design each half of the LM387A is wired as a non-inverting amplifier with bias and gain setting resistors as before. Resistors R_1 and R_2 set the input impedance at $2k\Omega$ (balanced). Potentiometer R_7 is used to set the output DC level at zero volts by matching the DC levels of pins 4 and 5 of the LM387A.

This allows direct coupling between the stages, thus eliminating the coupling capacitors and the associated matching problem for optimum CMRR. AC gain resistors R_8 and R_9 are grounded by the common capacitor, C_3, eliminating another capacitor and assuring AC gain match. Close resistor tolerance is necessary around the LM387A in order to preserve common-mode signals appearing at the input. The function of the LM387A is to amplify the low level signal adding as little noise as possible, and leave common-mode rejection to the LF356.

By substituting a LM381A and increasing its current density (see Section 2.7) a professional quality transformerless balanced mic preamp can be designed. With the exception of the additional components necessary to increase the current density, the circuit is the same as Figure 2.13.4. The improvement in noise performance is 7dB, yielding noise −74dB below a 2mV input level.

REFERENCES

1. Smith, D. A. and Wittman, P. H., "Design Considerations of Low-Noise Audio Input Circuitry for a Professional Microphone Mixer," *Jour. Aud. Eng. Soc.*, vol. 18, no. 2, April 1970, pp. 140-156.

2.14 TONE CONTROLS

2.14.1 Introduction

There are many reasons why a user of audio equipment may wish to alter the frequency response of the material being played. The purist will argue that he wants his amplifier "flat," i.e., no alteration of the source material's frequency response; hence, amplifiers with tone controls often have a FLAT position or a switch which bypasses the circuitry. The realist will argue that he wants the music to reach his ears "flat." This position recognizes that such parameters as room acoustics, speaker response, etc., affect the output of the amplifier and it becomes necessary to compensate for these effects if the listener is to "hear" the music "flat," i.e., as recorded. And there is simply the matter of personal taste (which is not simple): one person prefers "bassy" music; another prefers it "trebley."

2.14.2 Passive Design

Passive tone controls offer the advantages of lowest cost and minimum parts count while suffering from severe insertion loss which often creates the need for a tone recovery amplifier. The insertion loss is approximately equal to the amount of available boost, e.g., if the controls have +20dB of boost, then they will have about −20dB insertion loss. This is because passive tone controls work as AC voltage dividers and really only cut the signal.

2.14.3 Bass Control

The most popular bass control appears as Figure 2.14.1 along with its associated frequency response curve. The curve shown is the ideal case and can only be approximated. The corner frequencies f_1 and f_2 denote the half-power points and therefore represent the frequencies at which the relative magnitude of the signal has been reduced (or increased) by 3dB.

Passive tone controls require "audio taper" (logarithmic) potentiometers, i.e., at the 50% rotation point the slider splits the resistive element into two portions equal to 90% and 10% of the total value. This is represented in the figures by "0.9" and "0.1" about the wiper arm.

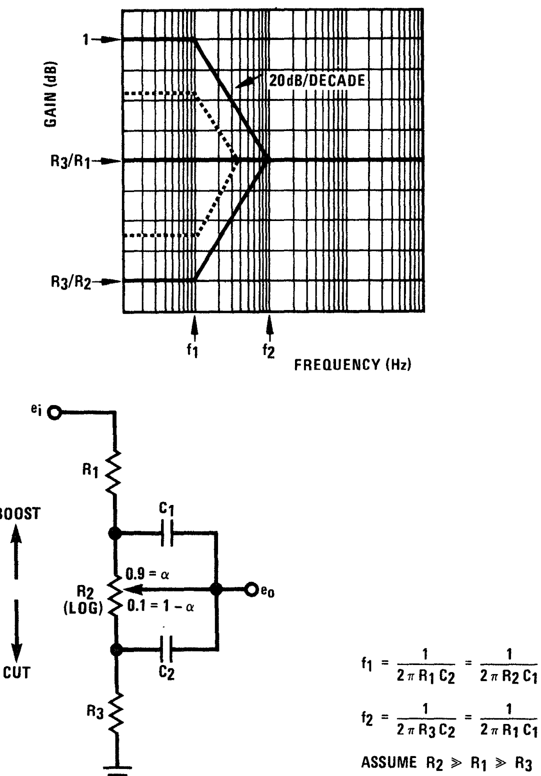


FIGURE 2.14.1 Bass Tone Control — General Circuit

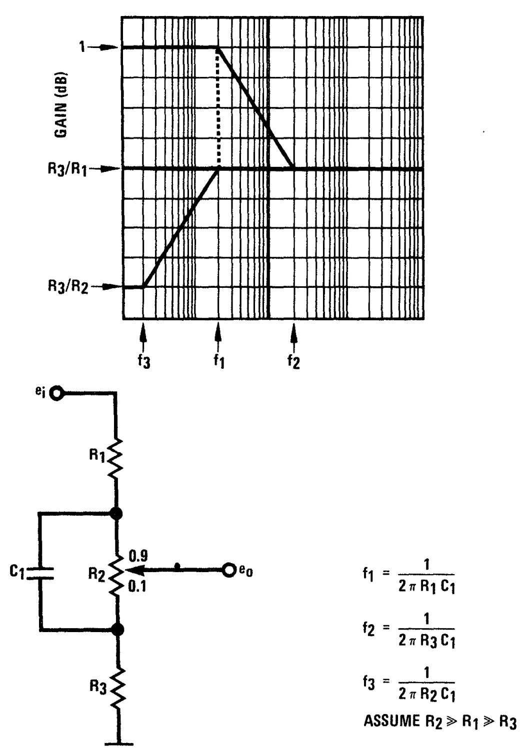


FIGURE 2.14.2 Minimum-Parts Bass Tone Control

For designs satisfying $R_2 \gg R_1 \gg R_3$, the amount of available boost or cut of the signal given by Figure 2.14.1 is set by the following component ratios:

$$\frac{R_1}{R_2} = \frac{R_3}{R_1} = \frac{C_1}{C_2} = \text{bass boost or cut amount} \qquad (2.14.1)$$

The turnover frequency f_2 occurs when the reactance of C_1 equals R_1 and the reactance of C_2 equals R_3 (assuming $R_2 \gg R_1 \gg R_3$):

$$C_1 = \frac{1}{2\pi f_2 R_1} \qquad (2.14.2)$$

$$C_2 = \frac{1}{2\pi f_2 R_3} \qquad (2.14.3)$$

The frequency response will be accentuated or attenuated at the rate of ±20dB/decade = ±6dB/octave (single pole response) until f_1 is reached. This occurs when the limiting impedance is dominant, i.e., when the reactance of C_1 equals R_2 and the reactance of C_2 equals R_1:

$$f_1 = \frac{1}{2\pi R_1 C_2} = \frac{1}{2\pi R_2 C_1} \qquad (2.14.4)$$

Note that Equations (2.14.1)-(2.14.4) are not independent but all relate to each other and that selection of boost/cut amount and corner frequency f_2 fixes the reamining parameters. Also of passing interest is the fact that f_2 is dependent upon the wiper position of R_2. The solid-line response of Figure 2.14.1 is only valid at the extreme ends of potentiometer R_2; at other positions the response changes as depicted by the dotted line response. The relevant time constants involved are $(1 - \alpha)R_2C_1$ and αR_2C_2, where α equals the fractional rotation of the wiper as shown in Figure 2.14.1. While this effect might appear to be undesirable, in practice it is quite acceptable and this design continues to dominate all others.

Figure 2.14.2 shows an alternate approach to bass tone control which offers the cost advantage of one less capacitor and the disadvantage of asymmetric boost and cut response. The degree of boost or cut is set by the same resistor ratios as in Figure 2.14.1.

$$\frac{R_2}{R_1} = \frac{R_1}{R_3} = \text{bass boost or cut amount} \qquad (2.14.5)$$

assumes $R_2 \gg R_1 \gg R_3$

The boost turnover frequency f_2 occurs when the reactance of C_1 equals R_3:

$$C_1 = \frac{1}{2\pi f_2 R_3} \qquad (2.14.6)$$

Maximum boost occurs at f_1, which also equals the cut turnover frequency. This occurs when the reactance of C_1 equals R_1, and maximum cut is achieved where $X_{C_1} = R_2$. Again, all relevant frequencies and the degree of boost or cut are related and interact. Since in practice most tone controls are used in their boost mode, Figure 2.14.2 is not as troublesome as it may first appear.

2.14.4 Treble Control

The treble control of Figure 2.14.3 represents the electrical analogue of Figure 2.14.1, i.e., resistors and capacitors interchanged, and gives analogous performance. The amount of boost or cut is set by the following ratios:

$$\frac{R_3}{R_1} = \frac{C_1}{C_2} = \text{treble boost or cut amount} \qquad (2.14.7)$$

assumes $R_2 \gg R_1 \gg R_3$

Treble turnover frequency f_1 occurs when the reactance of C_1 equals R_1 and the reactance of C_2 equals R_3:

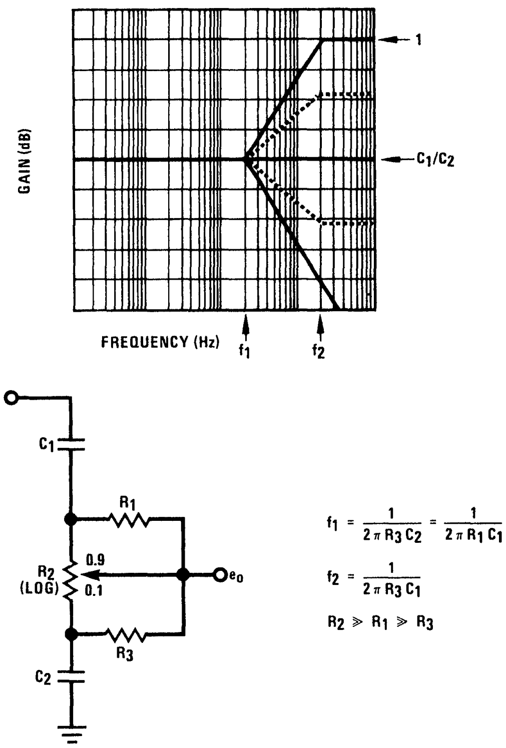


FIGURE 2.14.3 Treble Tone Control — General Circuit

$$C_1 = \frac{1}{2\pi f_1 R_1} \qquad (2.14.8)$$

$$C_2 = \frac{1}{2\pi f_1 R_3} \qquad (2.14.9)$$

The amount of available boost is reached at frequency f_2 and is determined when the reactance of C_1 equals R_3.

$$f_2 = \frac{1}{2\pi R_3 C_1} \qquad (2.14.10)$$

In order for Equations (2.14.8) and (2.14.9) to remain valid, it is necessary for R_2 to be designed such that it is much larger than either R_1 or R_3. For designs that will not permit this condition, Equations (2.14.8) and (2.14.9) must be modified by replacing the R_1 and R_3 terms with $R_1 \| R_2$ and $R_3 \| R_2$ respectively. Unlike the bass control, f_1 is not dependent upon the wiper position of R_2, as indicated by the dotted lines shown in Figure 2.14.3. Note that in the full cut position attenuation tends toward zero without the shelf effect of the boost characteristic.

