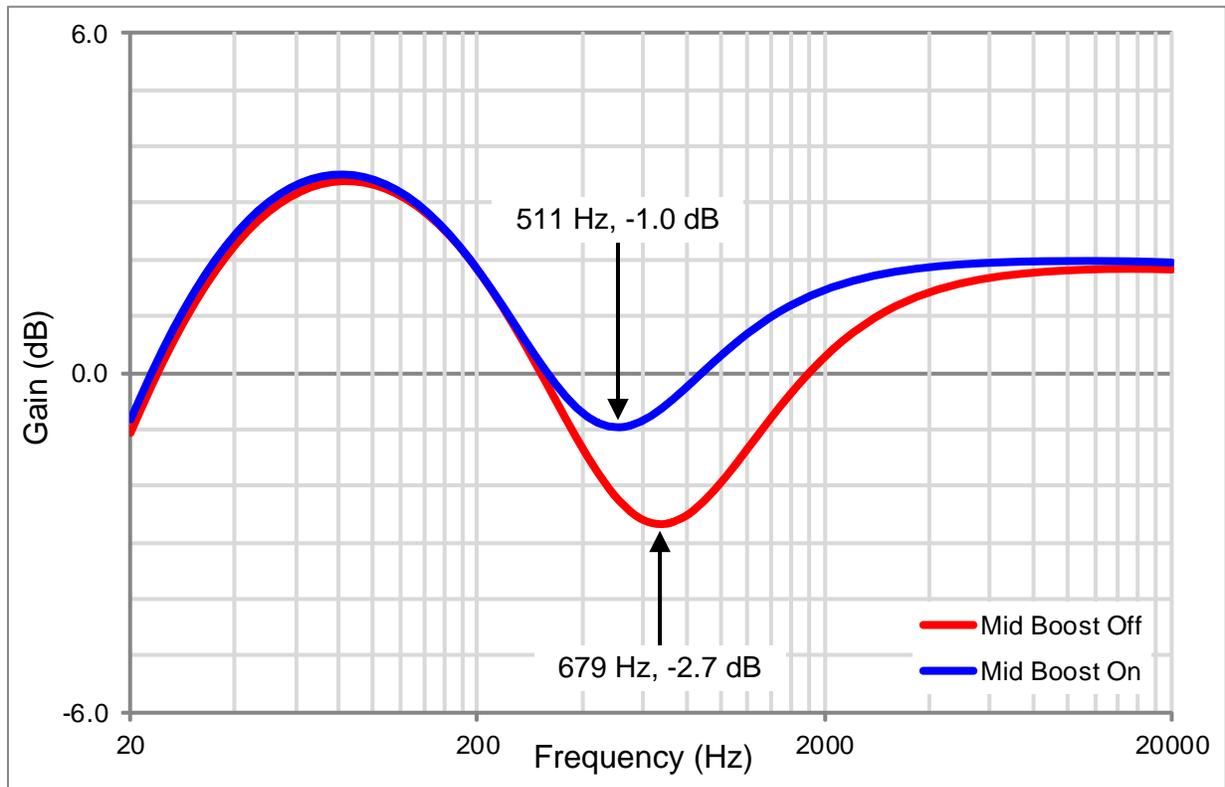
Figure 22: Measured AC Gain Analysis – Mid Boost

Activating the mid boost lowered the mid cut frequency from 679 Hz to 511 Hz and boosted the gain at the cut frequency from -2.7 dB to -1.0 dB.

The result of the measured ac phase analysis under the same conditions is shown in section A.3.

