AUDIO/RADIO HANDBOOK

NATIONAL SEMICONDUCTOR

1980 AUDIO/RADIO HANDBOOK National Semiconductor

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ISBN 1-882580-35-4
51495
9 781882 580354

US $14.95

DA-CP37RRD53M50/Printed in U.S.A.

AUDIO/RADIO HANDBOOK

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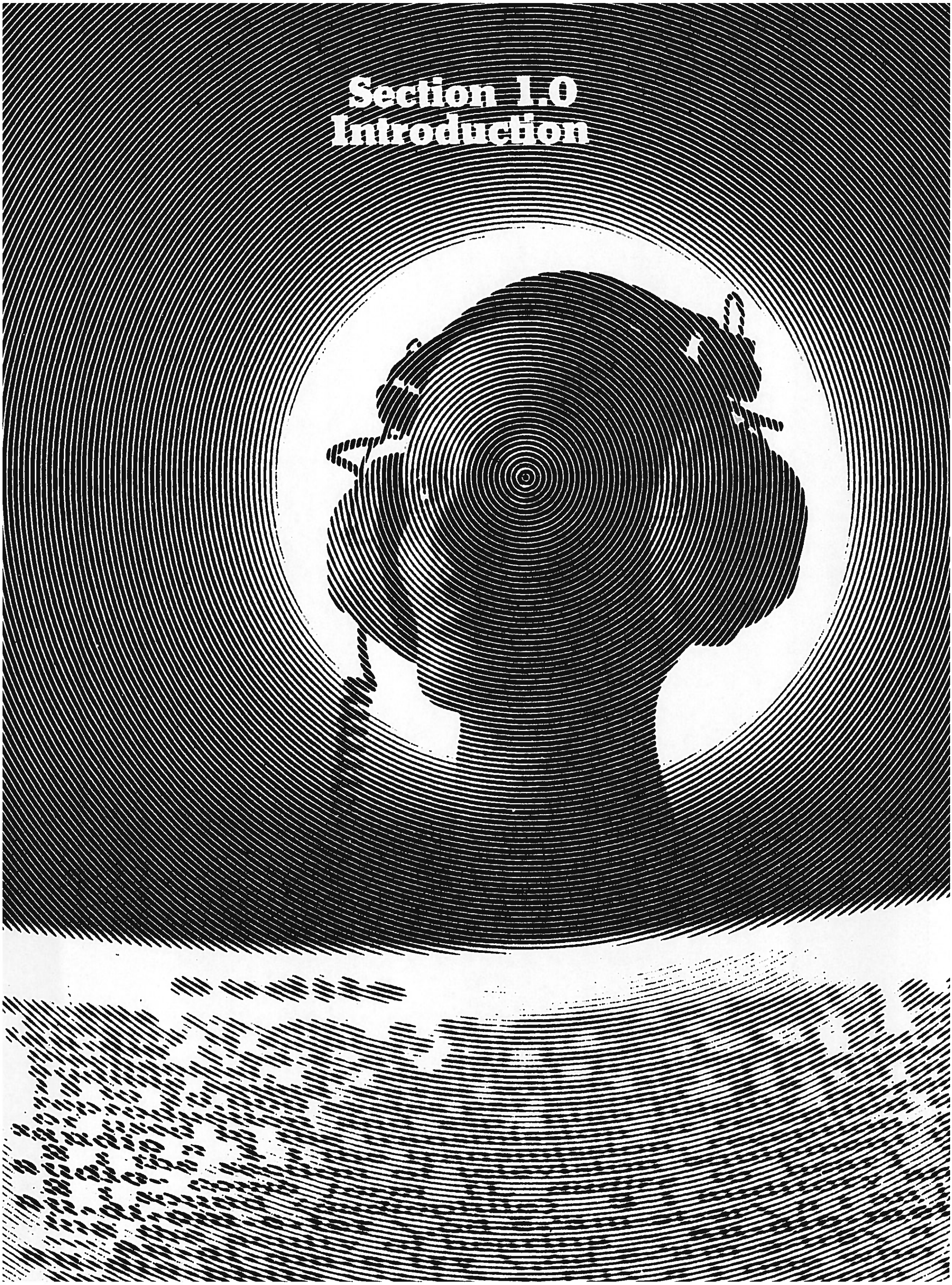
*"Floobydust" is a contemporary term derived from the archaic Latin *miscellaneus*, whose disputed history probably springs from Greek origins (influenced, of course, by Egyptian linguists) — meaning here "a mixed bag."

Device Index

Section 1.0 Introduction



1.0 Introduction

In just a few years time, National Semiconductor Corporation has emerged as a leader – indeed, if not *the* leader in all areas of integrated circuit products. National's well-known linear and digital ICs have become industry standards in all areas of design. This handbook exists to acquaint those involved in audio systems design with National Semiconductor's broad selection of integrated circuits specifically designed to meet the stringent requirements of accurate audio reproduction. Far from just a collection of data sheets, this manual contains detailed discussions, including complete design particulars, covering many areas of audio. Thorough explanations, complete with real-world design examples, make clear several audio areas never before available to the general public.

1.1 SCOPE OF HANDBOOK

Between the hobbyist and the engineer, the amateur and the professional, the casual experimenter and the serious product designer there exists a chaotic space filled with Laplace transforms, Fourier analysis, complex calculus, Maxwell's equations, solid-state physics, wave mechanics, holes, electrons, about four miles of effete mysticism, and, maybe, one inch of compassion. This audio handbook attempts to disperse some of the mist. Its contents cover many of the multidimensional fields of audio, with emphasis placed on intuition rather than rigor, favoring the practical over the theoretical. Each area is treated at the minimum depth felt necessary for adequate comprehension. Mathematics is not avoided – only reserved for just those areas demanding it. Some areas are more "cookbook" than others, the choice being dictated by the material and Mother Nature.

General concepts receive the same thorough treatment as do specific devices, based upon the belief that the more informed integrated circuit user has fewer problems using integrated circuits. Scanning the Table of Contents will indicate the diversity and relevance of what is inside. Within the broad scope of audio, only a few areas could be covered in a book this size; those omitted tend to be ones not requiring active devices for implementation (e.g., loudspeakers, microphones, transformers, styli, etc.).

Have fun.

1.2 IC PARAMETERS APPLIED TO AUDIO

Audio circuits place unique requirements upon IC parameters which, if understood, make proper selection of a specific device easier. Most linear integrated circuits fall into the "operational amplifier" category where design emphasis has traditionally been placed upon perfecting those parameters most applicable to DC performance. But what about AC performance? Specifically, what about audio performance?

Audio is really a rather specialized area, and its requirements upon an integrated circuit may be stated quite concisely: The IC must process complex AC signals comprised of frequencies ranging from 20 hertz to 20k hertz, whose amplitudes vary from a few hundred microvolts to several volts, with a transient nature characterized by steep, compound wavefronts separated by unknown periods of absolute silence. This must be done without adding distortion of any sort, either harmonic, amplitude, or phase; and it must be done noiselessly – in the sun, and in the snow – forever.

Unfortunately, this IC doesn't exist; we're working on it, but it's not ready for immediate release. Meanwhile, the problem remains of how to choose from what is available. For the most part, DC parameters such as offset voltages and currents, input bias currents and drift rates may be ignored. Capacitively coupling for bandwidth control and single supply operation negates the need for concern about DC characteristics. Among the various specifications applicable to AC operation, perhaps slew rate is the most important.

1.2.1 Slew Rate

The slew rate limit is the maximum rate of change of the amplifier's output voltage and is due to the fact that the compensation capacitor inside the amplifier only has finite currents[1] available for charging and discharging (see Section 4.1.2). A sinusoidal output signal will cease being small signal when its maximum rate of change equals the slew rate limit S_r of the amplifier. The maximum rate of change for a sine wave occurs at the zero crossing and may be derived as follows:

$$v_o = V_p \sin 2\pi ft \quad (1.2.1)$$

$$\frac{dv_o}{dt} = 2\pi f V_p \cos 2\pi ft \quad (1.2.2)$$

$$\left.\frac{dv_o}{dt}\right|_{t=0} = 2\pi f V_p \quad (1.2.3)$$

$$S_r = 2\pi f_{max} V_p \quad (1.2.4)$$

where: v_o = output voltage

V_p = peak output voltage

$S_r = \text{maximum } \frac{dv_o}{dt}$

The maximum sine wave frequency an amplifier with a given slew rate will sustain without causing the output to take on a triangular shape is therefore a function of the peak amplitude of the output and is expressed as:

$$f_{max} = \frac{S_r}{2\pi V_p} \quad (1.2.5)$$

Equation (1.2.5) demonstrates that the borderline between small signal response and slew rate limited response is not just a function of the peak output signal but that by trading off either frequency or peak amplitude one can continue to have a distortion free output. Figure (1.2.1) shows a quick reference graphical presentation of Equation (1.2.5) with the area above any V_{PEAK} line representing an undistorted small signal response and the area below a given V_{PEAK} line representing a distorted sine wave response due to slew rate limiting.

As a matter of convenience, amplifier manufacturers often give a "full-power bandwidth" or "large signal response" on their specification sheets.

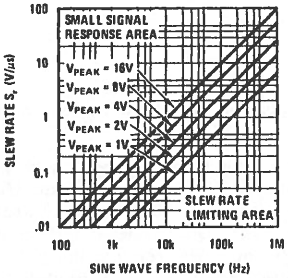


FIGURE 1.2.1 Sine Wave Response

This frequency can be derived by inserting the amplifier slew rate and peak rated output voltage into Equation (1.2.5). The bandwidth from DC to the resulting f_{max} is the full-power bandwidth or "large signal response" of the amplifier. For example, the full-power bandwidth of the LM741 with a 0.5V/µs S_r is approximately 6kHz while the full-power bandwidth of the LF356 with a S_r of 12V/µs is approximately 160kHz.