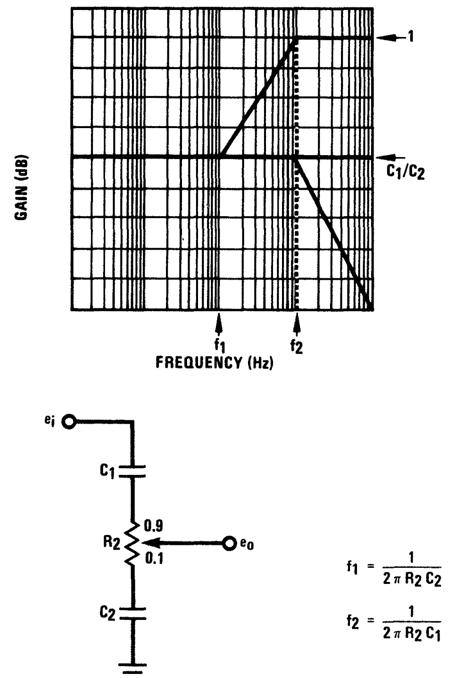


FIGURE 2.14.4 Minimum-Parts Treble Tone Control

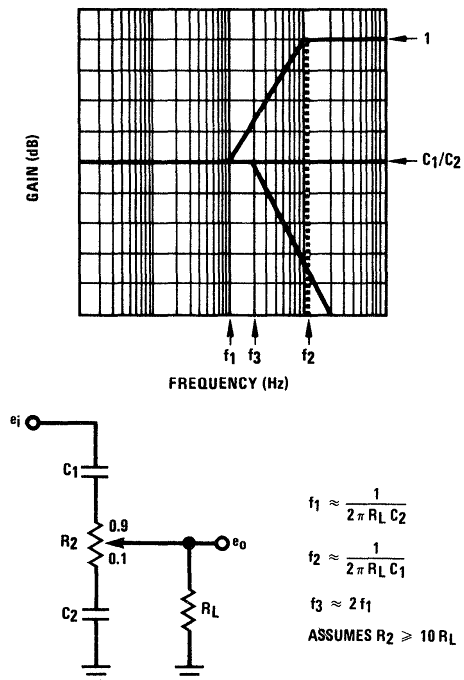


FIGURE 2.14.5 Effect of Loading Treble Tone Control

It is possible to omit R_1 and R_3 for low cost systems. Figure 2.14.4 shows this design with the modified equations and frequency response curve. The obvious drawback appears to be that the turnover frequency for treble cut occurs a decade later (for ±20dB designs) than the boost point. As noted previously, most controls are used in their boost mode, which lessens this drawback, but probably more important is the effect of finite loads on the wiper of R_2.

Figure 2.14.5 shows the loading effect of R_L upon the frequency response of Figure 2.14.4. Examination of these two figures shows that the presence of low impedance (relative to R_2) on the slider changes the break points significantly. If R_L is 1/10 of R_2 then the break points shift a full decade higher. The equations given in Figure 2.14.5 hold for values of $R_2 \geqslant 10\,R_L$. A distinct advantage of Figure 2.14.5 over Figure 2.14.4 is seen in the cut performance. R_L tends to pull the cut turnover frequency back toward the boost corner — a nice feature, and with two fewer resistors. Design becomes straightforward once R_L is known. C_1 and C_2 are calculated from Equations (2.14.11) and (2.14.12).

$$C_1 = \frac{1}{2\pi f_2 R_L} \qquad (2.14.11)$$

$$C_2 = \frac{1}{2\pi f_1 R_L} \qquad (2.14.12)$$

Here again, gain and turnover frequencies are related and fixed by each other.

Example 2.14.1

Design a passive, symmetrical bass and treble tone control circuit having 20dB boost and cut at 50Hz and 10kHz, relative to midband gain.

Solution

1. For symmetrical controls, combine Figures 2.14.1 and 2.14.3.

BASS (Figure 2.14.1):

2. From Equation (2.14.1):

$$\frac{R_1}{R_2} = \frac{R_3}{R_1} = \frac{C_1}{C_2} = \frac{1}{10}\,(-20\text{dB})$$

$$f_1 = 50\,\text{Hz and } f_2 = 500\,\text{Hz}$$

3. Let $R_2 = 100\text{k}$ (audio taper).

4. From Step 2:

$$R_1 = \frac{R_2}{10} = \frac{100\text{k}}{10} = 10\text{k}$$

$$R_3 = \frac{R_1}{10} = \frac{10\text{k}}{10} = 1\text{k}$$

5. From Equation (2.14.2) and Step 2:

$$C_1 = \frac{1}{2\pi f_2 R_1} = \frac{1}{(2\pi)(500)(10\text{k})} = 3.18 \times 10^{-8}$$

Use $C_1 = 0.033\mu F$

$C_2 = 10C_1$

$C_2 = 0.33\mu F$

TREBLE (Figure 2.14.3):

6. From Equation (2.14.7):

$$\frac{R_3}{R_1} = \frac{C_1}{C_2} = \frac{1}{10} \quad (-20\,dB)$$

$$f_1 = 1\,kHz, f_2 = 10\,kHz$$

7. Let $R_2 = 100k$ (audio taper).
8. Select $R_1 = 10k$ (satisfying $R_2 \gg R_1$ and minimizing component spread).

   Then:

$$R_3 = \frac{R_1}{10} = \frac{10k}{10} = 1k$$

9. From Equation (2.14.8) and Step 6:

$$C_1 = \frac{1}{2\pi f_1 R_1} = \frac{1}{(2\pi)(1k)(10k)} = 1.59 \times 10^{-8}$$

   Use $C_1 = 0.015\mu F$

$$C_2 = 10C_1$$

$$C_2 = 0.15\mu F$$

The completed design appears as Figure 2.14.6, where R_I has been included to isolate the two control circuits, and C_O is provided to block all DC voltages from the circuit – insuring the controls are not "scratchy," which results from DC charge currents in the capacitors and on the sliders. C_O is selected to agree with system low frequency response:

$$C_O = \frac{1}{(2\pi)(20\,Hz)(10k + 100k + 1k)} = 7.17 \times 10^{-8}$$

Use $C_O = 0.1\mu F$

2.14.5 Use of Passive Tone Controls with LM387 Preamp

A typical application of passive tone controls (Figure 2.14.7) involves a discrete transistor used following the circuit to further amplify the signal as compensation for the loss through the passive circuitry. While this is an acceptable practice, a more judicious placement of the same transistor results in a superior design without increasing parts count or cost.

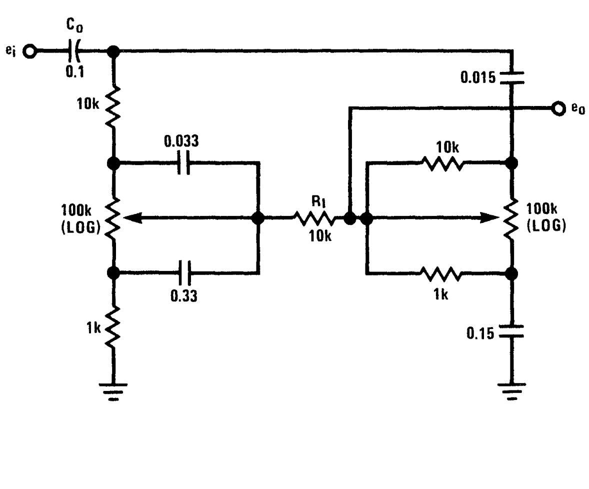
Placing the transistor *ahead* of the LM387 phono or tape preamplifier (Figure 2.14.8) improves the S/N ratio by boosting the signal before equalizing. An improvement of at least 3dB can be expected (analogous to operating a LM381A with single-ended biasing). The transistor selected must be low-noise, but in quantity the difference in price becomes negligible. The only precaution necessary is to allow sufficient headroom in each stage to minimize transient clipping. However, due to the excellent open-loop gain and large output swing capability of the LM387, this is not difficult to achieve.

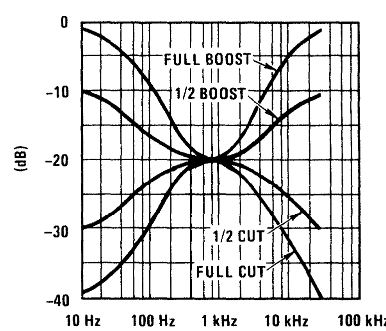
An alternative to the transistor is to use an LM381A selected low-noise preamp. Superior noise performance is possible. (See Section 2.7.) The large gain and output swing are adequate enough to allow sufficient single-stage gain to overcome the loss of the tone controls. Figure 2.14.9 shows an application of this concept where the LM381A is used differentially. Single-ended biasing and increased current density may be used for even quieter noise voltage performance.