6.3.2.4   Bass Control

The result of the measured ac gain analysis of the complete circuit as the bass potentiometer P_3 is rotated, when gain = 6 dB, the treble and mid potentiometers are set to 50%, and mid boost is on, is shown in Figure 23.
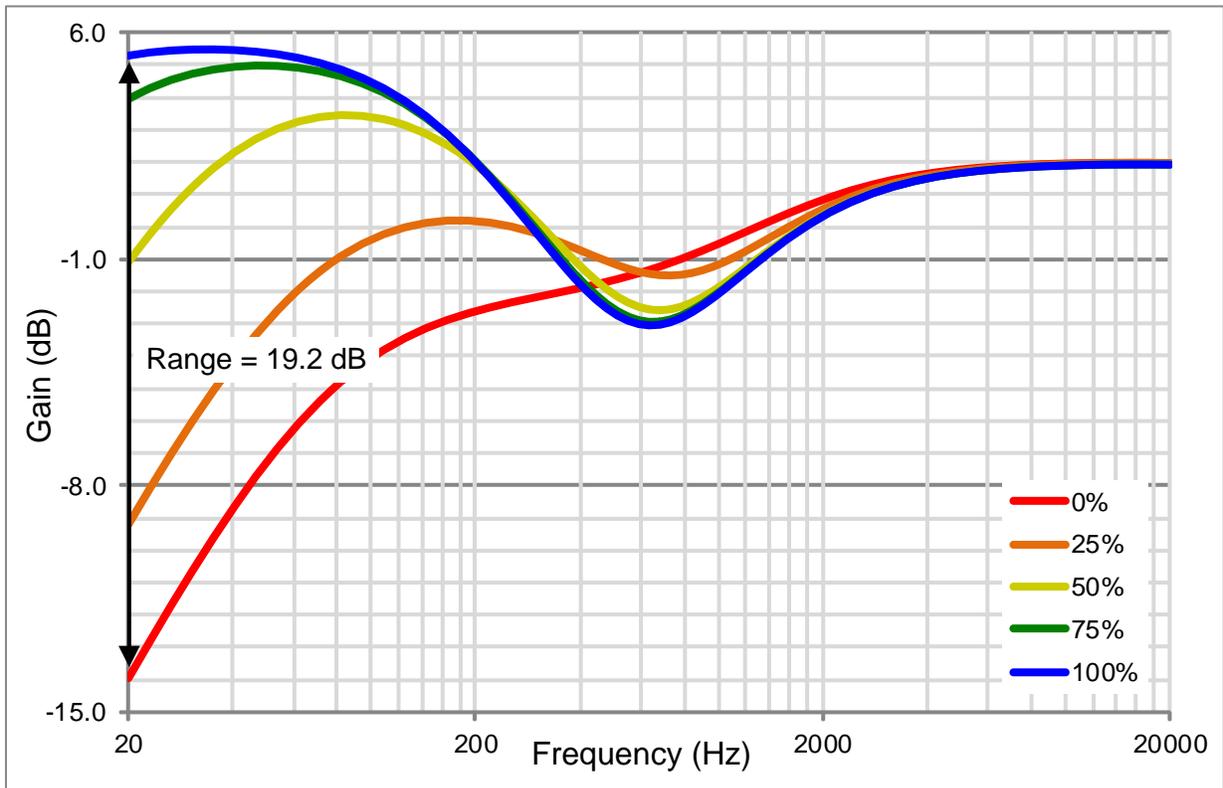
Figure 23: Measured AC Gain Analysis – Bass Control

In this condition, the bass gain varies from -13.9 dB at 0% potentiometer rotation to +5.3 dB at 100% potentiometer rotation. This gives an adjustment range of 19.2 dB, which meets the design requirement of 15 dB.

The result of the measured ac phase analysis under the same conditions is shown in section A.3.

6.4 THD+N Performance

The result of the THD+N measurement over frequency with gain = 6 dB, all potentiometers set to 100%, and mid boost off is shown in Figure 24. The audio analyzer is set to a measurement bandwidth of 90 kHz and no additional filtering or weighting is applied.
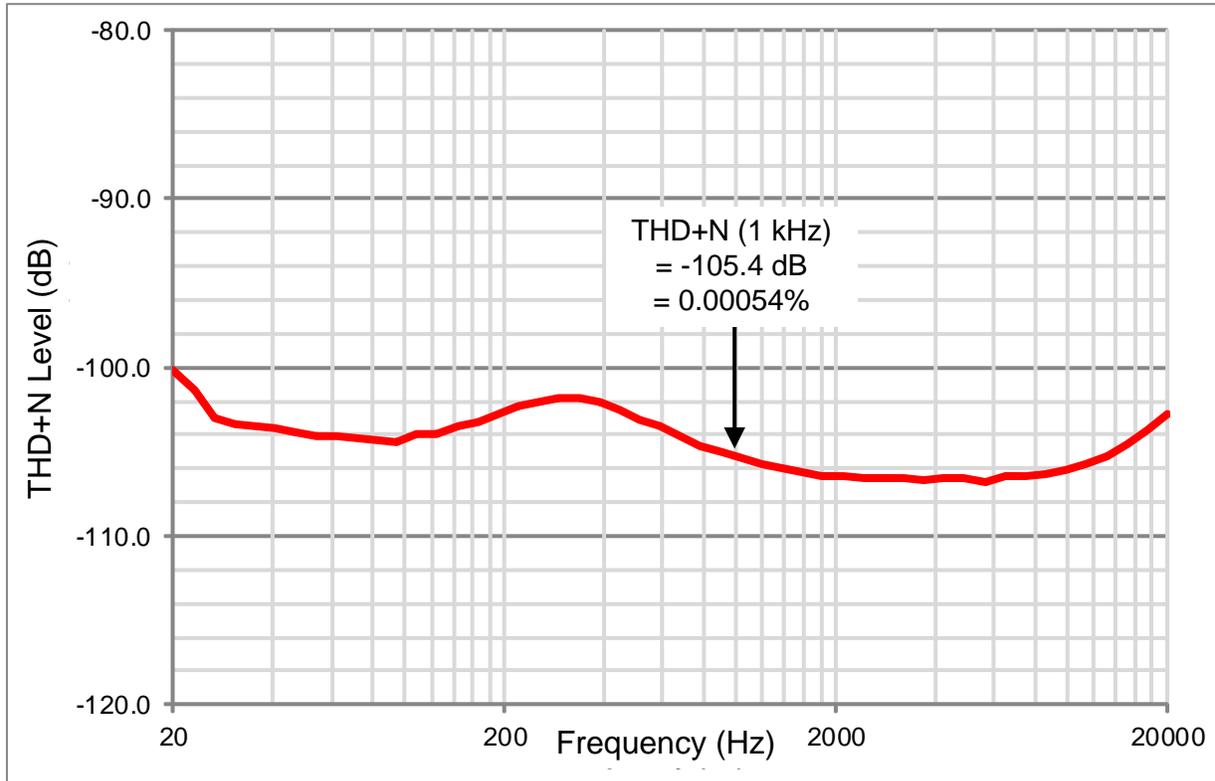
Figure 24: Measured THD+N Level vs. Frequency