1.2.2 Open Loop Gain

Since virtually all of an amplifier's closed loop performance depends heavily upon the amount of loop-gain available, open loop gain becomes very important. Input impedance, output impedance, harmonic distortion and frequency response all are determined by the difference between open loop gain and closed loop gain, i.e., the loop gain (in dB). Details of this relationship are covered in Section 2.1. What is desired is high open loop gain – the higher the better.

1.2.3 Bandwidth and Gain-Bandwidth

Closely related to the slew rate capabilities of an amplifier is its unity gain bandwidth, or just "bandwidth." The "bandwidth" is defined as the frequency where the open loop gain crosses unity. High slew rate devices will exhibit wide bandwidths.

Because the size of the capacitor required for internally compensated devices determines the slew rate – hence, the bandwidth – one method used to design faster amplifiers is to simply make the capacitor smaller. This creates a faster IC but at the expense of unity-gain stability. Known as a *decompensated* (as opposed to *uncompensated* – no capacitor) amplifier, it is ideal for most audio applications requiring gain.

The term *gain-bandwidth* is used frequently in place of "unity gain bandwidth." The two terms are equal numerically but convey slightly different information. Gain-bandwidth, or gain-bandwidth product, is a combined measure of open loop gain and frequency response – being the product of the available gain at any frequency times that frequency. For example, an LM381 with gain of around 2000V/V at 10kHz yields a GBW equal to 20MHz. The GBW requirement for accurate audio reproduction may be derived for general use by requiring a minimum loop-gain of 40dB (for distortion reduction) at 20kHz for an amplifier with a closed loop gain of 20dB. This means a minimum open loop gain of 60dB (1000V/V) at 20kHz, or a GBW equal to 20MHz. Requirements for lo-fi and mid-fi designs, where reduced frequency response and higher distortion are allowable, would, of course, be less.

1.2.4 Noise

The importance of noise performance from an integrated circuit used to process audio is obvious and needs little discussion. Noise specifications normally appear as "Total Equivalent Input Noise Voltage," stated for a certain source impedance and bandwidth. This is the most useful number, since it is what gets amplified by the closed loop gain of the amplifier. For high source impedances, noise current becomes important and must be considered, but most driving impedances are less than 600Ω, so knowledge of noise voltage is sufficient.

1.2.5 Total Harmonic Distortion

Need for low total harmonic distortion (THD) is also obvious and need not be belabored. THD performance for preamplifier ICs will state the closed loop gain and frequency at which it was measured, while audio power amplifiers will also include the power output.

1.2.6 Supply Voltage

Consideration of supply voltage limits may be more important than casual thought would indicate. For preamplifier ICs and general purpose op amps, attention needs to be directed to supply voltage from a dynamic range, or "headroom," standpoint. Much of audio processing requires headroom on the order of 20-40dB if transient clipping is to be avoided. For a design needing 26dB dynamic range with a nominal input of 50mV and operating at a closed loop gain of 20dB, a supply voltage of at least 30V would be required. It is important, therefore, to be sure the IC has a supply voltage rating adequate to handle the worst case conditions. These occur for high power line cases and low current drain, requiring the IC user to check the "absolute maximum" ratings for supply voltage to be sure there are no conditions under which they will be exceeded. Remember, "absolute maximum" means just that – it is not the largest supply you can apply; it is the value which, if exceeded, causes all bets to be cancelled. This problem is more acute for audio power devices since their supplies tend to sag greatly, i.e., the difference between no power out and full power out can cause variations in power supply level of several volts.

1.2.7 Ripple Rejection

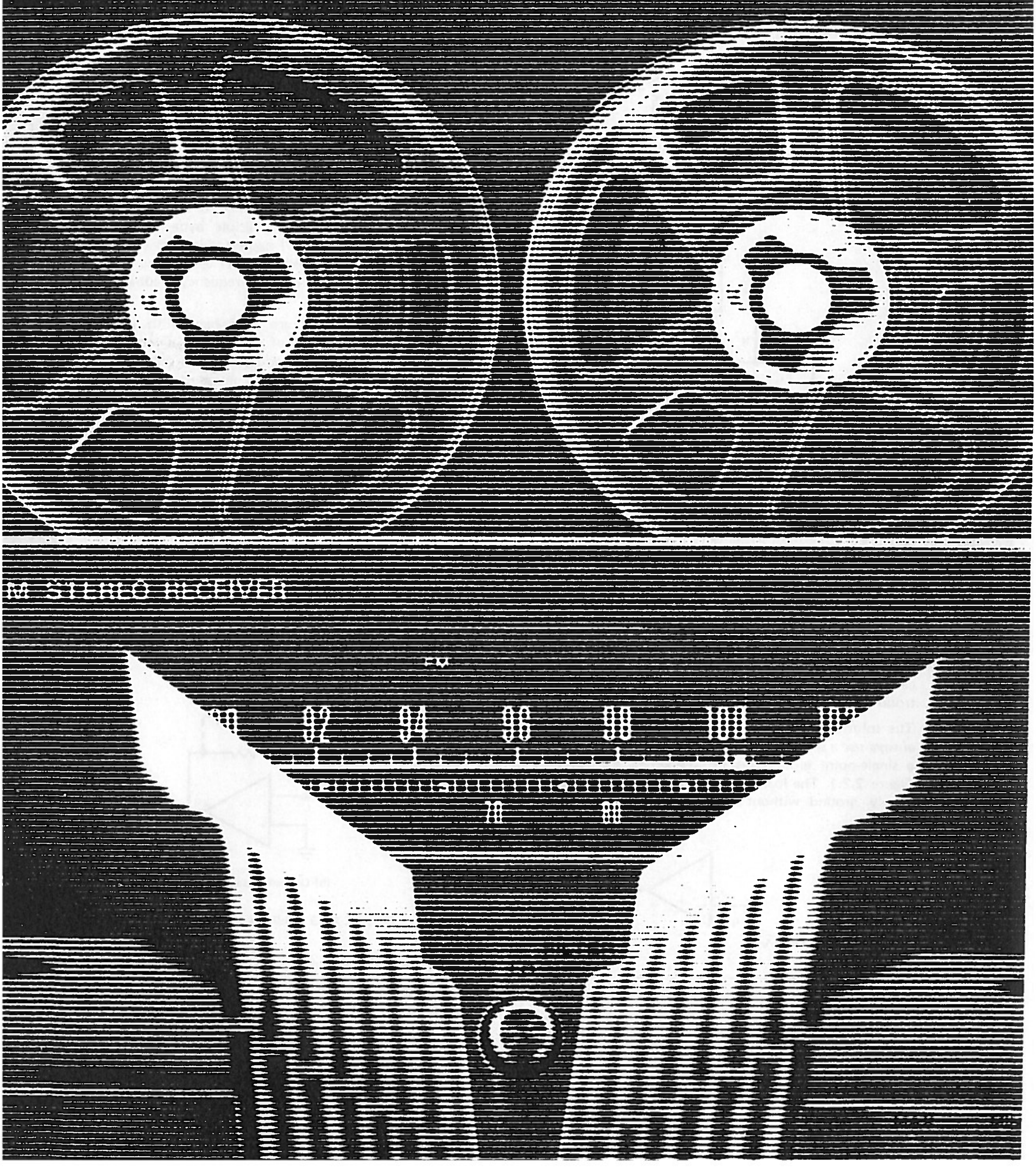
An integrated circuit's ability to reject supply ripple is important in audio applications. The reason has to do with minimizing hum within the system – high ripple rejection means low ripple bleedthrough to the output, where it adds to the signal as hum. Relaxed power supply design (i.e., ability to tolerate large amounts of ripple) is allowed with high ripple rejection parts.

Supply ripple rejection specifications cite the amount of rejection to be expected at a particular frequency (normally 120Hz), or over a frequency band, and is usually stated in dB. The figure may be "input referred" or "output referred." If input referred, then it is analogous to input referred noise and this amount of ripple will be multiplied by the gain of the amplifier. If output referred, then it is the amount of ripple expected at the output for the given conditions.

REFERENCES

1. Solomon, J. E., Davis, W. R., and Lee, P. L., "A Self-Compensated Monolithic Operational Amplifier with Low Input Current and High Slew Rate," *ISSCC Digest Tech. Papers,* February 1969, pp. 14-15.

Section 2.0
Preamplifiers



Real-world ground leads possess finite resistance, and the currents running through them will cause finite voltage drops. If two ground return lines tie into the same path at different points there will be a voltage drop between them. Figure 2.2.1a shows a common-ground example where the positive input ground and the load ground are returned to the supply ground point via the same wire. The addition of the finite wire resistance (Figure 2.2.1b) results in a voltage difference between the two points as shown.