2.14.6 Loudness Control

A loudness control circuit compensates for the logarithmic nature of the human ear. Fletcher and Munson[1] published curves (Figure 2.14.10) demonstrating this effect. Without loudness correction, the listening experience is characterized by a pronounced loss of bass response accompanied by a slight loss of treble response as the volume level is decreased. Compensation consists of boosting the high and

2





Bass & Treble Tone Control Response

FIGURE 2.14.6 Complete Passive Bass & Treble Tone Control

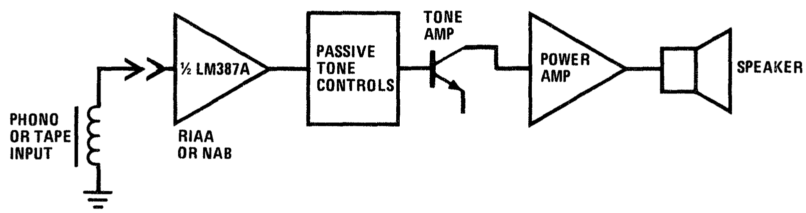


FIGURE 2.14.7 Typical Passive Tone Control Application

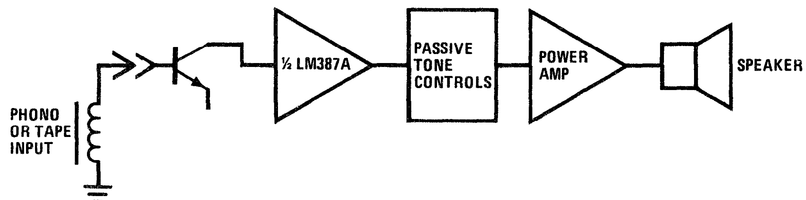


FIGURE 2.14.8 Improved Circuit Using Passive Tone Controls

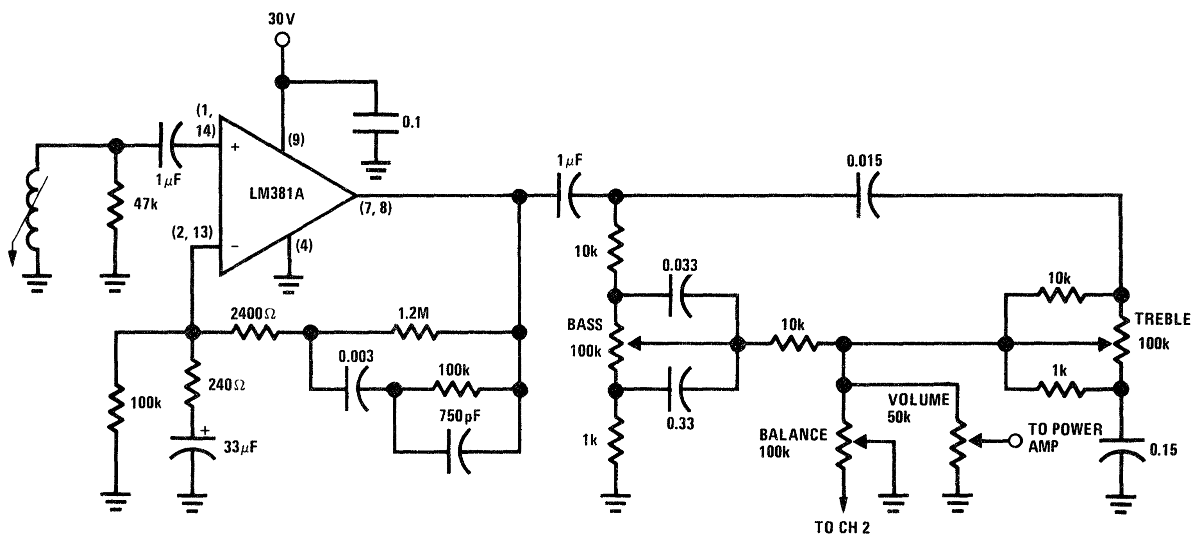


FIGURE 2.14.9 Single Channel of Complete Phono Preamp

low ends of the audio frequency band as an inverse function of volume control setting. One commonly used circuit appears as Figure 2.14.11 and uses a tapped volume pot (tap @ 10% resistance). The switchable R-C network paralleling the pot produces the frequency response shown in Figure 2.14.12 when the wiper is positioned at the tap point (i.e., mid-position for audio taper pot). As the wiper is moved further away from the tap point (louder) the paralleling circuit has less and less effect, resulting in a volume sensitive compensation scheme.

2.14.7 Active Design

Active tone control circuits offer many attractive advantages: they are inherently symmetrical about the axis in boost and cut operation; they have very low THD due to being incorporated into the negative feedback loop of the gain block, as opposed to the relatively high THD exhibited by a tone recovery transistor; and the component spread, i.e., range of values, is low.

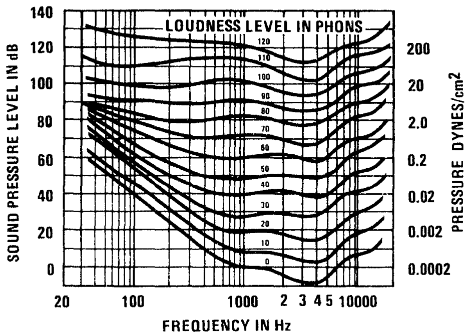


FIGURE 2.14.10 Fletcher-Munson Curves (USA). (Courtesy, Acoustical Society of America)

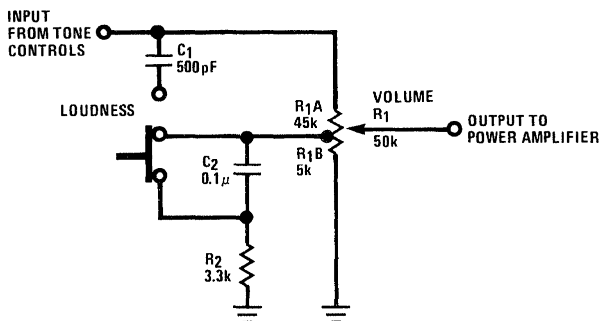


FIGURE 2.14.11 Loudness Control

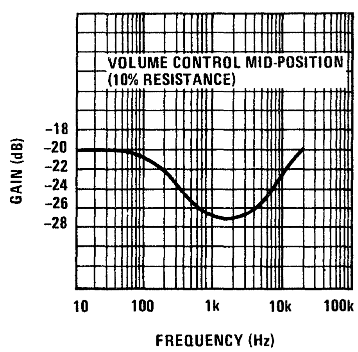


FIGURE 2.14.12 Loudness Control Frequency Response

The most common active tone control circuit is the so-called "Americanized" version of the Baxandall (1952)[2] negative feedback tone controls. A complete bass and treble active tone control circuit is given in Figure 2.14.13a. At very low frequencies the impedance of the capacitors is large enough that they may be considered open circuits, and the gain is controlled by the bass pot, being equal to Equations (2.14.13) and (2.14.14) at the extreme ends of travel.

$$A_{VB} = \frac{R_1 + R_2}{R_1} \quad \text{(max bass boost)} \qquad (2.14.13)$$

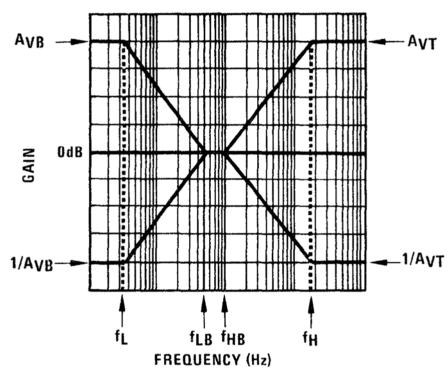
$$\frac{1}{A_{VB}} = \frac{R_1}{R_1 + R_2} \quad \text{(max bass cut)} \qquad (2.14.14)$$

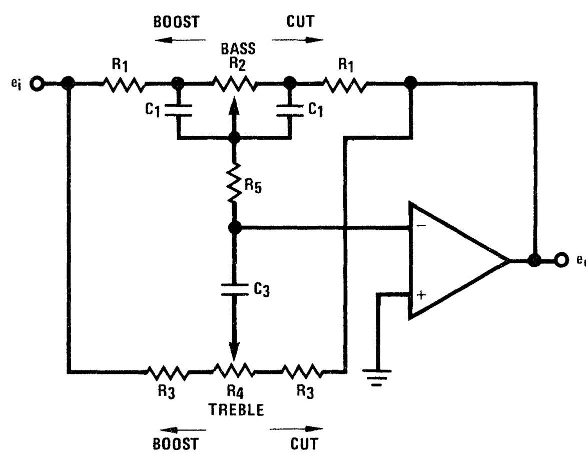
At very high frequencies the impedance of the capacitors is small enough that they may be considered short circuits, and the gain is controlled by the treble pot, being equal to Equations (2.14.15) and (2.14.16) at the extreme ends of travel.

$$A_{VT} = \frac{R_3 + R_1 + 2R_5}{R_3} \quad \text{(max treble boost)} \qquad (2.14.15)$$

$$\frac{1}{A_{VT}} = \frac{R_3}{R_3 + R_1 + 2R_5} \quad \text{(max treble cut)} \qquad (2.14.16)$$

Equations (2.14.15) and (2.14.16) are best understood by recognizing that the bass circuit at high frequencies forms a wye-connected load across the treble circuit. By doing a wye-delta transformation (see Appendix A3), the effective loading resistor is found to be (R_1 + $2R_5$) which is in parallel with (R_3 + R_4) and dominates the expression. (See Figure 2.14.13b.) This defines a constraint upon R_4 which is expressed as Equation (2.14.17).