This THD+N level at 1 kHz is measured to be -105.4 dB (0.00054%), which meets the design requirement of -100 dB. The THD+N levels at 20 Hz and 20 kHz are measured to be -100.1 dB (0.00099%) and -102.8 dB (0.00072%), respectively.

The Fast Fourier transform (FFT) measurement with V_{IN} at 1 kHz, gain = 6 dB, all potentiometers set to 100%, and mid boost off is shown in Figure 25. The FFT measurement is set to 192k points and 4x averaging.
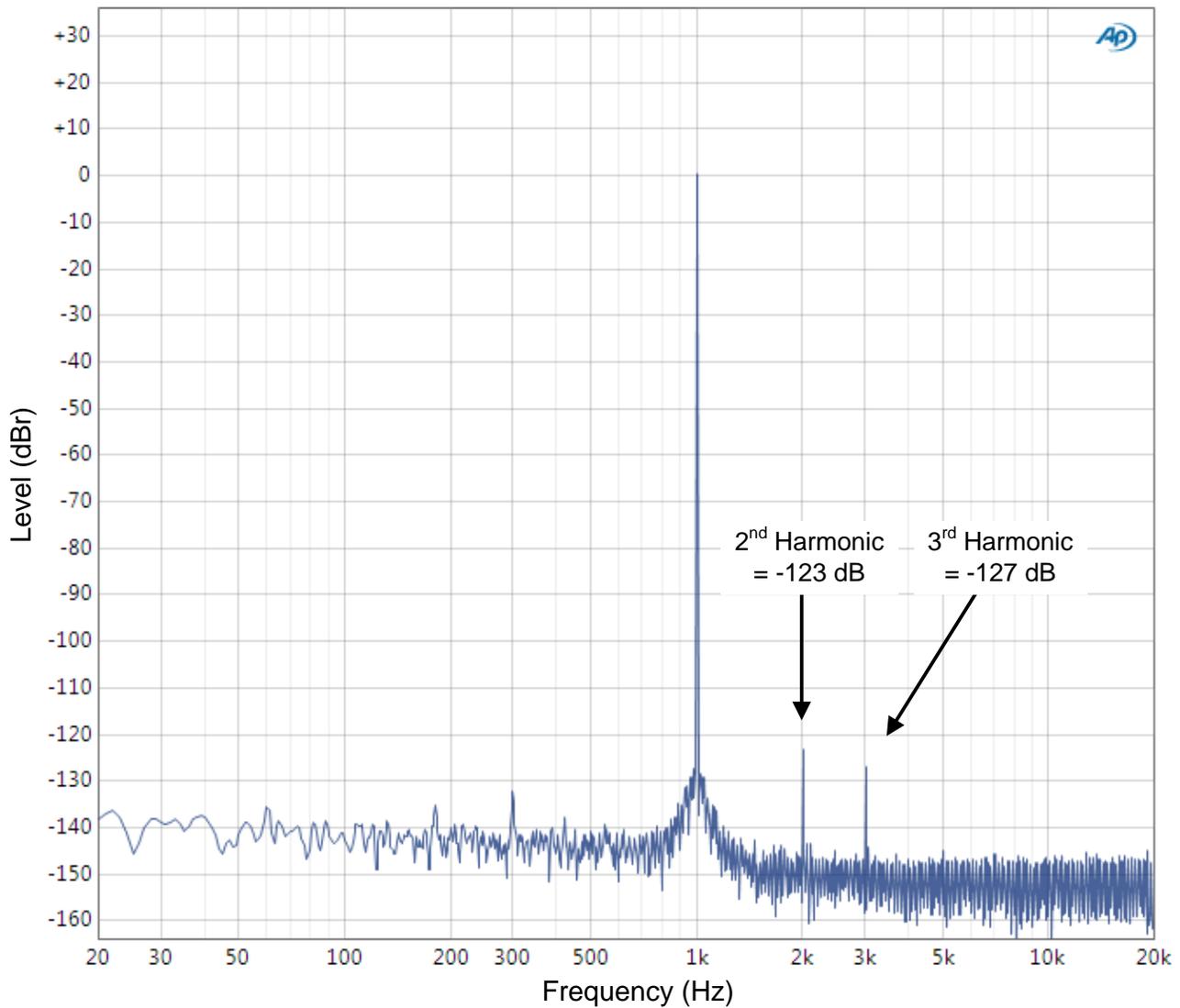
Figure 25: Measured Fast Fourier Transform (FFT)

The y-axis is referenced to the fundamental frequency output level of 1.58 V_{RMS}. The second harmonic is measured at 123 dB below the fundamental, while the third harmonic is measured at 127 dB below the fundamental.