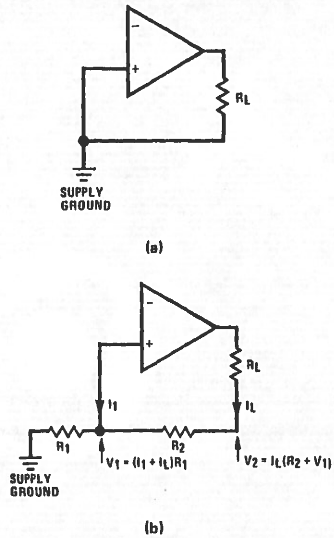


FIGURE 2.2.1 Ground Loop Example

Load current I_L will be much larger than input bias current I_1, thus V_1 will follow the output voltage directly, i.e., in phase. Therefore the voltage appearing at the non-inverting input is effectively positive feedback and the circuit may oscillate. If there were only one device to worry about then the values of R1 and R2 would probably be small enough to be ignored; however, several devices normally comprise a total system. Any ground return of a separate device, whose output is in phase, can feedback in a similar manner and cause instabilities. Out of phase ground loops also are troublesome, causing unexpected gain and phase errors.

The solution to this and other ground loop problems is to *always use a single-point ground system*. Figure 2.2.2 shows a single-point ground system applied to the example of Figure 2.2.1. The load current now returns directly to the supply ground without inducing a feedback voltage as before.

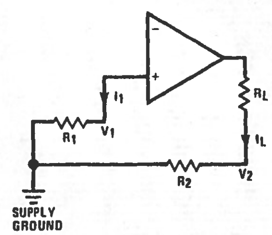


FIGURE 2.2.2 Single-Point Ground System

The single-point ground concept should be applied rigorously to all components and all circuits. Violations of single-point grounding are most common among printed circuit board designs. Since the circuit is surrounded by large ground areas the temptation to run a device to the closest ground spot is high. This temptation must be avoided if stable circuits are to result.

A final rule is to make all ground returns low resistance and low inductance by using large wire and wide traces.

2.2.3 Supply Bypassing

Many IC circuits appearing in print (including many in this handbook) do not show the power supply connections or the associated bypass capacitors for reasons of circuit clarity. *Shown or not, bypass capacitors are always required.* Ceramic disc capacitors (0.1μF) or solid tantalum (1μF) with short leads, and located close (within one inch) to the integrated circuit are usually necessary to prevent interstage coupling through the power supply internal impedance. Inadequate bypassing will manifest itself by a low frequency oscillation called "motorboating" or by high frequency instabilities. Occasionally multiple bypassing is required where a 10μF (or larger) capacitor is used to absorb low frequency variations and a smaller 0.1μF disc is paralleled across it to prevent any high frequency feedback through the power supply lines.

In general, audio ICs are wide bandwidth (~ 10MHz) devices and decoupling of each device is required. Some applications and layouts will allow one set of supply bypassing capacitors to be used common to several ICs. This condition cannot be assumed, but must be checked out prior to acceptance of the layout. Motorboating will be audible, while high frequency oscillations must be observed with an oscilloscope.

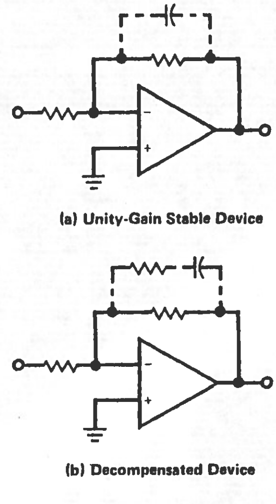


FIGURE 2.2.3 Addition of Feedback Capacitor

2.2.4 Additional Stabilizing Tips

If all of the previous rules are followed closely, no instabilities should occur within the circuit; however, Murphy being the way he is, some circuits defy these rules and oscillate anyway. Several additional techniques may be required when persistent oscillations plague a circuit:

- Reduce high impedance positive inputs to the minimum allowable value (e.g., replace 1 Meg biasing resistors with 47k ohm, etc.).
- Add small (< 100pF) capacitors across feedback resistors to reduce amplifier gain at high frequencies (Figure 2.2.3). **Caution: this assumes the amplifier is unity-gain stable.** If not, addition of this capacitor will *guarantee* oscillations. (For amplifiers that are not unity-gain stable, place a resistor in series with the capacitor such that the gain does not drop below where it is stable.)
- Add a small capacitor (size is a function of source resistance) at the positive input to reduce the impedance to high frequencies and effectively shunt them to ground.

2.3 NOISE

2.3.1 Introduction

The noise performance of IC amplifiers is determined by four primary noise sources: thermal noise, shot noise, 1/f, and popcorn noise. These four sources of noise are briefly discussed. Their contribution to overall noise performance is represented by equivalent input generators. In addition to these equivalent input generators, the effects of feedback and frequency compensation on noise are also examined. The noise behavior of the differential amplifier is noted since most op amps today use a differential pair. Finally noise measurement techniques are presented.

2.3.2 Thermal Noise

Thermal noise is generated by any passive resistive element. This noise is "white," meaning it has a constant spectral density. Thermal noise can be represented by a mean-square voltage generator $\overline{e_R^2}$ in series with a noiseless resistor, where $\overline{e_R^2}$ is given by Equation (2.3.1).

$$\overline{e_R^2} = 4k\,TRB \text{ (volts)}^2$$

where: T = temperature in °K

R = resistor value in ohms

B = noise bandwidth in Hz

k = Boltzmann's constant (1.38×10^{-23} W-sec/°K)

The RMS value of Equation (2.3.1) is plotted in Figure 2.3.1 for a one Hz bandwidth. If the bandwidth is increased, the plot is still valid so long as e_R is multiplied by $\sqrt{B}$.

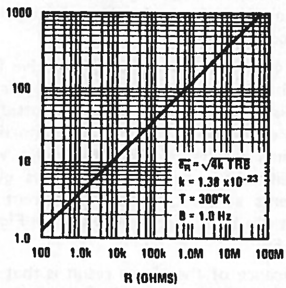


FIGURE 2.3.1 Thermal Noise of Resistor

Actual resistor noise measurements may have more noise than shown in Figure 2.3.1. This additional noise component is known as *excess noise*. Excess noise has a 1/f spectral response, and is proportional to the voltage drop across the resistor. It is convenient to define a *noise index* when referring to excess noise in resistors. *The noise index is the RMS value in µV of noise in the resistor per volt of DC drop across the resistor in a decade of frequency.* Noise index expressed in dB is:

$$NI = 20 \log\left(\frac{E_{ex}}{V_{DC}} \times 10^6\right) \text{dB} \qquad (2.3.1)$$

where: E_{ex} = resistor excess noise in µV per frequency decade.

V_{DC} = DC voltage drop across the resistor.

Excess noise in carbon composition resistors corresponds to a large noise index of +10dB to −20dB. Carbon film resistors have a noise index of −10dB to −25dB. Metal film and wire wound resistors show the least amount of excess noise, with a noise index figure of −15dB to −40dB. For a complete discussion of excess noise see Reference 2.

2.3.3 Noise Bandwidth

Noise bandwidth is not the same as the common amplifier or transfer function −3dB bandwidth. Instead, noise bandwidth has a "brick-wall" filter response. The maximum power gain of a transfer function T(jω) multiplied by the noise bandwidth must equal the total noise which passes through the transfer function. Since the transfer function power gain is related to the square of its voltage gain we have:

$$(T_{MAX}{}^2)B = \int_0^\infty |T(j\omega)|^2 d\omega \qquad (2.3.2)$$

where: T_{MAX} = maximum value of T(jω)

T(jω) = transfer function voltage gain

B = noise bandwidth in Hz

For a single RC roll-off, the noise bandwidth B is $\pi/2\, f_{-3dB}$, and for higher order maximally flat filters, see Table 2.3.1.

TABLE 2.3.1 Noise Bandwidth Filter Order

Filter Order	Noise Bandwidth B
1	$1.57 f_{-3dB}$
2	$1.11 f_{-3dB}$
3	$1.05 f_{-3dB}$
4	$1.025 f_{-3dB}$
"Brick-wall"	$1.00 f_{-3dB}$

2.3.4 Shot Noise

Shot noise is generated by charge crossing a potential barrier. It is the dominant noise mechanism in transistors and op amps at medium and high frequencies. The mean square value of shot noise is given by:

$$\overline{I_S^2} = 2q\, I_{DC}\, B \text{ (amps)}^2 \qquad (2.3.3)$$

where: q = charge of an electron in coulombs

I_{DC} = direct current in amps

B = noise bandwidth in Hz

Like thermal noise, shot noise has a constant spectral density.

2

2.3.5 1/f Noise

1/f or flicker noise is similar to shot noise and thermal noise since its amplitude is random. Unlike thermal and shot noise, 1/f noise has a 1/f spectral density. This means that the noise increases at low frequencies. 1/f noise is caused by material and manufacturing imperfections, and is usually associated with a direct current:

$$\overline{I_f^2} = K \frac{(I_{DC})^a}{f} B \text{ (amps)}^2 \qquad (2.3.4)$$

where: I_{DC} = direct current in amps

K and a = constants

f = frequency in Hz

B = noise bandwidth in Hz

2.3.6 Popcorn Noise (PCN)

Popcorn noise derives its name from the popcorn-like sound made when connected to a loudspeaker. It is characterized by a sudden change in output DC level, lasting from milliseconds to seconds, recurring randomly. Although there is no clear explanation of PCN to date, it is usually reduced by cleaner processing (see Reference 5). Extensive testing techniques are used to screen for PCN units.

2.3.7 Modelling

Every element in an amplifier is a potential source of noise. Each transistor, for instance, shows all three of the above mentioned noise sources. The net effect is that noise sources are distributed throughout the amplifier, making analysis of amplifier noise extremely difficult. Consequently, amplifier noise is completely specified by a noise voltage and a noise current generator at the input of a noiseless amplifier. Such a model is shown in Figure 2.3.2. Correlation between generators is neglected unless otherwise noted.