BASS

$$f_L = \frac{1}{2\pi R_2 C_1}$$

$$f_{LB} = \frac{1}{2\pi R_1 C_1}$$

$$A_{VB} = 1 + \frac{R_2}{R_1}$$

ASSUMES $R_2 \gg R_1$

TREBLE

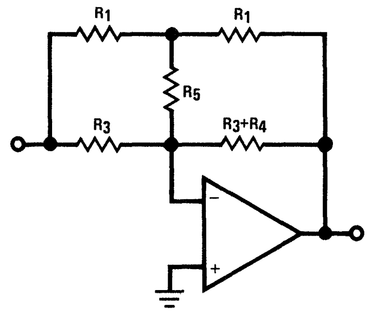
$$f_H = \frac{1}{2\pi R_3 C_3}$$

$$f_{HB} = \frac{1}{2\pi (R_1 + R_3 + 2R_5) C_3}$$

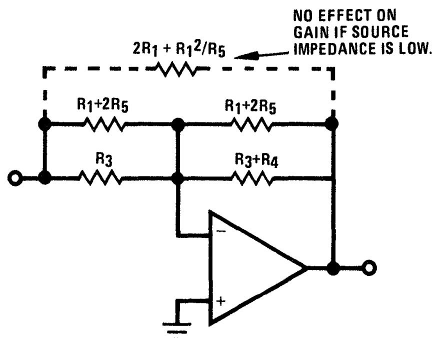
$$A_{VT} = 1 + \frac{R_1 + 2R_5}{R_3}$$

ASSUMES $R_4 \gg R_1 + R_3 + 2R_5$

FIGURE 2.14.13a Bass and Treble Active Tone Control



(a) High Frequency Max Treble Boost Equivalent Circuit



$$A_V = \frac{(R_1 + 2R_5) \| (R_3 + R_4)}{(R_1 + 2R_5) \| R_3} = \frac{R_3 + R_1 + 2R_5}{R_3}$$

IF $R_4 \gg R_1 + R_3 + 2R_5$

(b) High Frequency Circuit After Wye-Delta Transformation

FIGURE 2.14.13b Development of Max Treble Gain

$$R_4 \geqslant 10(R_3 + R_1 + 2R_5) \tag{2.14.17}$$

At low-to-middle frequencies the impedance of C_1 decreases at the rate of −6dB/octave, and is in parallel with R_2, so the effective resistance reduces correspondingly, thereby reducing the gain. This process continues until the resistance of R_1 becomes dominant and the gain levels off at unity.

The action of the treble circuit is similar and stops when the resistance of R_3 becomes dominant. The design equations follow directly from the above.

$$C_1 = \frac{1}{2\pi f_{LB} R_1} \quad \text{assumes } R_2 \gg R_1 \tag{2.14.18}$$

$$R_2 = \frac{1}{2\pi f_L C_1} \tag{2.14.19}$$

$$C_3 = \frac{1}{2\pi f_H R_3} \tag{2.14.20}$$

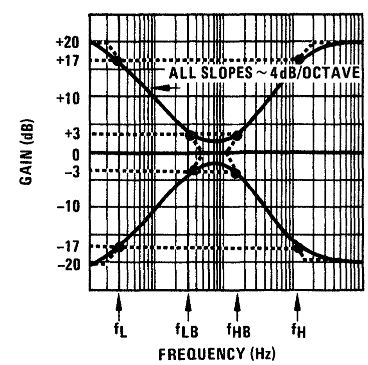
$$R_5 = \frac{1}{2}\left(\frac{1}{2\pi f_{HB} C_3} - R_1 - R_3\right) \tag{2.14.21}$$

The relationship between f_L and f_{LB} and between f_H and f_{HB} is not as clear as it may first appear. As used here these frequencies represent the ±3dB points relative to gain at midband and the extremes. To understand their relationship in the most common tone control design of ±20dB at extremes, reference is made to Figure 2.14.14. Here it is seen what shape the frequency response will actually have. Note that the flat (or midband) gain is not unity but approximately ±2dB. This is due to the close proximity of the poles and zeros of the transfer function. Another effect of this close proximity is that the slopes of the curves are not the expected ±6dB/octave, but actually are closer to ±4dB/octave. Knowing that f_L and f_{LB} are 14dB apart in magnitude, and the slope of the response is 4dB/octave, it is possible to relate the two. This relationship is given as Equation (2.14.22).

$$\frac{f_{LB}}{f_L} = \frac{f_H}{f_{HB}} \approx 10 \tag{2.14.22}$$

Example 2.14.2

Design a bass and treble active tone control circuit having ±20dB gain with low frequency upper 3dB corner at 30Hz and high frequency upper 3dB corner at 10kHz.



$$\frac{f_{LB}}{f_L} = \frac{f_H}{f_{HB}} \approx 10$$

FIGURE 2.14.14 Relationship Between Frequency Breakpoints of Active Tone Control Circuit

Solution

BASS DESIGN:

1. Select R_2 = 100k (linear). This is an arbitrary choice.
2. From Equation (2.14.13):

$$A_{VB} = 1 + \frac{R_2}{R_1} = 10 \text{ (+20dB)}$$

$$R_1 = \frac{R_2}{10 - 1} = \frac{100k}{9} = 1.11 \times 10^4$$

$$R_1 = 11k$$

3. Given f_L = 30Hz and from Equations (2.14.22) and (2.14.18):

$$f_{LB} = 10 f_L = 300\text{Hz}$$

$$C_1 = \frac{1}{2\pi f_{LB} R_1} = \frac{1}{(2\pi)(300)(11k)} = 4.82 \times 10^{-8}$$

$$C_1 = 0.05\mu F$$

TREBLE DESIGN:

4. Let $R_5 = R_1$ = 11k. This also is an arbitrary choice.

5. From Equation (2.14.15):

$$A_{VT} = 1 + \frac{R_1 + 2R_5}{R_3} = 10 \ (+20\text{dB})$$

$$R_3 = \frac{R_1 + 2R_5}{10 - 1} = \frac{11\text{k} + 2(11\text{k})}{9} = 3.67 \times 10^3$$

$$R_3 = 3.6\text{k}$$

6. Given $f_H = 10\text{kHz}$ and from Equation (2.14.20):

$$C_3 = \frac{1}{2\pi f_H R_3} = \frac{1}{(2\pi)(10\text{kHz})(3.6\text{k})} = 4.42 \times 10^{-9}$$

$$C_3 = 0.005\mu\text{F}$$

7. From Equation (2.14.17):

$$R_4 \geqslant 10(R_3 + R_1 + 2R_5)$$

$$\geqslant 10(3.6\text{k} + 11\text{k} + 22\text{k})$$

$$\geqslant 3.66 \times 10^5$$

$$R_4 = 500\text{k}$$

The completed design is shown in Figure 2.14.15, where the quad op amp LM349 has been chosen for the active element. The use of a quad makes for a single IC, stereo tone control circuit that is very compact and economical. The buffer amplifier is necessary to insure a low driving impedance for the tone control circuit and creates a high input impedance (100kΩ) for the source. The LM349 was chosen for its fast slew rate (2.5V/µs), allowing undistorted, full-swing performance out to > 25kHz. Measured THD was typically 0.05% @ 0dBm (0.77V) across the audio band. Resistors R_6 and R_7 were added to insure stability at unity gain since the LM349 is internally compensated for positive gains of five or greater. R_6 and R_7 act as input voltage dividers at high frequencies such that the actual input-to-output gain is never less than five (four if used inverting). Coupling capacitors C_4 and C_6 serve to block DC and establish low-frequency roll-off of the system; they may be omitted for direct-coupled designs.

2.14.8 Alternate Active Bass Control

Figure 2.14.16 shows an alternate design for bass control, offering the advantage of one less capacitor while retaining identical performance to that shown in Figure 2.14.13. The development of Figure 2.14.16 follows immediately from Figure 2.14.13 once it is recognized that at the extreme wiper positions one of the C_1 capacitors is shorted out and the other bridges R_2.

The modifications necessary for application with the LM387 are shown in Figure 2.14.17 for a supply voltage of 24V. Resistors R_4 and R_5 are added to supply negative input bias as discussed in Section 2.8. The feedback coupling capacitor C_O is necessary to block DC voltages from being fed back into the tone control circuitry and upsetting the DC bias, also to insure quiet pot operation since there are no DC level changes occurring across the capacitors, which