6.5 Measured Results Summary

Table 4 summarizes the measured performance of the design.

Table 4. Comparison of Design Goals and Measured Performance

	Goal	Measured
THD+N ratio at 1 kHz	-100 dB (0.001%)	-105.4 dB (0.00054%)
Treble adjustment range	10 dB	10.6 dB
Mid adjustment range	6 dB	8.9 dB
Bass adjustment range	15 dB	19.2 dB

7 Audio Recordings

While frequency response curves and FFTs can be useful in measuring the performance of a circuit, in audio applications many times "hearing is believing." The following audio recordings capture the tonal differences between circuit settings as the author plays an E major chord on a Gibson SG through its neck pickup.

Listen online here.

7.1 Audio Recording Downloads

1. All controls 100% vs. all controls 0%: Download

2. Bass 100% vs. bass 0% (mid and treble 50%): Download

3. Mid 100% vs. mid 0% (bass and treble 50%): Download

4. Treble 100% vs. treble 0% (bass and mid 50%): Download

5. Mid boost off vs. mid boost on (all controls 50%): Download

8 Modifications

The components selected for this design were based on the design goals outlined at the beginning of the design process.

This design specifies an input impedance of 6 kΩ. While this is a reasonable specification for passive electric guitar pickups, the actual value will vary across electric guitar and pickup manufacturers. It may be necessary to adjust the values of R_4, C_4 and C_7 in the input filter to achieve the desired cutoff frequencies.

If modifications to the frequency response of the tone stack are desired, the component values of the FMV tone stack may easily be modified using the equations given in section 2.3.2. Duncan's Tone Stack Calculator is a free software tool which may also be used to model the response of different tone stack topologies, component values and potentiometer settings [8].

A JFET-input amplifier was selected for this application because of the high impedances present in the circuit. The extremely low input bias current (I_B) and input current noise (I_n) of FET-input devices prevent large offset and noise voltages from developing and degrading audio performance.

Among the FET-input audio amplifiers offered by Texas Instruments, the OPA1642 was selected for this application because of its extremely stable input common-mode capacitance which preserves excellent distortion performance even with high source impedances. The OPA1652 is another FET-input audio amplifier with excellent THD+N performance, low noise, and low cost; however its in-circuit distortion performance may be reduced compared to the OPA1642. Table 5 summarizes the key specs between these two devices.

Table 5. Brief Comparison of Audio Operational Amplifiers

Operational Amplifier	THD+N Level at 1kHz	e_N at 1 kHz	I_Q / Channel	Input Type	Approx. Cost / Channel
OPA1642	-126 dB (0.00005%)	5.1 nV/√Hz	1.8 mA	JFET	$0.70 / 1ku
OPA1652	-126 dB (0.00005%)	4.5 nV/√Hz	2.5 mA	CMOS	$0.33 / 1ku

It is often desirable for a gain or volume control to have a response which is linear-in-dB with respect to the rotation of the controlling potentiometer. This results in a very natural change in perceived volume as the user rotates the volume knob.

Because the gain control in this circuit is a potentiometer with a linear taper, the characteristic is not linear-in-dB. If linear-in-dB behavior is desired a potentiometer with an audio taper may be used, or the gain control circuit can be modified to include a Baxandall active volume control as described in TIPD136 [9].

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9 About the Author

Ian Williams (ian@ti.com) is an applications engineer in the Precision Analog – Linear team at Texas Instruments where he supports industrial products and applications. Ian graduated from the University of Texas, Dallas, where he earned a Bachelor of Science in Electrical Engineering with a concentration in Microelectronics.

10 Acknowledgements & References

10.1 Acknowledgements

The author wishes to acknowledge John Caldwell for his assistance in the completion of this design.

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9. I. Williams (2013, December). Active Volume Control for Professional Audio. Available: http://www.ti.com/lit/ug/tidu034/tidu034.pdf

Appendix A.