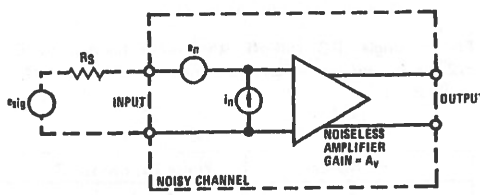


FIGURE 2.3.2 Noise Characterization of Amplifier

Noise voltage e_n, or more properly, equivalent short-circuit input RMS noise voltage, is simply that noise voltage which would appear to originate at the input of the noiseless amplifier if the input terminals were shorted. It is expressed in "nanovolts per root Hertz" (nV/$\sqrt{\text{Hz}}$) at a specified frequency, or in microvolts for a given frequency band. It is measured by shorting the input terminals, measuring the output RMS noise, dividing by amplifier gain, and referencing to the input – hence the term "equivalent input noise voltage." An output bandpass filter of known characteristic is used in measurements, and the measured value is divided by the square root of the bandwidth if data are to be expressed per unit bandwidth.

Figure 2.3.3 shows e_n of a typical op amp. For this amplifier, the region above 1 kHz is the shot noise region, and below 1 kHz is the amplifier's 1/f region.

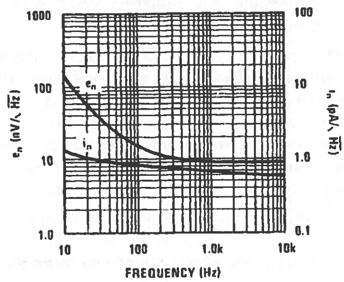


FIGURE 2.3.3 Noise Voltage and Current for an Op Amp

Noise Current, i_n, or more properly, equivalent open-circuit RMS noise current, is that noise which occurs apparently at the input of the noiseless amplifier due only to noise currents. It is expressed in "picoamps per root Hertz" (pA/$\sqrt{\text{Hz}}$) at a specified frequency or in nanoamps in a given frequency band. It is measured by shunting a capacitor or resistor across the input terminals such that the noise current will give rise to an additional noise voltage which is $i_n \times R_{in}$ (or XC_{in}). The output is measured, divided by amplifier gain, and that contribution known to be due to e_n and resistor noise is appropriately subtracted from the total measured noise. If a capacitor is used at the input, there is only e_n and $i_n XC_{in}$. The i_n is measured with a bandpass filter and converted to pA/$\sqrt{\text{Hz}}$ if appropriate. Again, note the 1/f and shot noise regions of Figure 2.3.3.

Now we can examine the relationship between e_n and i_n at the amplifier input. When the signal source is connected, the e_n appears in series with the e_{sig} and e_R. The i_n flows through R_s, thus producing another noise voltage of value $i_n \times R_s$. This noise voltage is clearly dependent upon the value of R_s. All of these noise voltages add at the input of Figure 2.3.2 in RMS fashion, that is, as the square root of the sum of the squares. Thus, neglecting possible correlation between e_n and i_n, the total input noise is:

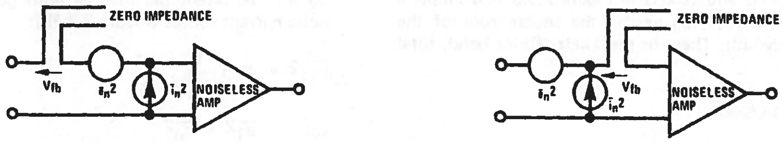
$$\overline{e_N^2} = \overline{e_n^2} + \overline{e_R^2} + \overline{i_n^2}\, R_s^2 \qquad (2.3.5)$$

2.3.8 Effects of Ideal Feedback on Noise

Extensive use of voltage and current feedback are common in op amp circuits today. Figures 2.3.4a and 2.3.4b can be used to show the effect of voltage feedback on the noise performance of an op amp.

Figure 2.3.4a shows application of negative feedback to an op amp with generators $\overline{e_n^2}$ and $\overline{i_n^2}$. Figure 2.3.4b shows that the noise generators can be moved outside the feedback loop. This operation is possible since shorting both amplifiers' inputs results in the same noise voltage at the outputs. Likewise, opening both inputs gives the same noise currents at the outputs. For current feedback, the same result can be found. This is seen in Figure 2.3.5a and Figure 2.3.5b.

The significance of the above result is that the equivalent input noise generators completely specify circuit noise. *The application of ideal negative feedback does not alter the noise performance of the circuit.* Feedback reduces the output noise, but it also reduces the output signal. *In other words, with ideal feedback, the equivalent input noise is independent of gain.*

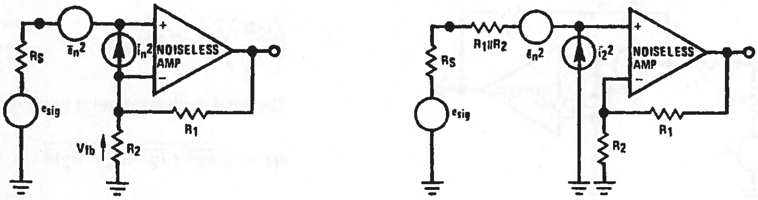


(a) Feedback Applied to Op Amp with Noise Generators **(b) Noise Generators Outside Feedback Loop**

FIGURE 2.3.4

(a) Current Feedback Applied to Op Amp **(b) Noise Generators Moved Outside Feedback Loop**

FIGURE 2.3.5



(a) Practical Voltage Feedback Amplifier **(b) Voltage Feedback with Noise Generators Moved Outside Feedback Loop**

FIGURE 2.3.6

2.3.9 Effects of Practical Feedback on Noise

Voltage feedback is implemented by series-shunt feedback as shown in Figure 2.3.6a.

The noise generators can be moved outside the feedback loop as shown in Figure 2.3.6b if the thermal noise of $R_1 \| R_2$ is included in $\overline{e_N^2}$. In addition, the noise generated by $i_n \times (R_1 \| R_2)$ must be added even though the (–) input is a virtual ground (see Appendix 7). The above effects can be easily included if $R_1 \| R_2$ is considered to be in series with R_s.

$$\overline{e_N^2} = \overline{e_n^2} + 4kT\,(R_s + R_1 \| R_2) + \overline{i_n^2}\,(R_s + R_1 \| R_2)^2$$

$$\overline{i_2^2} = \overline{i_n^2}$$

Example 2.3.1

Determine the total equivalent input noise per unit bandwidth for the amplifier of Figure 2.3.6a operating at 1kHz from a source resistance of 1kΩ. R_1 and R_2 are 100kΩ and 1kΩ respectively.

Solution:

Use data from Figure 2.3.1 and Figure 2.3.3.

1. Thermal noise from $R_s + R_1 \| R_2 \approx 2k$ is $5.65\text{nV}/\sqrt{\text{Hz}}$.
2. Read e_n from Figure 2.3.3 at 1kHz; this value is $9.5\text{nV}/\sqrt{\text{Hz}}$.
3. Read i_n from Figure 2.3.3 at 1kHz; this value is $0.68\text{pA}/\sqrt{\text{Hz}}$. Multiply this noise current by $R_s + R_1 \| R_2$ to obtain $1.36\text{nV}/\sqrt{\text{Hz}}$.
4. Square each term and enter into Equation (2.3.5).

$$e_N = \sqrt{\overline{e_n^2} + 4kT\,(R_s + R_1 \| R_2) + \overline{i_n^2}\,(R_s + R_1 \| R_2)^2}$$

$$e_N = \sqrt{(9.5)^2 + (5.65)^2 + (1.36)^2}$$

$$e_N = 11.1\text{nV}/\sqrt{\text{Hz}}$$

This is total RMS noise at the input in one Hertz bandwidth at 1kHz. If total noise in a given bandwidth is desired, one must integrate the noise over a bandwidth as specified. This is most easily done in a noise measurement set-up, but may be approximated as follows:

1. If the frequency range of interest is in the flat band, i.e., between 1kHz and 10kHz in Figure 2.3.3, it is simply a matter of multiplying e_N by the square root of the noise bandwidth. Then, in the 1kHz-10kHz band, total noise is:

$$e_N = 11.1\sqrt{9000}$$
$$= 1.05\mu V$$

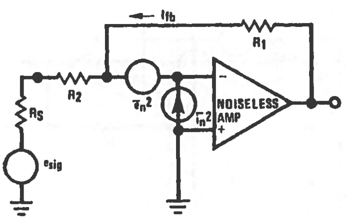
2. If the frequency band of interest is not in the flat band of Figure 2.3.3, one must break the band into sections, calculating average noise in each section, squaring, multiplying by section bandwidth, summing all sections, and finally taking square root of the sum as follows:

$$e_N = \sqrt{\overline{e_R}^2 B + \sum_{1}^{i} \left(\overline{e_N}^2 + \overline{i_n}^2 (R_s + R_1 || R_2)^2\right)_i B_i} \qquad (2.3.6)$$

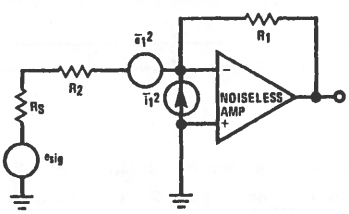
where: i is the total number of sub-blocks

For details and examples of this type of calculation, see application note AN-104, "Noise Specs Confusing?"

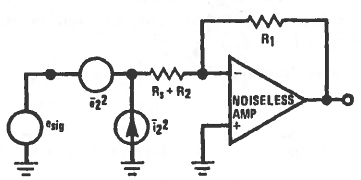
Current feedback is accomplished by shunt-shunt feedback as shown in Figure 2.3.7a.