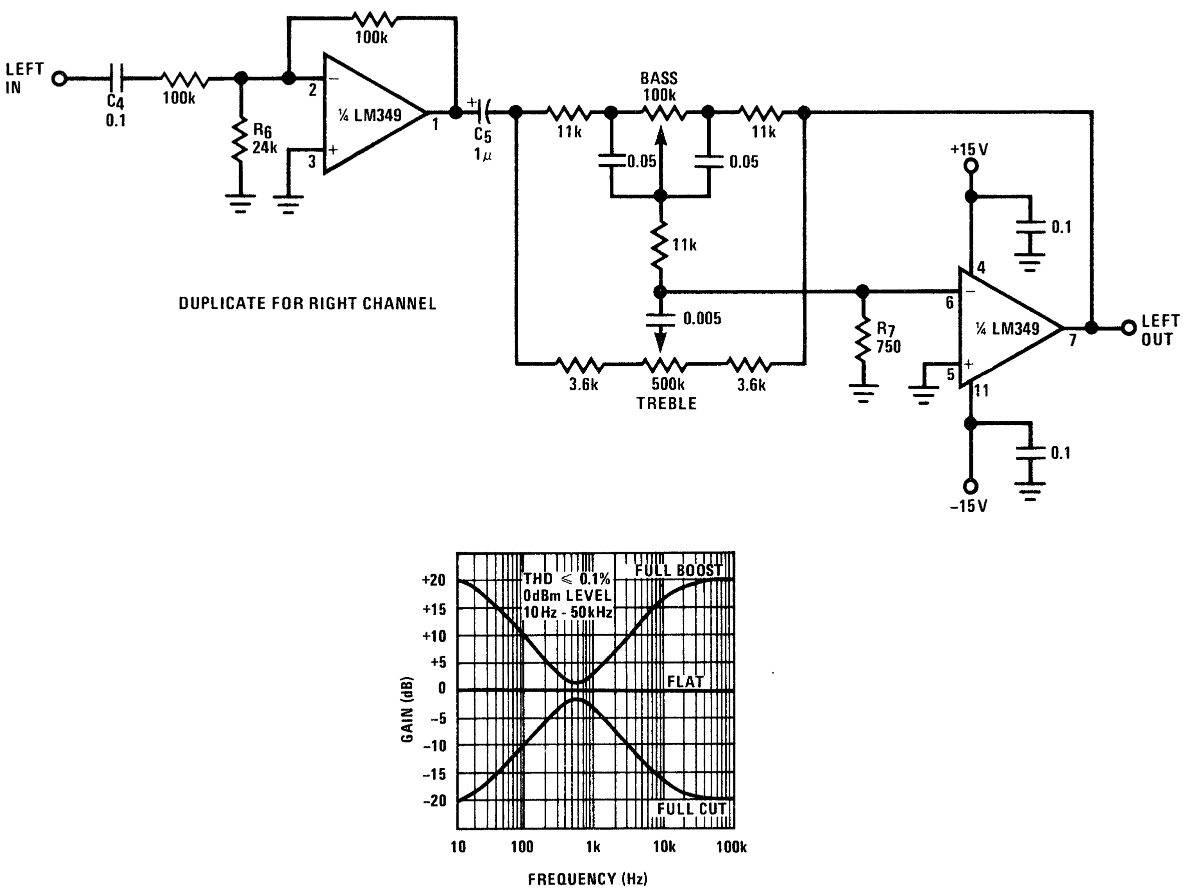


FIGURE 2.14.15 Typical Active Bass & Treble Tone Control with Buffer

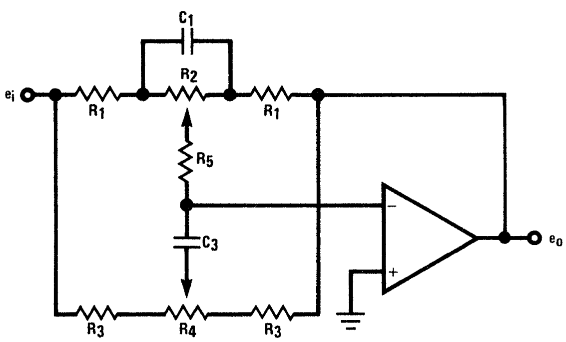
would cause "scratchiness." The R_7-C_3 network creates the input attenuation at high frequencies for stability.

For other supply voltages R_4 is recalculated as before, leaving R_5 equal to 240kΩ. It is not necessary to change R_7 since its value is dictated by the high frequency equivalent impedance seen by the inverting input (equals 33kΩ).

2.14.9 Midrange Control

The addition of a midrange control which acts to boost or cut the midrange frequencies in a manner similar to the bass and treble controls offers greater flexibility in tone control.

The midrange control circuitry appears in Figure 2.14.18. It is seen that the control is a merging together of the bass and treble controls, incorporating the bass bridging capacitor and the treble slider capacitor to form a combined network. If the bass control is, in fact, a low pass filter, and the treble control a high pass filter, then the midrange is a combination of both, i.e., a bandpass filter.



BASS

$$f_L = \frac{1}{2\pi R_2 C_1}$$

$$f_{LB} = \frac{1}{2\pi R_1 C_1}$$

$$A_{VB} = 1 + \frac{R_2}{R_1}$$

TREBLE

$$f_H = \frac{1}{2\pi R_3 C_3}$$

$$f_{HB} = \frac{1}{2\pi (R_1 + R_3 + 2 R_5) C_3}$$

$$A_{VT} = 1 + \frac{R_1 + 2 R_5}{R_3}$$

ASSUMES $R_4 \gg R_1 + R_3 + 2 R_5$

FIGURE 2.14.16 Alternate Bass Design Active Tone Control

While the additional circuitry appears simple enough, the resultant mathematics and design equations are not. In the bass and treble design of Figure 2.14.13 it is possible to include the loading effects of the bass control upon the treble circuit, make some convenient design rules, and obtain useful equations. (The treble control offers negligible load to the bass circuit.) This is possible, primarily because the frequencies of interest are far enough apart so as not to interfere with one another. Such is not the case with the midrange included. Any two of the controls appreciably loads the third. The equations that result from a detailed analysis of Figure 2.14.18 become so complex that they are useless for design. So, as is true with much of real-world engineering, design is accomplished by empirical (i.e., trial-and-error) methods. The circuit of Figure 2.14.18 gives the performance shown by the frequency plot, and should be optimum for most applications. For those who feel a change is necessary, the following guidelines should make it easier.

1. To increase (or decrease) midrange gain, decrease (increase) R_6. This will also shift the midrange center frequency higher (lower). (This change has minimal effect upon bass and treble controls.)
2. To move the midrange center frequency (while preserving gain, and with negligible change in bass and treble performance), change both C_4 and C_5. Maintain the relationship that $C_5 \approx 5C_4$. Increasing (decreasing) C_5 will decrease (increase) the center frequency. The amount of shift is approximately equal to the inverse ratio of the new capacitor to the old one. For example, if the original capacitor is C_5 and the original center frequency is f_o, and the new capacitor is C_5' with the new frequency being f_o', then

$$\frac{C_5'}{C_5} \approx \frac{f_o}{f_o'}$$

The remainder of Figure 2.14.18 is as previously described in Figure 2.14.15.

The temptation now arises to add a fourth section to the growing tone control circuitry. It should be avoided. Three paralleled sections appears to be the realistic limit to what can be expected with one gain block. Beyond three, it is best to separate the controls and use a separate op amp with each control and then sum the results. (See Section 2.17 on equalizers for details.)

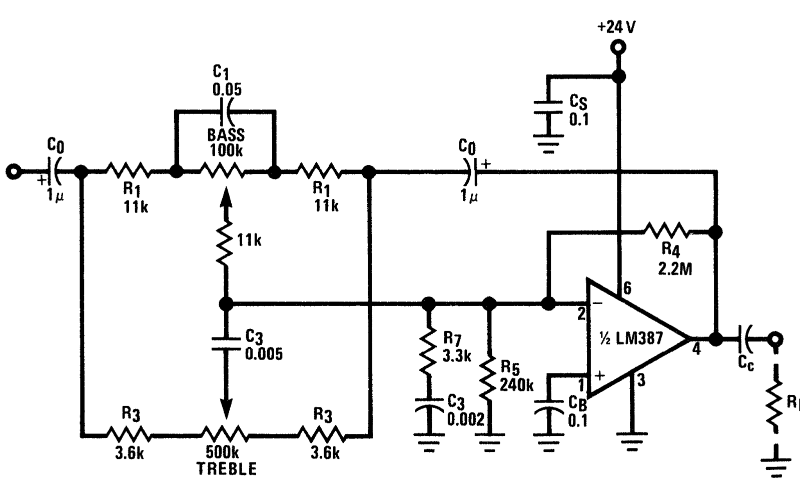
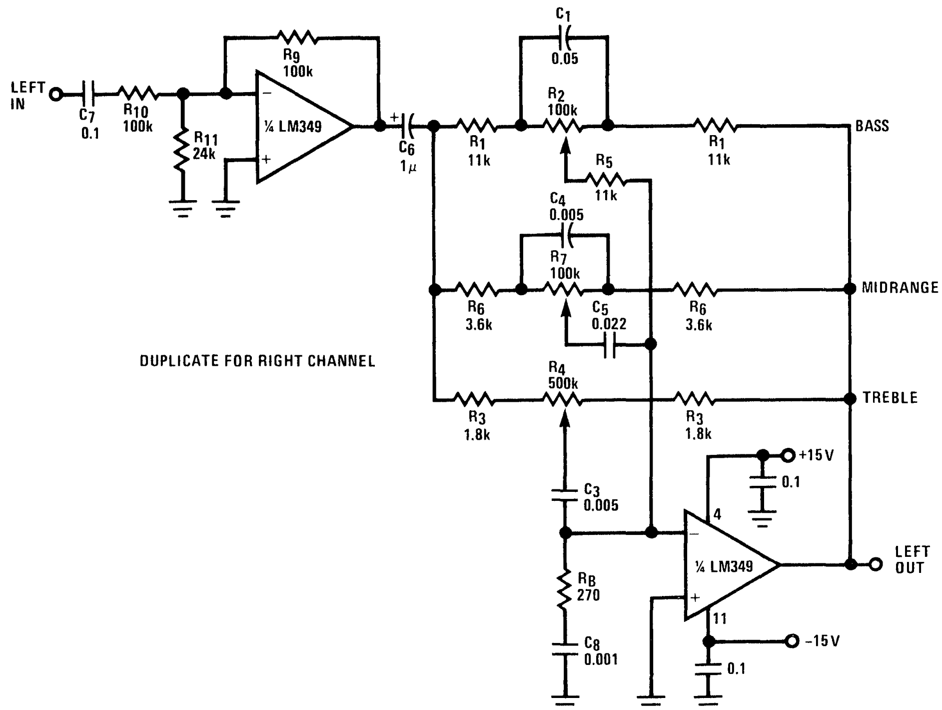


FIGURE 2.14.17 LM387 Feedback Tone Controls



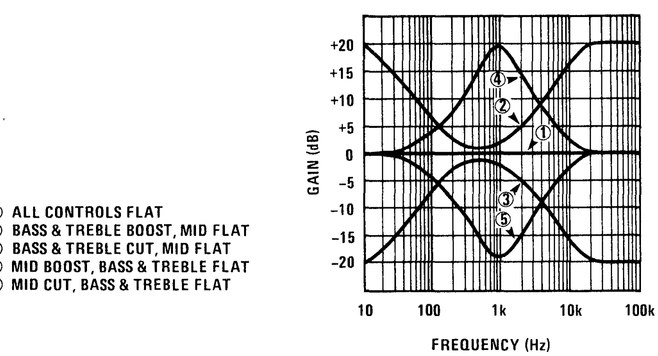


FIGURE 2.14.18 Three Band Active Tone Control (Bass, Midrange & Treble)

REFERENCES

1. Fletcher, H., and Munson, W. A., "Loudness, Its Definition, Measurement and Calculation," *J. Acoust. Soc. Am.*, vol. 5, p. 82, October 1933.
2. Baxandall, P. J., "Negative Feedback Tone Control — Independent Variation of Bass and Treble Without Switches," *Wireless World*, vol. 58, no. 10, October 1952, p. 402.