A.1 Electrical Schematic
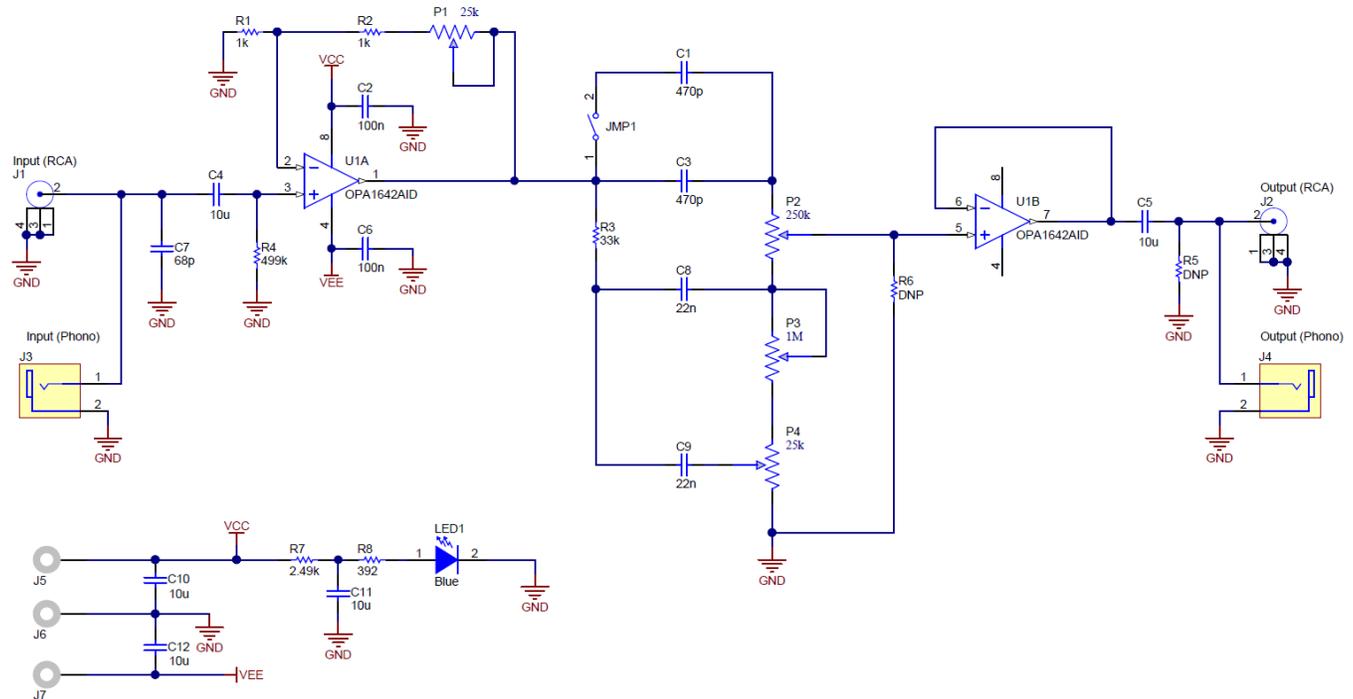
Figure A-1: Electrical Schematic