(a) Practical Current Feedback Amplifier



(b) Intermediate Move of Noise Generators



(c) Current Feedback with Noise Generators Moved Outside Feedback Loop

FIGURE 2.3.7

$\overline{e_n}^2$ and $\overline{i_n}^2$ can be moved outside the feedback loop if the noise generated by R_1 and R_2 are taken into account. First, move the noise generators outside feedback R_1. To do this, represent the thermal noise generated by R_1 as a noise current source (Figure 2.3.7b):

$$\overline{i_{R1}}^2 = 4k\,T\,\frac{1}{R_1}$$

so: $$\overline{e_1}^2 = \overline{e_n}^2$$

and: $$\overline{i_1}^2 = \overline{i_n}^2 + 4k\,T\,\frac{1}{R_1}$$

Now move these noise generators outside $R_s + R_2$ as shown in Figure 2.3.7c to obtain $\overline{e_2}^2$ and $\overline{i_2}^2$:

$$\overline{e_2}^2 = \overline{e_n}^2 + 4k\,T\,(R_s + R_2) \qquad (2.3.7)$$

$$\overline{i_2}^2 = \overline{i_n}^2 + 4k\,T\,\frac{1}{R_1} \qquad (2.3.8)$$

$\overline{e_2}^2$ and $\overline{i_2}^2$ are the equivalent input generators with feedback applied. The total equivalent input noise, e_N, is the sum of the noise produced with the input shorted, and the noise produced with the input opened. With the input of Figure 2.3.7c shorted, the input referred noise is $\overline{e_2}^2$. With the input opened, the input referred noise is:

$$\left(\frac{i_2 R_1}{A_V}\right)^2 = \overline{i_2}^2 (R_s + R_2)^2$$

The total equivalent input noise is:

$$e_N = \sqrt{\overline{e_2}^2 + \overline{i_2}^2 (R_s + R_2)^2}$$

Example 2.3.2

Determine the total equivalent input noise per unit bandwidth for the amp of Figure 2.3.7a operating at 1kHz from a 1kΩ source. Assume R_1 is 100kΩ and R_2 is 9kΩ.

Solution

Use data from Figures 2.3.1 and 2.3.3.

1. Thermal noise from $R_s + R_2$ is $12.7\text{nV}/\sqrt{\text{Hz}}$.
2. Read e_n from figure 2.3.3 at 1kHz; this value is $9.5\text{nV}/\sqrt{\text{Hz}}$. Enter these values into Equation (2.3.7).
3. Determine the thermal noise current contributed by R_1:

$$i_{R1} = \sqrt{4k\,T\,\frac{1}{R_1}\,B} = \sqrt{\frac{1.61 \times 10^{-20}}{100k}} = 0.401\text{pA}/\sqrt{\text{Hz}}$$

4. Read i_n from Figure 2.3.3 at 1kHz; this value is $0.68\text{pA}/\sqrt{\text{Hz}}$. Enter these values into Equation (2.3.7).

$$e_N = \sqrt{\overline{e_n}^2 + (R_s + R_2)^2 \left(\overline{i_n}^2 + 4k\,T\,\frac{1}{R_1}\right) + 4k\,T\,(R_s + R_2)}$$

$$e_N = \sqrt{(9.5)^2 + (10k)^2 (0.68^2 + 0.401^2) + (12.7)^2}\ \text{nV}/\sqrt{\text{Hz}}$$

$$e_N = 17.7\text{nV}/\sqrt{\text{Hz}}$$

For the noise in the bandwidth from 1kHz to 10kHz, $e_N = 17.7\text{nV}\sqrt{9000} = 1.68\mu V$. If the noise is not constant with frequency, the method shown in Equation (2.3.6) should be used.

TABLE 2.3.2 Equivalent Input Noise Comparison

NON-INVERTING AMPLIFIER					INVERTING AMPLIFIER				
A_V	R_s	R_1	R_2	e_N (nV$\sqrt{Hz}$)	A_V	R_s	R_1	R_2	e_N (nV$\sqrt{Hz}$)
101	1k	100k	1k	11.1	100	1k	100k	0	10.3
11	1k	100k	10k	17.3	10	1k	100k	9k	17.7
2	1k	100k	100k	46.0	2	1k	100k	49k	49.5
1	1k	100k	∞	80.2	1	1k	100k	99k	89.1

Example 2.3.3

Compare the noise performance of the non-inverting amplifier of Figure 2.3.6a to the inverting amplifier of Figure 2.3.7a.

Solution:

The best way to proceed here is to make a table and compare the noise performance with various gains.

Table 2.3.2 shows only a small difference in equivalent input noise for the two amplifiers. There is, however, a large difference in the flexibility of the two amplifiers. The gain of the inverting amplifier is a function of its input resistance, R_2. Thus, for a given gain and input resistance, R_1 is fixed. This is not the case for the non-inverting amplifier. The designer is free to pick R_1 and R_2 independent of the amplifier's input impedance. Thus in the case of unity gain, where $R_2 = \infty$, R_1 can be zero ohms. The equivalent input noise is:

$$e_N = \sqrt{e_n^2 + 4kTR_s + i_r^2 R_s^2}$$

$$e_N = 10.3\text{nV}/\sqrt{\text{Hz}}$$

There is now a large difference in the noise performance of the two amplifiers. Table 2.3.2 also shows that *the equivalent input noise for practical feedback can change as a function of closed loop gain A_V. This result is somewhat different from the case of ideal feedback.*

Example 2.3.4

Determine the signal-to-noise ratio for the amplifier of Example 2.3.2 if e_{SIG} has a nominal value of 100mV.

Solution:

Signal to noise ratio is defined as:

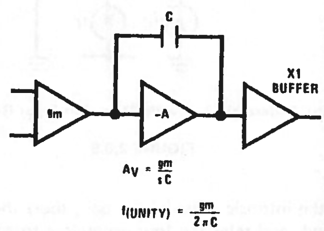
$$S/N = 20\log\frac{e_{SIG}}{e_N} \qquad (2.3.9)$$

$$= 20\log\frac{100\text{mV}}{1.68\mu\text{V}} = 95.5\text{dB}$$

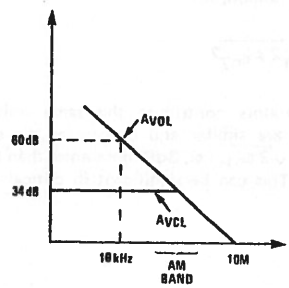
2.3.10 RF Precautions

A source of potential RF interference that needs to be considered in AM radio applications lies in the radiated wideband noise voltage developed at the speaker terminals. The method of amplifier compensation (Figure 2.3.8a) fixes the point of unity gain cross at approximately 10MHz (Figure 2.3.8b). A wideband design is essential in achieving low distortion performance at high audio frequencies, since it allows adequate loop-gain to reduce THD. (Figure 2.3.8b shows that for a closed-loop gain of 34dB there still exists 26dB of loop-gain at 10kHz.)

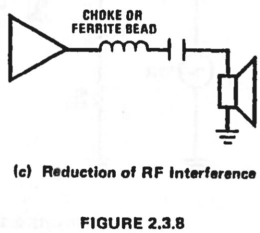
The undesirable consequence of a single-pole roll-off, wideband design is the excess gain beyond audio frequencies, which includes the AM band; hence, noise of this frequency is amplified and delivered to the load where it can radiate back to the AM (magnetic) antenna and sensitive RF circuits. A simple and economical remedy is shown in Figure 2.3.8c, where a ferrite bead, or small RF choke is added in series with the output lead. Experiments have demonstrated that this is an effective method in suppressing the unwanted RF signals.



(a) Typical Compensation



(b) Source of RF Interference



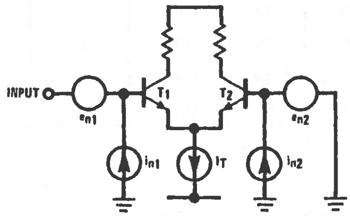
(c) Reduction of RF Interference

FIGURE 2.3.8

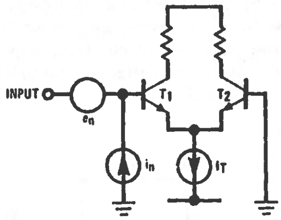
2

2.3.11 Noise in the Differential Pair

Figure 2.3.9a shows a differential amplifier with noise generators e_{n1}, i_{n1}, e_{n2}, and i_{n2}.



(a) Differential Pair with Noise Generators



(b) Differential Pair with Generators Input Referred

FIGURE 2.3.9

To see the intrinsic noise of the pair, short the base of T_2 to ground, and refer the four generators to an input noise voltage and noise current as shown in Figure 2.3.9b. To determine e_n, short the input of 9(a) and 9(b) to ground. e_n is then the series combination of e_{n1} and e_{n2}. These add in an RMS fashion, so:

$$e_n = \sqrt{e_{n1}^2 + e_{n2}^2}$$

Both generators contribute the same noise, since the transistors are similar and operate at the same current; thus, $e_n = \sqrt{2}\, e_{n1}$, i.e., 3dB more noise than a single ended amplifier. This can be significant in critical noise applications.

In order to find the input noise current generator, i_n, open the input and equate the output noise from Figure 2.3.9a and Figure 2.3.9b. The result of this operation is $i_n = i_{n1}$. Thus, from a high impedance source, the differential pair gives similar noise current as a single transistor.

2.3.12 Noise Measurement Techniques

This section presents techniques for measuring e_n, i_n, and e_N. The method can be used to determine the spectral density of noise, or the noise in a given bandwidth. The circuit for measuring the noise of an LM387 is shown in Figure 2.3.10.