2.15 SCRATCH, RUMBLE AND SPEECH FILTERS

2.15.1 Introduction

Infinite-gain, multiple-feedback active filters using LM387 (or LM381) as the active element make simple low-cost audio filters. Two of the most popular filters found in audio equipment are SCRATCH (low pass), used to roll off excess high frequency noise appearing as hiss, ticks and pops from worn records, and RUMBLE (high pass), used to roll off low frequency noise associated with worn turntable and tape transport mechanisms. By combining low and high pass filter sections, a broadband bandpass filter is created such as that required to limit the audio bandwidth to include only speech frequencies (300 Hz-3 kHz)

2.15.2 Definition of ω_c and ω_o for 2-Pole Active Filters

When working with active filter equations, much confusion exists about the difference between the terms ω_o and ω_c. The center frequency, f_o, equals $\omega_o/2\pi$ and has meaning only for *bandpass* filters. The term ω_c and its associated frequency, f_c, is the cutoff frequency of a high or low pass filter defined as the point at which the magnitude of the response is −3 dB from that of the passband (i.e., 0.707 times the passband value). Figure 2.15.1 illustrates the two cases for two-pole filters.

Equally confusing is the concept of "Q" in relation to high and low pass two-pole active filters. The design equations contain Q; therefore it must be determined before a filter can be realized — but what does it mean? For bandpass filters the meaning of Q is clear; it is the ratio of the center frequency, f_O, to the −3dB bandwidth. *For low and high pass filters, Q only has meaning with regard to the amount of peaking occurring at f_O and the relationship between the −3dB frequency, f_C, and f_O.*

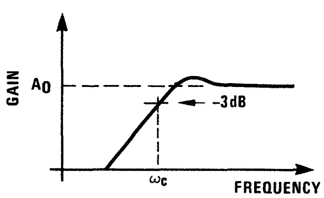
The relationship that exists between ω_O and ω_C follows:

High Pass $$\omega_C = \frac{\omega_O}{\beta} \tag{2.15.1}$$

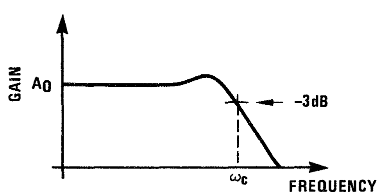
Low Pass $$\omega_C = \beta\omega_O \tag{2.15.2}$$

$$\beta = \sqrt{\left(1 - \frac{1}{2Q^2}\right) + \sqrt{\left(1 - \frac{1}{2Q^2}\right)^2 + 1}} \tag{2.15.3}$$

A table showing various values of β for several different values of Q is provided for convenience (Table 2.15.1). Notice that $\omega_C = \omega_O$ *only* for the Butterworth case (Q = 0.707). Since Butterworth filters are characterized by a maximally flat response (no peaking like that diagrammed in Figure 2.15.1), they are used most often in audio systems.



(a) High Pass



(b) Low Pass

FIGURE 2.15.1 Definition of ω_C for Low and High Pass Filters

TABLE 2.15.1 ω_C vs. Q

Q	ω_C Low-Pass	ω_O High-Pass
0.707*	$1.000\omega_O$	$1.000\omega_O$
1	$1.272\omega_O$	$0.786\omega_O$
2	$1.498\omega_O$	$0.668\omega_O$
3	$1.523\omega_O$	$0.657\omega_O$
4	$1.537\omega_O$	$0.651\omega_O$
5	$1.543\omega_O$	$0.648\omega_O$
10	$1.551\omega_O$	$0.645\omega_O$
100	$1.554\omega_O$	$0.644\omega_O$

* Butterworth

Substitution of f_C for f_O in Butterworth filter design equations is therefore permissible and experimental results will agree with calculations — but only for Butterworth. Always use Equations (2.15.1)-(2.15.3) (or Table 2.15.1) when Q equals anything other than 0.707.

2.15.3 High Pass Design

An LM387 configured as a high-pass filter is shown in Figure 2.15.2. Design procedure is to select R_2 and R_3 per Section 2.8 to provide proper bias; then, knowing desired passband gain, A_O, the Q and the corner frequency f_C, the remaining components are calculated from the following:

Calculate ω_O from $\omega_C = 2\pi f_C$ and Q using Equations (2.15.1) and (2.15.3) (or Table 2.15.1).

Let $C_1 = C_3$

Then:

$$C_1 = \frac{Q}{\omega_O R_2}(2A_O + 1) \tag{2.15.4}$$

$$C_2 = \frac{C_1}{A_O} \tag{2.15.5}$$

$$R_1 = \frac{1}{Q\,\omega_O\,C_1\,(2A_O + 1)} \tag{2.15.6}$$

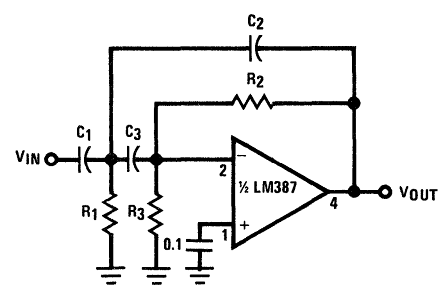
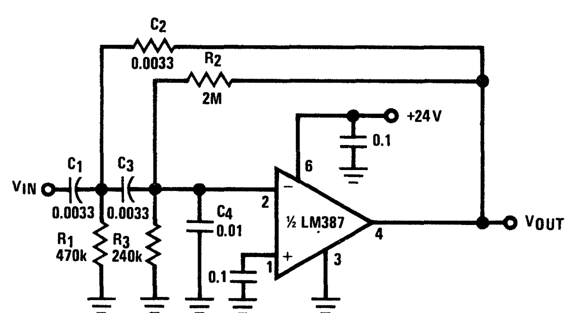


FIGURE 2.15.2 LM387 High Pass Active Filter



f_c = 50Hz
SLOPE = −12dB/OCTAVE
A_o = −1
THD ⩽ 0.1%

FIGURE 2.15.3 Rumble Filter Using LM387

Example 2.15.1

Design a two-pole active high pass filter for use as a rumble filter. Passband gain, A_O = 1, Q = 0.707 (Butterworth) and corner frequency, f_C = 50 Hz. Supply V_S = +24 V.

Solution

1. Select $R_3 = 240k$.
2. From Section 2.8,

$$R_2 = \left(\frac{V_s}{2.6'} - 1\right) R_3 = \left(\frac{24}{2.6} - 1\right) 240k = 1.98 \times 10^6$$

Use $R_2 = 2M$

3. Since $Q = 0.707$, $\omega_o = \omega_c = 2\pi f_c$ (see Table 2.15.1).
4. Let $C_1 = C_3$.
5. From Equation (2.15.4):

$$C_1 = \frac{(0.707)(2+1)}{(2\pi)(50)(2 \times 10^6)} = 3.38 \times 10^{-9}$$

Use $C_1 = C_3 = 0.0033\mu F$

6. From Equation (2.15.5):

$$C_2 = \frac{C_1}{(1)} = C_1 = 0.0033\mu F$$

7. From Equation (2.15.6):

$$R_1 = \frac{1}{(0.707)(2\pi)(50)(0.0033 \times 10^{-6})(2+1)}$$

$$= 45.5 \times 10^4$$

Use $R_1 = 470k\Omega$.

The final design appears as Figure 2.15.3. For checking and trimming purposes Equation (2.15.7) is useful:

$$f_c = \frac{1}{2\pi C_1 \sqrt{R_1 R_2}} \qquad (2.15.7)$$

Capacitor $C_4 = 0.01$ is included to guarantee high frequency stability for unity gain designs (required for $A_o \leqslant 10$).

2.15.4 Low Pass Design

The low pass configuration for a LM387 is shown in Figure 2.15.4. Design procedure is almost the reverse of the high pass case since biasing resistor R_4 will be selected last. Knowing A_o, Q and f_c, proceed by calculating a constant K per Equation (2.15.8).

$$K = \frac{1}{4Q^2(A_o + 1)} \qquad (2.15.8)$$

Arbitrarily select C_1 to be a convenient value.

Then: $C_2 = KC_1$ (2.15.9)

Calculate ω_o from $\omega_c = 2\pi f_c$ and Q using Equations (2.15.1) and (2.15.3) (or Table 2.15.1).

Then:

$$R_2 = \frac{1}{2Q\omega_o C_1 K} \qquad (2.15.10)$$

$$R_3 = \frac{R_2}{A_o + 1} \qquad (2.15.11)$$

$$R_1 = \frac{R_2}{A_o} \qquad (2.15.12)$$

$$R_4 = \frac{R_2 + R_3}{\left(\frac{V_s}{2.6} - 1\right)} \qquad (2.15.13)$$

Example 2.15.2

Design a two-pole active low-pass filter for use as a scratch filter. Passband gain, $A_o = 1$, $Q = 0.707$ (Butterworth) and corner frequency $f_c = 10kHz$. Supply $V_s = +24V$.

Solution

1. From Equation (2.15.8):

$$K = \frac{1}{(4)(0.707)^2(1+1)} = 0.25$$

2. Select $C_1 = 560pF$ (arbitrary choice).
3. From Equation (2.15.9):

$$C_2 = KC_1 = (0.25)(560pF) = 140pF$$

Use $C_2 = 150pF$

4. Since $Q = 0.707$, $\omega_o = \omega_c = 2\pi f_c$ (see Table 2.15.1).
5. From Equation (2.15.10):

$$R_2 = \frac{1}{(2)(0.707)(2\pi)(10kHz)(560pF)(0.25)} = 80.4k$$

Use $R_2 = 82k$

6. From Equation (2.15.11):

$$R_3 = \frac{82k}{2} = 41k$$

Use $R_3 = 39k$

7. From Equation (2.15.12):

$$R_1 = \frac{R_2}{1} = R_2 = 82k$$

8. From Equation (2.15.13):

$$R_4 = \frac{82k + 39k}{\left(\frac{24}{2.6} - 1\right)} = 14.7k$$

Use $R_4 = 15k$

The complete design (Figure 2.15.5) includes C_3 for stability and input blocking capacitor C_4. Checking and trimming can be done with the aid of Equation (2.15.14).

$$f_o = \frac{Q}{\pi C_1} \sqrt{\frac{A_o + 1}{R_2 R_3}} \qquad (2.15.14)$$

2.15.5 Speech Filter

A speech filter consisting of a highpass filter based on Section 2.15.2, in cascade with a low pass based on Section 2.15.3, is shown in Figure 2.15.6 with its frequency response as Figure 2.15.7. The corner frequencies are 300Hz and 3kHz with roll-off of –40dB/decade beyond the corners. Measured THD was 0.07% with a 0dBm signal of 1kHz. Total output noise with input shorted was 150μV and is