A.2 Bill of Materials

Item #	Quantity	Value	Designator	Description	Manufacturer	Part Number	Supplier	Supplier Part Number
1	2	470p	C1, C3	CAP CER 470PF 50V 5% NP0 0603	Samsung Electro-Mechanics America	CL10C471JB8NCNC	Digi-Key	1276-2302-1-ND
2	2	100n	C2, C6	CAP CER 0.1UF 50V 10% X7R 0603	Samsung Electro-Mechanics America	CL10B104KB8SFNC	Digi-Key	1276-1936-1-ND
3	5	10u	C4, C5, C10-C12	CAP CER 10UF 35V 10% X7R 1206	Samsung Electro-Mechanics America	CL31B106KLHNNNE	Digi-Key	1276-3103-1-ND
4	1	68p	C7	CAP CER 68PF 50V 5% NP0 0603	Samsung Electro-Mechanics America	CL10C680JB8NCNC	Digi-Key	1276-2324-1-ND
5	2	22n	C8, C9	CAP CER 0.022UF 50V 5% NP0 1206	TDK Corporation	C3216C0G1H223J060AA	Digi-Key	445-7690-1-ND
6	2	1k	R1, R2	RES SMD 1K OHM 1% 1/10W 0603	Panasonic Electronic Components	ERJ-3EKF1001V	Digi-Key	P1.00KHCT-ND
7	1	33k	R3	RES SMD 33K OHM 1% 1/10W 0603	Panasonic Electronic Components	ERJ-3EKF3302V	Digi-Key	P33.0KHCT-ND
8	1	499k	R4	RES SMD 499K OHM 1% 1/10W 0603	Panasonic Electronic Components	ERJ-3EKF4993V	Digi-Key	P499KHCT-ND
9	1	2.49k	R7	RES SMD 2.49K OHM 1% 1/8W 0805	Panasonic Electronic Components	ERJ-6ENF2491V	Digi-Key	P2.49KCCT-ND
10	1	392	R8	RES SMD 392 OHM 1% 1/8W 0805	Panasonic Electronic Components	ERJ-6ENF3920V	Digi-Key	P392CCT-ND
11	2	25k	P1, P4	25K LINEAR POT RIGHT ANGLE PC MOUNT	Alpha	RV16AF-41-15R1-B25K	Mammoth Electronics	210-100-B-25K
12	1	250k	P2	250K LINEAR POT RIGHT ANGLE PC MOUNT	Alpha	RV16AF-41-15R1-B250K	Mammoth Electronics	210-100-B-250K
13	1	1M	P3	1M LOG POT RIGHT ANGLE PC MOUNT	Alpha	RV16AF-41-15R1-A1M	Mammoth Electronics	210-100-A-1M
14	3	KNOB	N/A	BLK ALUM KNURLED KNOB 15MM X 13MM	Unknown	4SKA-15X13KBK	Mammoth Electronics	4SKA-15X13KBK
15	1	LED	LED1	LED BLUE CLEAR 3MM 470NM	KingBright	WP710A10QBC/D	Digi-Key	754-1596-ND
16	1	OPA1642	U1	OPA1642 SOUND-PLUS OP AMP	Texas Instruments	OPA1642AID	N/A	N/A
17	3	BANANA	J5-J7	CONN JACK BANANA UNINS PANEL MOU	Emerson Network Power	108-0740-001	Digi-Key	J147-ND
18	2	RCA	J1, J2	CONN RCA JACK METAL R/A BLK PCB	CUI Inc	RCJ-011	Digi-Key	CP-1400-ND
19	2	PHONO PLUG	J3, J4	CONN JACK PHONE 1/4" 2POS RA OPN	Switchcraft Inc.	RN111PC	Digi-Key	SC1121-ND
20	1	JUMPER	JMP1	CONN HEADER 2POS .100" SGL GOLD	Samtec	TSW-102-07-G-S	Digi-Key	SAM1029-02-ND
21	1	JUMPER SHUNT	JMP1	JUMPER SHORTING UNITS	TE Connectivity	881545-2	Digi-Key	A26242-ND
22	4	SCREW	Screws	SCREW MACHINE PHILLIPS 4-40X3/8	B&F Fastener Supply	PMS 440 0038 PH	Digi-Key	H781-ND
23	4	STANDOFF	H1-H4	STANDOFF HEX 4-40THR ALUM .500"L	Keystone Electronics	2203	Digi-Key	2203K-ND
24	1	DNP	R5, R6	DO NOT POPULATE	n/a	N/A	N/A	N/A