The system gain, V_{OUT}/e_n, of the circuit in Figure 2.3.10 is large – 80dB. This large gain is required since we are trying to measure *input referred noise generators* on the order of $5nV/\sqrt{Hz}$, which corresponds to $50\mu V/\sqrt{Hz}$ at the output. R_1 and R_2 form a 100:1 attenuator to provide a low input signal for measuring the system gain. The gain should be measured in both the e_n and i_n positions, since LM387 has a 250k bias resistor which is between input and ground. The LM387 of Figure 2.3.10 has a closed loop gain of 40dB which is set by feedback elements R_5 and R_6. 40dB provides adequate gain for the input referred generators of the LM387. The output noise of the LM387 is large compared to the input referred generators of the LM381; consequently, noise at the output of the LM381 will be due to the LM387. To measure the noise voltage e_n, and noise current $i_n \times R_3$, a wave analyzer or noise filter set is connected. In addition the noise in a given bandwidth can be measured by using a bandpass filter and an RMS voltmeter. If a true RMS voltmeter is not available, an average responding meter works well. When using an average responding meter, the measured noise must be multiplied by 1.13 since the meter is calibrated to measure RMS *sine waves*. The meter used for measuring noise should have a crest factor (ratio of peak to RMS value) from 3 to 5, as the peak to RMS ratio of noise is on that order. Thus, if an average responding meter measures 1mV of noise, the RMS value would be $1.13mV_{RMS}$, and the peak-to-peak value observed on an oscilloscope could be as high as 11.3mV (1.13mV x 2 x 5).

Some construction tips for the circuit of Figure 2.3.10 are as follows:

1. R_4 and R_6 should be metal film resistors, as they exhibit lower excess noise than carbon film resistors.
2. C_1 should be large, to provide low capacitive reactance at low frequency, in order to accurately observe the 1/f noise in e_n.

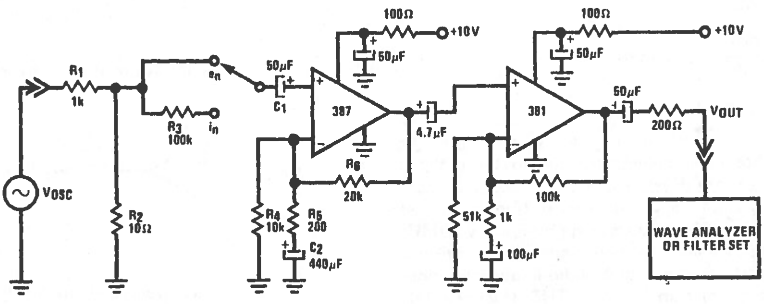


FIGURE 2.3.10 Noise Test Setup for Measuring e_n and i_n of an LM387

3. C_2 should be large to maintain the gain of 80dB down to low frequencies for accurate 1/f measurements.
4. The circuit should be built in a small grounded metal box to eliminate hum and noise pick-up, especially in i_n.
5. The LM387 and LM381 should be separated by a metal divider within the metal box. This is to prevent output to input oscillations.

Typical LM387 noise voltage and noise current are plotted in Figure 2.3.11.

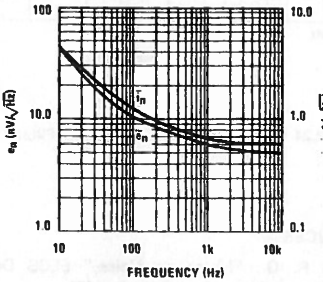


FIGURE 2.3.11 LM387 Noise Voltage and Noise Current

Many times we do not care about the actual spectral distribution of noise, rather we want to know the noise voltage in a given bandwidth for comparison purposes. For audio frequencies, we are interested only in a 20kHz bandwidth. The noise voltage is often the dominant noise source since many systems use a low impedance voltage drive as the signal. For this common case we use a test set-up as shown in Figure 2.3.12.

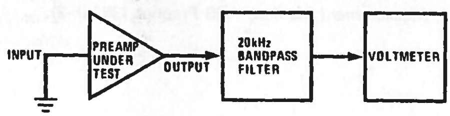


FIGURE 2.3.12 Test Setup for Measuring Equivalent Input Noise for a 20 kHz Bandwidth

Example 2.3.5

Determine the equivalent input noise voltage for the preamp of Figure 2.3.12. The gain, A_V, of the preamp is 40dB and the voltmeter reads 0.2mV. Assume the voltmeter is average responding and the 20kHz low-pass filter has a single R-C roll-off.

Solution:

Since the voltmeter is average responding, the RMS voltage is $V_{RMS} = 0.2\text{mV} \times 1.13 = 0.226\text{mV}$. Using an average responding meter causes only a 13% error. The filter has a single R-C roll-off, so the noise bandwidth is $\pi/2 \times 20\text{kHz} = 31.4\text{kHz}$, i.e., the true noise bandwidth is 31.4kHz and not 20kHz. Since RMS noise is related to the square root of the noise bandwidth, we can correct for this difference:

$$V_{OUT} = \sqrt{\frac{0.226}{\pi/2}} = 0.18\text{mV}$$

The equivalent input noise is:

$$\frac{V_{OUT}}{A_V} = \frac{0.18\text{mV}}{100} = 1.8\mu\text{V in a 20kHz bandwidth.}$$

If this preamp had RIAA playback equalization, the output noise, V_{OUT}, would have been divided by the gain at 1kHz.

Typical values of noise, measured by the technique of Figure 2.3.12, are shown in Table 2.3.3. For this data, $B = 10\text{kHz}$ and $R_s = 600\Omega$.

TABLE 2.3.3 Typical Flat Band Equivalent Input Noise

Type	e_N (μV)
LM381	0.70
LM381A	0.50
LM382	0.80
LM387	0.80
LM387A	0.65

2.3.13 Noise Measurement for Consumer Audio Equipment — The Use of Weighting Filters

The previous discussion of noise and its measurement has been mainly concerned with obtaining a noise voltage "number" over a given frequency bandwidth in order to provide a S/N ratio for signals that can occupy all or part of the same bandwidth. The usefulness of this is restricted by the fact that there is no indication from this "number" of the subjective annoyance of noise spectra present within this bandwidth of interest. For example, two systems with measured identical signal/noise ratios can sound very different because one may have a uniform distribution of noise spectra whereas the other may have most of the noise concentrated in one particular portion of the frequency band. The total noise voltage is the same in each case but the audible effect is that one system sounds "noisier" than the other.

To understand why this should be, we need to investigate in a little more detail the relative sensitivity of the human ear and the effects of auditory masking phenomena. Readers familiar with the Fletcher-Munson equal loudness contours (Section 2.14.7) and the more recent work by Robinson and Dadson[6] will already know that the ear is not uniformly sensitive to all frequencies in the audio band, an effect that is emphasized at extremely low sound levels. Further, in a steady state condition, the threshold of hearing for a given tone is changed by the presence of another tone (the masker). The amount of change is dependent on the relative pitch and loudness of the masker and the maskee. Noise will also raise the threshold of hearing for tones — i.e. the tone has to be louder to be heard if noise is also present in some part of the frequency band. Figure 2.12.23 is a plot of the hearing threshold of acute ears for noise in a typical home environment (noise spectra below this curve are inaudible). Below 200Hz and above 6kHz the shape of this curve is caused by the hearing mechanism, and between 200Hz and 6kHz is caused by the masking effect of room noise. This means that if noise is just audible at 1kHz, the amplitude of noise at 100Hz has to be 30db higher to be equally audible. A further complication is that the audibility of the noise is not necessarily indicative of its obtrusiveness or annoyance.

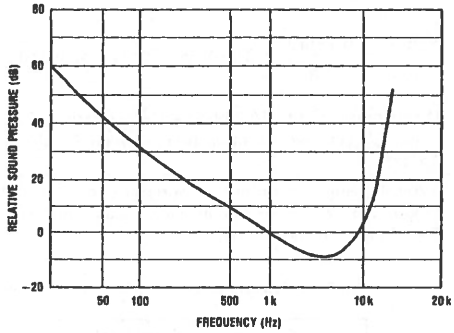


Figure 2.12.23 Threshold of Hearing for Noise in the Home Environment

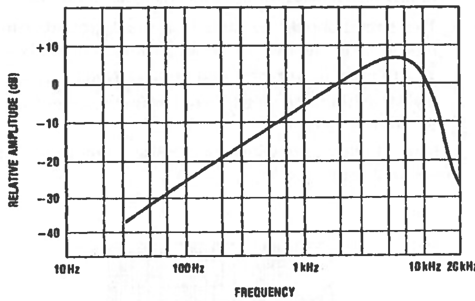


Figure 2.12.24 CCIR/ARM Noise Weighting Filter Characteristic

To make comparative S/N measurements more meaningful, several filters have been used to weight the contribution of the noise spectra over the frequency band of interest (N.A.B. A-Weighting Curve, D.I.N. 45405 for example) so that the S/N numbers correlate better to the subjective impression gained in listening tests. Recently the CCIR adopted a weighting filter (Recommendation 468-1) with the characteristic shape shown in Figure 2.12.24 which is based on the obtrusiveness as well as the level of different kinds of noise. While this filter is normally used with a quasi-peak reading meter to derive consistent readings with all types of noise (including clicks, pops and whistles as well as broad spectrum noise), for typical audio equipment such as tape decks and amplifiers, an average responding meter has been found to give equally consistent results.[7]

The filter characteristic of Figure 2.12.24 is known as the CCIR/ARM filter and is currently used by Dolby Laboratories for measurements on their Dolby® B-Type noise reduction units. Note that the 0dB reference frequency is 2kHz instead of the more conventional 1kHz — so that S/N ratios obtained by this method are numerically close to the S/N ratios obtained by earlier methods and which are considered commercially acceptable for the quality of equipment being measured. Without the reference frequency shift the S/N ratios obtained with the CCIR/ARM filter would be several dB below the expected number.