2

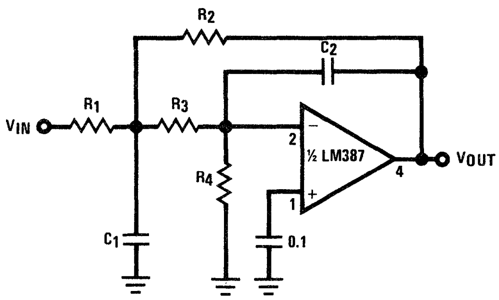


FIGURE 2.15.4 LM387 Low Pass Active Filter

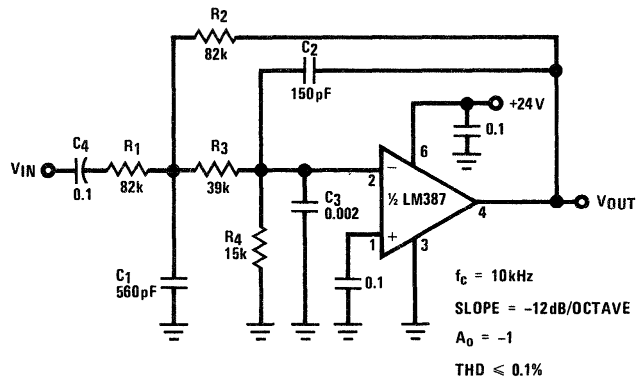


FIGURE 2.15.5 Scratch Filter Using LM387

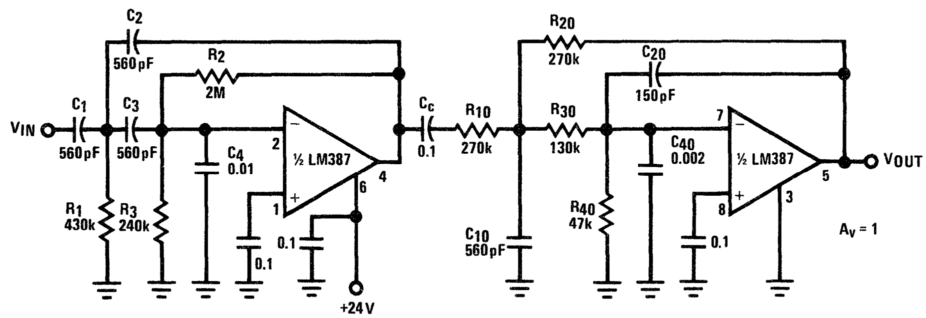


FIGURE 2.15.6 Speech Filter (300Hz-3kHz Bandpass)

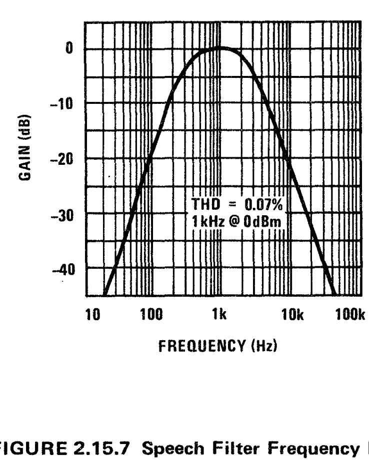


FIGURE 2.15.7 Speech Filter Frequency Response

due mostly to thermal noise of the resistors, yielding S/N of 74dBm. The whole filter is very compact since the LM387 dual preamp is packaged in the 8-pin minidip, making tight layout possible.

2.16 BANDPASS ACTIVE FILTERS

Narrow bandwidth bandpass active filters do not require cascading of low and high pass sections as described in Section 2.15.4. A single amplifier bandpass filter using the LM387 (Figure 2.16.1) is capable of $Q \leq 10$ for audio frequency low distortion applications. The wide gain bandwidth (20MHz) and large open loop gain (104dB) allow high frequency, low distortion performance unobtainable with conventional op amps.

Beginning with the desired f_O, A_O and Q, design is straightforward. Start by selecting R_3 and R_4 per Section 2.8, *except use* $24k\Omega$ *as an upper limit of* R_4 (instead of $240k\Omega$). This minimizes loading effects of the LM387 for high Q designs.

Let $C_1 = C_2$. Then:

$$R_1 = \frac{R_3}{2A_O} \tag{2.16.1}$$

$$C_1 = \frac{Q}{A_O\,\omega_O\,R_1} \tag{2.16.2}$$

$$R_2 = \frac{Q}{(2Q^2 - A_o)\,\omega_o C_1} \qquad (2.16.3)$$

For checking and trimming, use the following:

$$A_o = \frac{R_3}{2R_1} \qquad (2.16.4)$$

$$f_o = \frac{1}{2\pi C_1}\sqrt{\frac{R_1 + R_2}{R_1 R_2 R_3}} \qquad (2.16.5)$$

$$Q = \frac{1}{2}\omega_o R_3 C_1 \qquad (2.16.6)$$

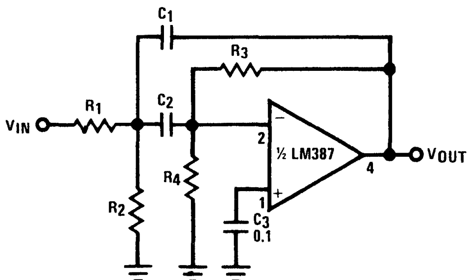


FIGURE 2.16.1 LM387 Bandpass Active Filter

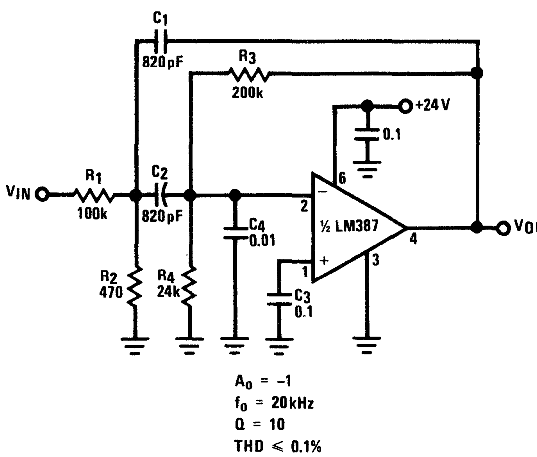


FIGURE 2.16.2 20kHz Bandpass Active Filter

Example 2.16.1

Design a two-pole active bandpass filter with a center frequency $f_o = 20\text{kHz}$, midband gain $A_o = 1$, and a bandwidth of 2000Hz. A single supply, $V_s = 24\text{V}$, is to be used.

Solution

1. $Q \triangleq \frac{f_o}{BW} = \frac{20\text{kHz}}{2000\text{Hz}} = 10, \quad \omega_o = 2\pi f_o$

2. Let $R_4 = 24\text{k}\Omega$.

3. $R_3 = \left(\frac{V_s}{2.6} - 1\right) R_4 = \left(\frac{24}{2.6} - 1\right) 24\text{k} = 1.98 \times 10^5$

   Use $R_3 = 200\text{k}$

4. From Equation (2.16.1):

   $$R_1 = \frac{R_3}{2A_o} = \frac{200\text{k}}{2} = 100\text{k}$$

   $R_1 = 100\text{k}$

5. Let $C_1 = C_2$; then, from Equation (2.16.2):

   $$C_1 = \frac{Q}{A_o \omega_o R_1} = \frac{10}{(1)(2\pi)(20\text{k})(1 \times 10^5)} = 796\text{pF}$$

   Use $C_1 = 820\text{pF}$

6. From Equation (2.16.3):

   $$R_2 = \frac{Q}{(2Q^2 - A_o)\,\omega_o C_1}$$

   $$= \frac{10}{[(2)(10)^2 - 1]\,(2\pi)(20\text{k})(820\text{pF})} = 488\Omega$$

   Use $R_2 = 470\Omega$

The final design appears as Figure 2.16.2. Capacitor C_3 is used to AC ground the positive input and can be made equal to $0.1\mu F$ for all designs. Input shunting capacitor C_4 is included for stability since the design gain is less than 10.

2.17 OCTAVE EQUALIZER

An octave equalizer offers the user several bands of tone control, separated an octave apart in frequency with independent adjustment of each. It is designed to compensate for any unwanted amplitude-frequency or phase-frequency characteristics of an audio system.

The midrange tone control circuit described in Section 2.14 can be used separately to make a convenient ten band octave equalizer. Design equations result from a detailed analysis of Figure 2.17.1, where a typical section is shown. Resistors R_3 have been added to supply negative input DC bias currents, and to guarantee unity gain at low frequencies. This circuit is particularly suited for equalizer applications since it offers a unique combination of results depending upon the slider position of R_2. With R_2 in the flat position (i.e., centered) the circuit becomes an all-pass with unity gain; moving R_2 to full boost results in a bandpass characteristic, while positioning R_2 in full cut creates a band-reject (notch) filter.