Figure A-2: Bill of Materials

A.3 Phase Response Measurements
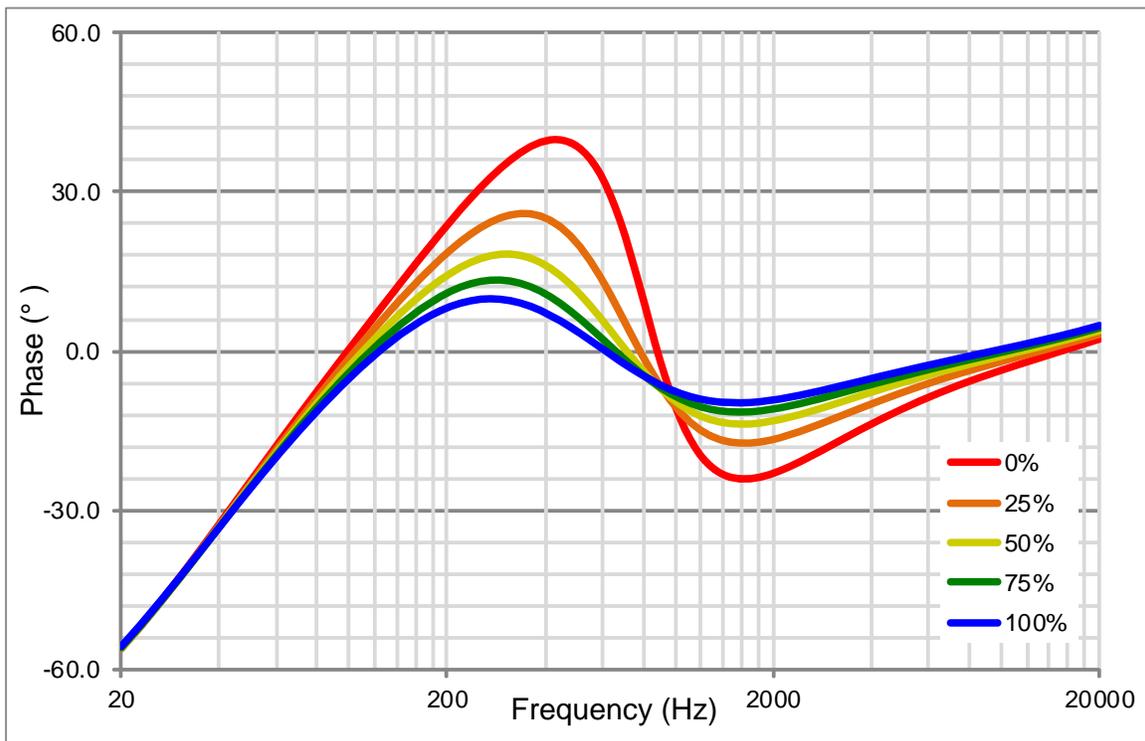
Figure A-3: Measured AC Phase Analysis – Treble Control
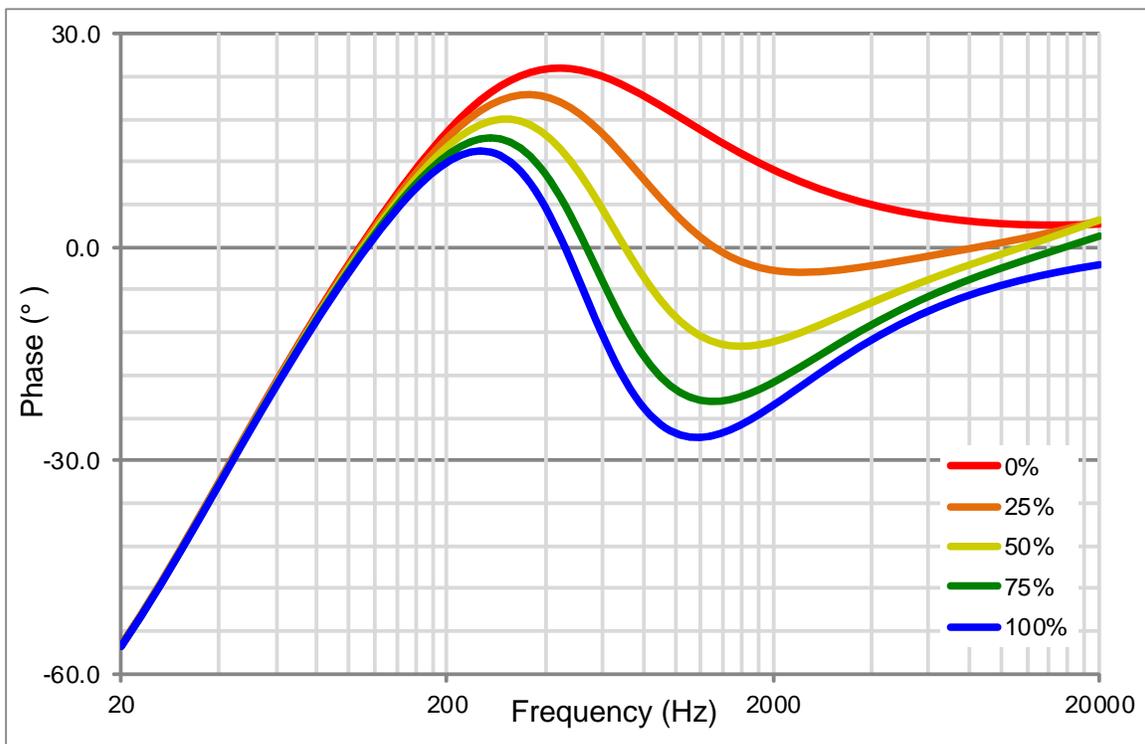
Figure A-4: Measured AC Phase Analysis – Mid Control