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2.4 AUDIO RECTIFICATION
Or, "How Come My Phono Detects AM?"

Audio rectification refers to the phenomenon of RF signals being picked up, rectified, and amplified by audio circuits – notably by high-gain preamplifiers. Of all types of interference possible to plague a hi-fi system, audio rectification remains the most slippery and troublesome. A common occurrence of audio rectification is to turn on a phonograph and discover you are listening to your local AM radio station instead. There exist four main sources of interference, each with a unique character: If it is clearly audible through the speaker then AM radio stations are probably the source; if the interference is audible but garbled then suspect SSB and amateur radio equipment; a decrease in volume can be produced by FM pickup; and if buzzing occurs, then RADAR or TV is being received. Whatever the source, the approaches to eliminating it are similar.

Commonly, the rectification occurs at the first non-linear, high gain, wide bandwidth transistor encountered by the incoming signal. The signal may travel in unshielded or improperly grounded input cables; it may be picked up through the air by long, poorly routed wires; or it may enter on the AC power lines. It is rectified by the first stage transistor acting as a detector diode, subsequently amplified by the remaining circuitry, and finally delivered to the speaker. Bad solder joints can detect the RF just well as transistors and must be avoided (or suspected). The following list should be consulted when seeking to eliminate audio rectification from existing equipment. For new designs, keep input leads short and shielded, with the shield grounded only at one point; make good clean solder connections; avoid loops created by multiple ground points; and make ground connections close to the IC or transistor that they associate with.

Audio Rectification Elimination Tips (Figure 2.4.1).

- Reduce input impedance.
- Place capacitor to ground close to input pin or base (~ 10-300pF).
- Use ceramic capacitors.
- Put ferrite bead on input lead close to the device input.
- Use RF choke in series with input (~ 10μH).
- Use RF choke (or ferrite bead) *and* capacitor to ground.
- Pray.

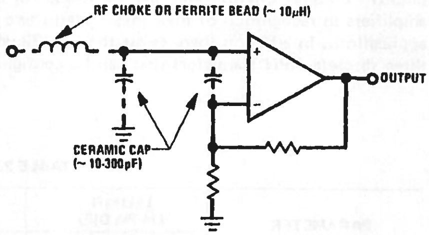


FIGURE 2.4.1 Audio Rectification Elimination Tips

A particularly successful technique is uniquely possible with the LM381 since both base and emitter points of the input transistor are available. A ceramic capacitor is mounted very close to the IC from pin 1 to pin 3, shorting base to emitter at RF frequencies (see Figure 2.4.2).

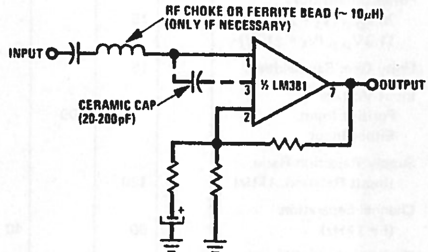


FIGURE 2.4.2 LM381 Audio Rectification Correction

2.5 DUAL PREAMPLIFIER SELECTION

National Semiconductor's line of integrated circuits designed specifically to be used as audio preamplifiers consists of the LM381, LM382, LM387, and the LM1303. All are dual amplifiers in recognition of their major use in two channel applications. In addition there exists the LM389 which has three discrete NPN transistors that can be configured into a low noise monaural preamplifier for minimum parts count mono systems (Section 4.11). Table 2.5.1 shows the major electrical characteristics of each of the dual preamps offered. A detailed description of each amplifier follows, where the individual traits and operating requirements are presented.

TABLE 2.5.1 Dual Preamplifier Characteristics

PARAMETER	LM381N (14 Pin DIP)			LM382N (14 Pin DIP)			LM387N (8 Pin DIP)			LM1303N (14 Pin DIP)			UNITS
	MIN	TYP	MAX	MIN	TYP	MAX	MIN	TYP	MAX	MIN	TYP	MAX	
Supply Voltage	9		40	9		40	9		30/40[6]	±4.5		±15	V
Quiescent Supply Current		10			10	16		10				15	mA
Input Resistance (open loop)													
Positive Input		100k			100k		50k	100k			25k		Ω
Negative Input		200k			200k			200k			25k		Ω
Open Loop Gain		104			100			104		76	80		dB
Output Voltage Swing $R_L = 10k\Omega$		$V_s - 2$			$V_s - 2$			$V_s - 2$		11.3	15.6		$V_{p\text{-}p}$
Output Current													
Source		8[2]			8[2]			8[2]		0.6	0.8		mA
Sink		2			2			2		0.6	0.8		mA
Output Resistance (open loop)		150			150			150			4k		Ω
Slew Rate (A_v = 40dB)		4.7			4.7			4.7			5.0[7]		V/μs
Power Bandwidth													
$20V_{p\text{-}p}$ (V_s = 24V)		75			75			75					kHz
$11.3V_{p\text{-}p}$ (V_s = ±13V)											100		kHz
Unity Gain Bandwidth		15			15			15			20		MHz
Input Voltage													
Positive Input			300			300			300				mV_{RMS}
Either Input												±5	V
Supply Rejection Ratio (Input Referred, 1kHz)		120			120			110					dB
Channel Separation (f = 1kHz)		60		40	60		40	60		60	70		dB
Total Harmonic Distortion (f = 1kHz)[3]		0.1			0.1	0.3		0.1	0.5		0.1		%
Total Equivalent Input Noise		0.5[4]	1.0[4]		0.8	1.2		0.8	1.2				μV_{RMS}
(R_s = 600Ω, 10-10k Hz)		0.5[4,5]	0.7[4,5]					0.65[6]	0.9[6]				μV_{RMS}
Total NAB[8] Output Noise		190						230					μV_{RMS}
(R_s = 600Ω, 10-10k Hz)		140[5]						180[6]					μV_{RMS}

1. Specifications apply for T_A = 25°C with V_s = +14V for LM381/382/387 and V_s = ±13V for LM1303, unless otherwise noted.
2. DC current; symmetrical AC current = $2mA_{p\text{-}p}$.
3. LM381 & LM387: Gain = 60dB; LM382: Gain = 60dB; LM1303: Gain = 40dB.
4. Single ended input biasing.
5. LM381AN.
6. LM387AN.
7. Frequency Compensation: C = 0.0047μF, Pins 3 to 4.
8. NAB reference level: 37dBV Gain at 1kHz. Tape Playback Circuit.

2.6 LM381 LOW NOISE DUAL PREAMPLIFIER

2.6.1 Introduction

The LM381 is a dual preamplifier expressly designed to meet the requirements of amplifying low level signals in low noise applications. Total equivalent input noise is typically $0.5\mu V_{RMS}$ ($R_s = 600\Omega$, 10-10,000 Hz).

Each of the two amplifiers is completely independent, with an internal power supply decoupler-regulator, providing 120dB supply rejection and 60dB channel separation. Other outstanding features include high gain (112dB), large output voltage swing (V_{CC} – 2V) p-p, and wide power bandwidth (75kHz, $20V_{p\text{-}p}$). The LM381 operates from a single supply across the wide range of 9 to 40V. The amplifier is internally compensated and short-circuit protected.

Attempts have been made to fill this function with selected operational amplifiers. However, due to the many special requirements of this application, these recharacterizations have not adequately met the need.

With the low output level of magnetic tape heads and phonograph cartridges, amplifier noise becomes critical in achieving an acceptable signal-to-noise ratio. This is a major deficiency of the op amp in this application. Other inadequacies of the op amp are insufficient power supply rejection, limited small-signal and power bandwidths, and excessive external components.

2.6.2 Circuit Description

To achieve low noise performance, special consideration must be taken in the design of the input stage. First, the input should be capable of being operated single ended, since both transistors contribute noise in a differential stage degrading input noise by the factor $\sqrt{2}$. (See Section 2.3.) Secondly, both the load and biasing elements must be resistive, since active components would each contribute as much noise as the input device.

The basic input stage, Figure 2.6.1, can operate as a differential or single ended amplifier. For optimum noise performance Q_2 is turned OFF and feedback is brought to the emitter of Q_1.

In applications where noise is less critical, Q_1 and Q_2 can be used in the differential configuration. This has the advantage of higher impedance at the feedback summing point, allowing the use of larger resistors and smaller capacitors in the tone control and equalization networks.

The voltage gain of the single ended input stage is given by:

$$A_{V(AC)} = \frac{R_L}{r_e} = \frac{200k}{1.25k} = 160 \quad (2.6.1)$$

where: $r_e = \dfrac{KT}{q\, I_E} \approx 1.25 \times 10^3$ at 25°C, $I_E \approx 20\mu A$

The voltage gain of the differential input stage is:

$$A_V = \frac{1}{2}\frac{R_L}{r_e} = \frac{1}{2}\frac{R_L\, q\, I_E}{KT} \approx 80 \quad (2.6.2)$$

The schematic diagram of the LM381, Figure 2.6.2, is divided into separate groups by function – first and second voltage gain stages, third current gain stage, and the bias regulator.