Writing the transfer function for Figure 2.17.1 in its general form for max boost (assuming only $R_3 \gg R_1$) results in Equation (2.17.1).

$$\frac{e_o}{e_i} = -\frac{S^2 + \left[\frac{2R_1R_2C_1 + R_3(R_1 + R_2)C_2}{R_1R_2R_3C_1C_2}\right]S + \frac{2R_1 + R_2}{R_1R_2R_3C_1C_2}}{S^2 + \left[\frac{(R_1 + R_2)C_2 + 2R_2C_1 + R_3C_2}{R_2R_3C_1C_2}\right]S + \frac{2R_1 + R_2}{R_1R_2R_3C_1C_2}} \qquad (2.17.1)$$

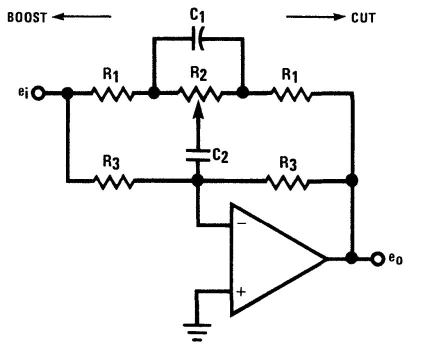


FIGURE 2.17.1 Typical Octave Equalizer Section

Equation (2.17.1) has the form of Equation (2.17.2):

$$\frac{e_o}{e_i} = -\frac{S^2 + K 2\rho\,\omega_o S + \omega_o^2}{S^2 + 2\rho\,\omega_o S + \omega_o^2} \quad (2.17.2)$$

where: $Q = \frac{1}{2\rho}$, A_o = gain @ f_o = K, $\omega_o = 2\pi f_o$

Equating coefficients yields Equations (2.17.3)-(2.17.5):

$$\omega_o = \sqrt{\frac{2R_1 + R_2}{R_1 R_2 R_3 C_1 C_2}} \quad (2.17.3)$$

$$A_o = -\frac{2R_1 R_2 C_1 + R_3(R_1 + R_2)C_2}{2R_1 R_2 C_1 + R_1(R_2 + R_3)C_2} \quad (2.17.4)$$

$$Q = \sqrt{\frac{2R_1 + R_2}{R_1 R_2 R_3 C_1 C_2}}\left(\frac{R_2 R_3 C_1 C_2}{(R_1 + R_2)C_2 + 2R_2 C_1 + R_3 C_2}\right) \quad (2.17.5)$$

In order to reduce these equations down to something useful, it is necessary to examine what is required of the finished equalizer in terms of performance. For normal home use, ±12dB of boost and cut is adequate, which means only a moderate amount of passband gain is necessary; and since the filters will be centered one octave apart in frequency a large Q is not necessary (Q = 1-2 works fine). What *is* desirable is for the passband ripple (when all filters are at maximum) to be less than 3dB.

Examination of Equation (2.17.5) in terms of optimizing the ratio of C_1 and C_2 in order to maximize Q shows a good choice is to let $C_1 = 10C_2$. A further design rule that is reasonable is to make $R_3 = 10R_2$, since R_3 is unnecessary for the filter section. Applying these rules to Equations (2.17.3)-(2.17.5) produces some useful results:

$$\omega_o = 2\pi f_o = \frac{1}{10R_2C_2}\sqrt{2 + \frac{R_2}{R_1}} \quad (2.17.6)$$

$$A_o = 1 + \frac{R_2}{3R_1} \quad (2.17.7)$$

$$Q = \sqrt{\frac{2R_1 + R_2}{9.61R_1}} \quad (2.17.8)$$

Rewriting (2.17.7) and (2.17.8) yields:

$$R_2 = 3(A_o - 1)R_1 \quad (2.17.9)$$

$$R_2 = (9.61Q^2 - 2)R_1 \quad (2.17.10)$$

Combining (2.17.9) and (2.17.10) gives:

$$A_o = \left(\frac{9.61Q^2 - 2}{3}\right) + 1 \quad (2.17.11)$$

From Equation (2.17.11) it is seen that gain and Q are intimately related and that large gains mean large Qs and vice versa. Equations (2.17.9) and (2.17.10) show that R_1 and R_2 are not independent, which means one may be arbitrarily selected and from it (knowing A_o and/or Q) the other is found.

Design

1. Select R_2 = 100k.
2. $R_3 = 10R_2 = 10(100k)$

   R_3 = 1 Meg
3. Let A_o = 12dB = 4V/V and from Equation (2.17.9):

   $$R_1 = \frac{R_2}{3(A_o - 1)} = \frac{100k}{3(4-1)} = 1.11 \times 10^4$$

   Use R_1 = 10k.
4. Check Q from Equation (2.17.8):

   $$Q = \sqrt{\frac{2(10k) + 100k}{(9.61)(10k)}}$$

   Q = 1.12, which is satisfactory.
5. Calculate C_2 from Equation (2.17.6) and $C_1 = 10C_2$:

   $$C_2 = \frac{1}{2\pi f_o (10R_2)}\sqrt{2 + \frac{R_2}{R_1}}$$

   $$C_2 = \frac{1}{2\pi f_o (10)(100k)}\sqrt{2 + \frac{100k}{10k}}$$

   $$C_2 = \frac{5.513 \times 10^{-7}}{f_o}$$

A table of standard values for C_1 and C_2 vs. f_o is given below:

TABLE 2.17.1

f_o (Hz)	C_1	C_2
32	0.18μF	0.018μF
64	0.1μF	0.01μF
125	0.047μF	0.0047μF
250	0.022μF	0.0022μF
500	0.012μF	0.0012μF
1k	0.0056μF	560pF
2k	0.0027μF	270pF
4k	0.0015μF	150pF
8k	680pF	68pF
16k	360pF	36pF

The complete design appears as Figure 2.17.2. While it appears complicated, it is really just repetitious. By using quad amplifier ICs, the whole thing consists of only three integrated circuits. Figure 2.17.2 is for one channel and would be duplicated for a stereo system. The input buffer amplifier guarantees a low source impedance to drive the equalizer and presents a large input impedance for the preamplifier. Resistor R_8 is necessary to stabilize the LM349 while retaining its fast slew rate (2V/μs). The output amplifier is a unity gain, inverting summer used to add each equalized octave of frequencies back together again. One aspect of the summing circuit that may appear odd is that the original signal is subtracted from the sum via R_{20}. (It is subtracted rather than added because each equalizer section inverts the signal relative to the output of the buffer and R_{20} delivers the original signal without inverting.) The reason this subtraction is necessary is in order to maintain a unity gain system. Without it the output would equal ten times the input, e.g., an input of 1V, with all pots flat, would produce 1V at each equalizer output – the sum of which is 10V. By scaling R_{20} such that the input signal is multiplied by 9 before the subtraction, the output now becomes 10V - 9V = 1V output, i.e., unity gain. The addition of R_4 to each section is for stability. Capacitor C_3 minimizes possibly large DC offset voltages from appearing at the output. If the driving source has a DC level then an input capacitor is necessary (0.1μF), and similarly, if the load has a DC level, then an output capacitor is required.

It is possible to generate just about any frequency response imaginable with this ten band octave equalizer. A few possibilities are given in Figure 2.17.3.

2

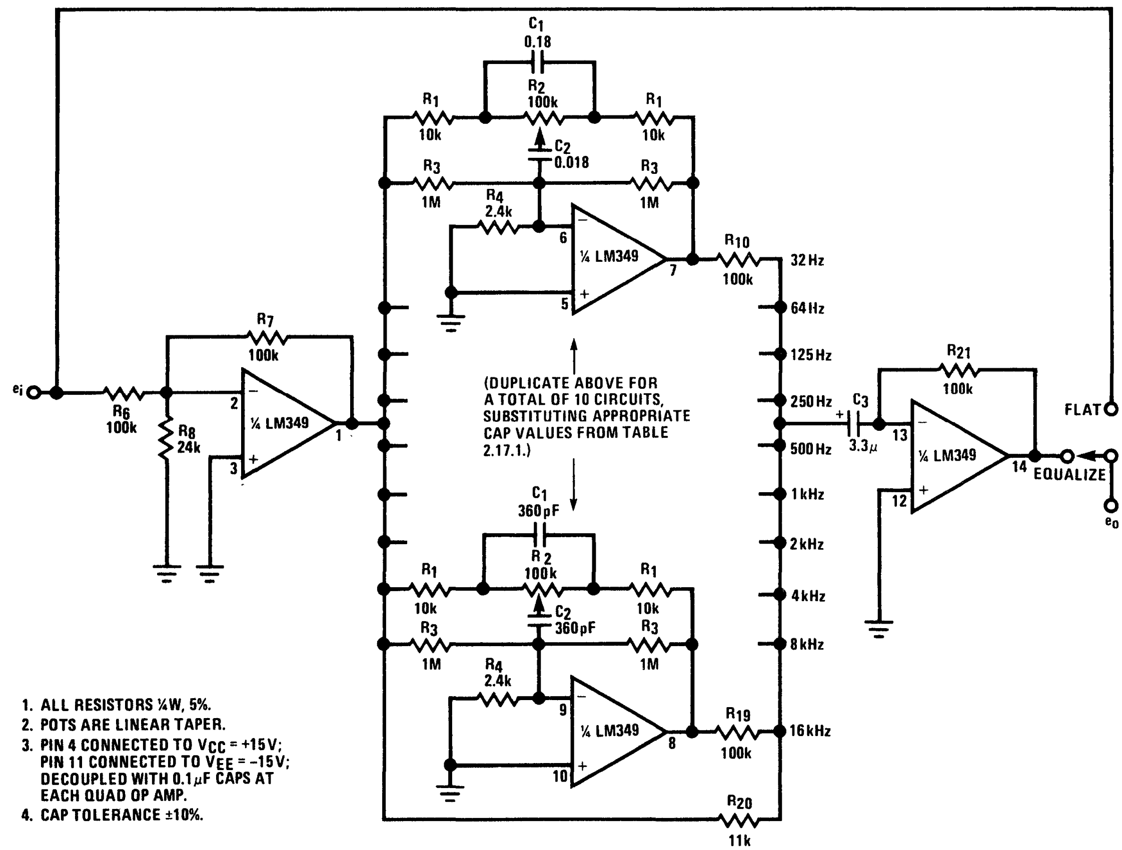


FIGURE 2.17.2 Ten Band Octave Equalizer

① ALL CONTROLS FLAT
② 500Hz BOOST/CUT, ALL OTHERS FLAT
③ 1kHz BOOST/CUT, ALL OTHERS FLAT
④ 500Hz, 1kHz, 2kHz, 4kHz BOOSTED, ALL OTHERS FLAT

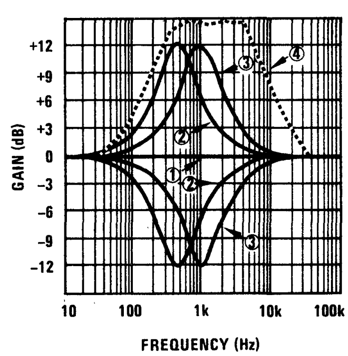


FIGURE 2.17.3 Typical Frequency Response of Equalizer