TIPD186 - Tone Stack for Guitar Amplifier Reference Design 33
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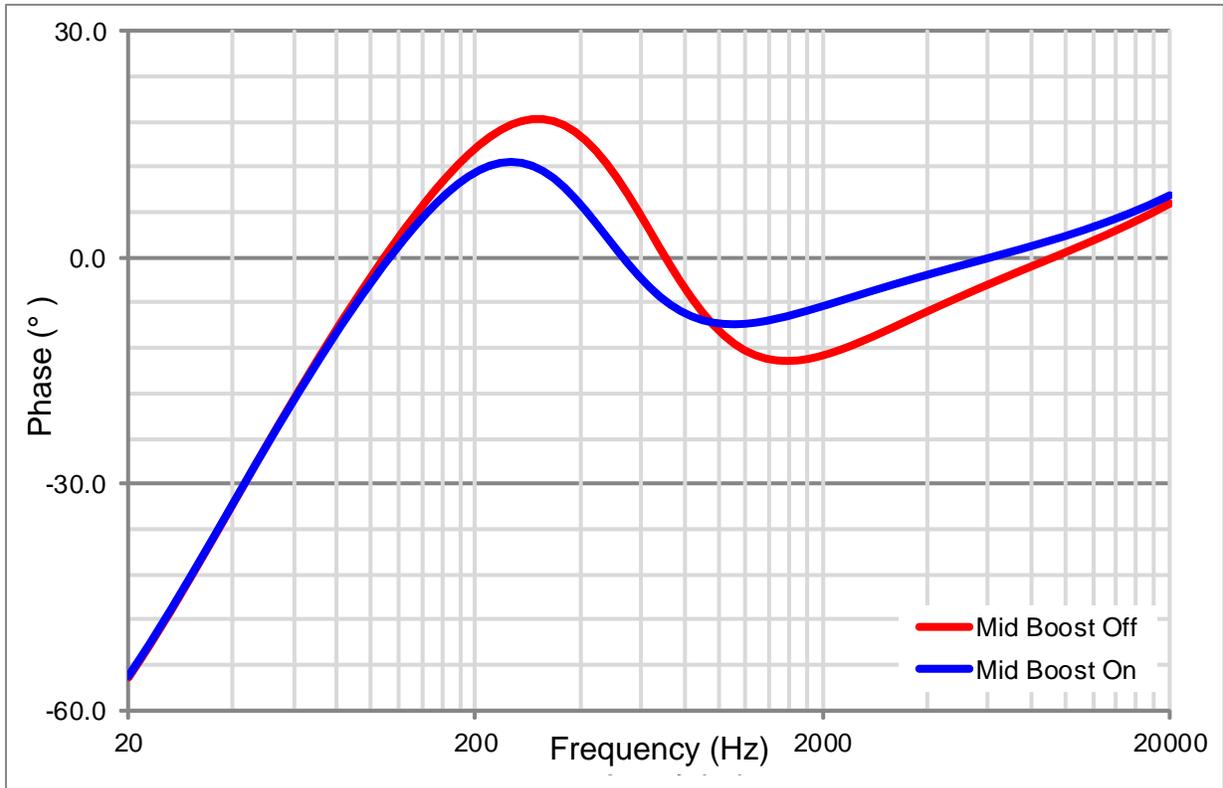
Figure A-5: Measured AC Phase Analysis – Mid Boost
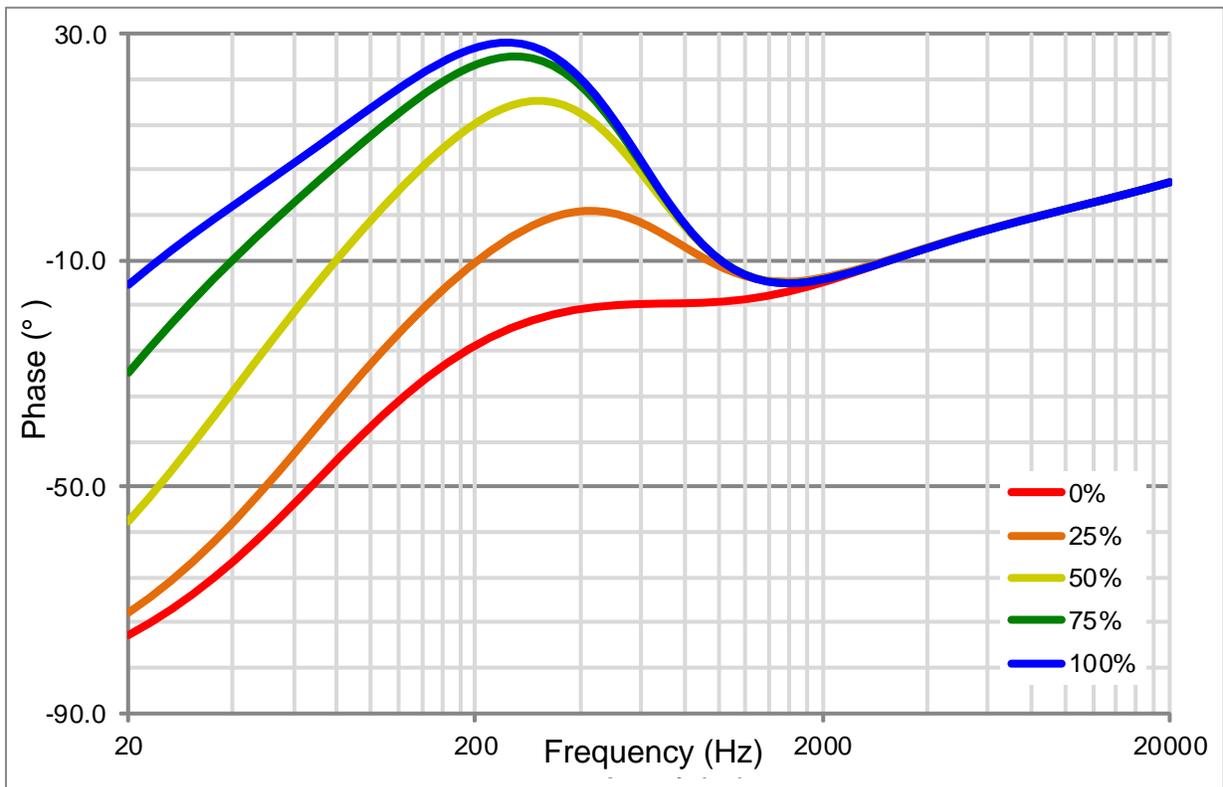
Figure A-6: Measured AC Phase Analysis – Bass Control