The second stage is a common-emitter amplifier (Q_5) with a current source load (Q_6). The Darlington emitter-follower

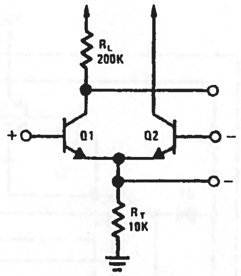


FIGURE 2.6.1 Input Stage

Q_3, Q_4 provides level shifting and current gain to the common-emitter stage (Q_5) and the output current sink (Q_7). The voltage gain of the second stage is approximately 2,000, making the total gain of the amplifier typically 160,000 in the differential input configuration.

The preamplifier is internally compensated with the pole-splitting capacitor, C_1. This compensates to unity gain at 15MHz. The compensation is adequate to preserve stability to a closed loop gain of 10. Compensation for unity gain closure may be provided with the addition of an external capacitor in parallel with C_1 between pins 5 and 6, 10 and 11.

Three basic compensation schemes are possible for this amplifier: first stage pole, second stage pole and pole-splitting. First stage compensation will cause an increase in high frequency noise because the first stage gain is reduced, allowing the second stage to contribute noise. Second stage compensation causes poor slew rate (power bandwidth) because the capacitor must swing the full output voltage. Pole-splitting overcomes both these deficiencies and has the advantage that a small monolithic compensation capacitor can be used.

The output stage is a Darlington emitter-follower (Q_8, Q_9) with an active current sink (Q_7). Transistor Q_{10} provides short-circuit protection by limiting the output to 12mA.

The biasing reference is a zener diode (Z_2) driven from a constant current source (Q_{11}). Supply decoupling is the ratio of the current source impedance to the zener impedance. To achieve the high current source impedance necessary for 120dB supply rejection, a cascade configuration is used (Q_{11} and Q_{12}). The reference voltage is used to power the first stages of the amplifier through emitter-followers Q_{14} and Q_{15}. Resistor R_1 and zener Z_1 provide the starting mechanism for the regulator. After starting, zero volts appears across D_1, taking it out of conduction.

2.6.3 Biasing

Figure 2.6.3 shows an AC equivalent circuit of the LM381. The non-inverting input, Q_1, is referenced to a voltage source two V_{BE} above ground. The output quiescent point is established by negative DC feedback through the external divider R_4/R_5 (Figure 2.6.4).

For bias stability, the current through R_5 is made ten times the input current of Q_2 ($\approx 0.5\mu A$). Then, for the differential input, resistors R_5 and R_4 are:

$$R_5 = \frac{2\,V_{BE}}{10\, I_{Q2}} = \frac{1.3}{5 \times 10^{-6}} = 260k\Omega \text{ maximum} \quad (2.6.3)$$

$$R_4 = \left(\frac{V_{CC}}{2.6} - 1\right) R_5 \quad (2.6.4)$$

2

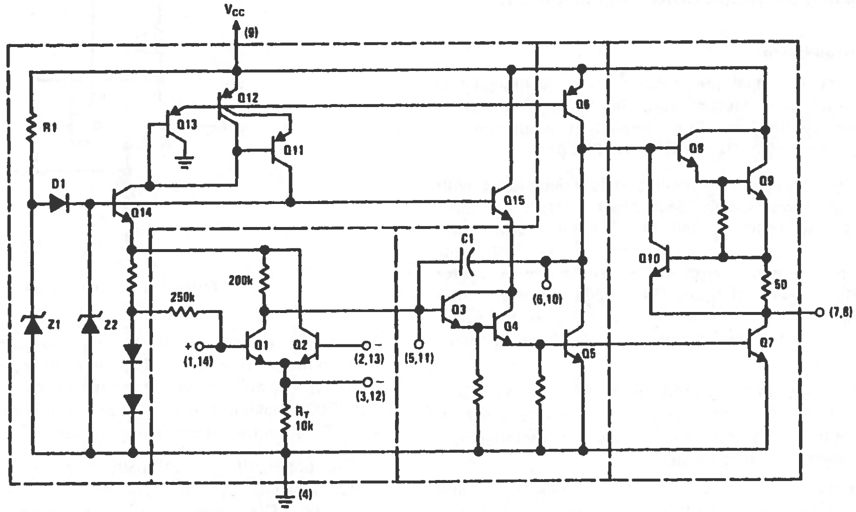


FIGURE 2.6.2 Schematic Diagram

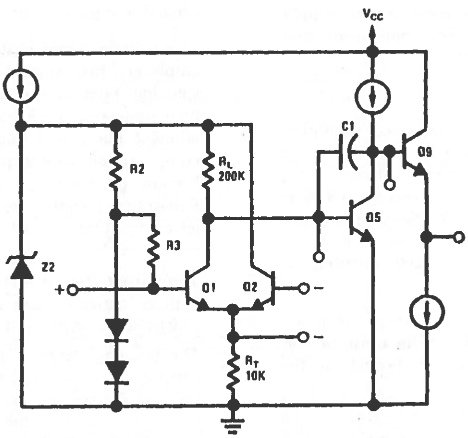


FIGURE 2.6.3 AC Equivalent Circuit

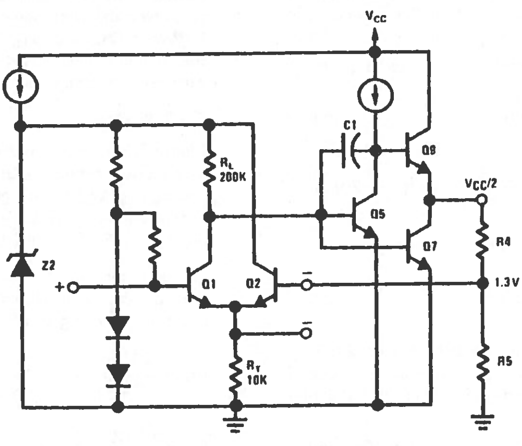


FIGURE 2.6.4 Differential Input Biasing

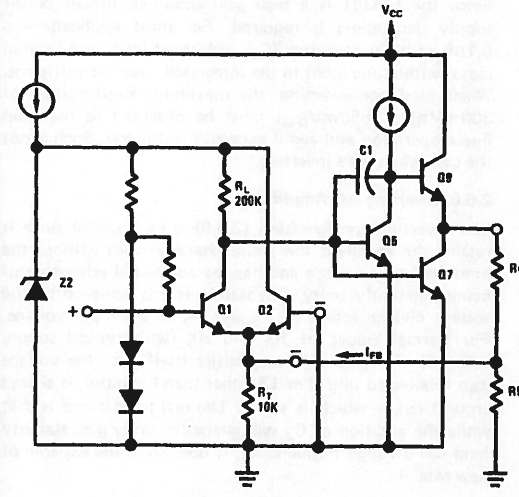


FIGURE 2.6.5 Single Ended Input Biasing

When using the single ended input, Q_2 is turned OFF and DC feedback is brought to the emitter of Q_1 (Figure 2.6.5). The impedance of the feedback summing point is now two orders of magnitude lower than the base of Q_2 ($\approx 10k\Omega$). Therefore, to preserve bias stability, the impedance of the feedback network must be decreased. In keeping with reasonable resistance values, the impedance of the feedback voltage source can be 1/5 the summing point impedance. The feedback current is $< 100\mu A$ worst case. Therefore, for single ended input, resistors R_5 and R_4 are:

$$R_5 = \frac{V_{BE}}{5\,I_{FB}} = \frac{0.65}{5 \times 10^{-4}} = 1300\Omega \text{ maximum} \qquad (2.6.5)$$

$$R_4 = \left(\frac{V_{CC}}{1.3} - 1\right) R_5 \qquad (2.6.6)$$

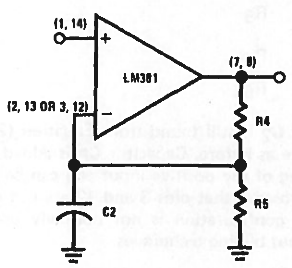


FIGURE 2.6.6 AC Open Loop

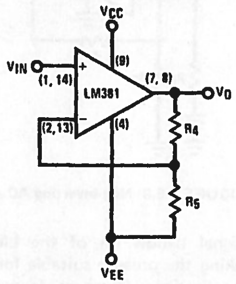
The circuits of Figures 2.6.4 and 2.6.5 have an AC and DC gain equal to the ratio R_4/R_5. To open the AC gain, capacitor C_2 is used to shunt R_5 (Figure 2.6.6). The AC gain now approaches open loop. The low frequency 3dB corner, f_o, is given by:

$$f_o = \frac{A_o}{2\pi C_2 R_4} \qquad (2.6.7)$$

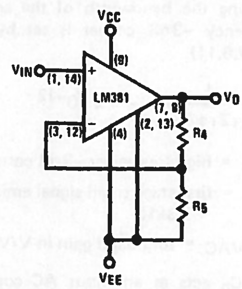
where: A_o = open loop gain

2.6.4 Split Supply Operation

Although designed for single supply operation, the LM381 may be operated from split supplies just as well. (A trade-off exists when unregulated negative supplies are used since the inputs are biased to the negative rail without supply rejection techniques and hum may be introduced.) All that is necessary is to apply the negative supply (V_{EE}) to the ground pin and return the biasing resistor R_5 to V_{EE} instead of ground. Equations (2.6.3) and (2.6.5) still hold, while the only change in Equations (2.6.4) and (2.6.6) is to recognize that V_{CC} represents the total potential across the LM381 and equals the absolute sum of the split supplies used, e.g., V_{CC} = 30 volts for ±15 volt supplies. Figure 2.6.7 shows a typical split supply application; both differential and single ended input biasing are shown. (Note that while the output DC voltage will be approximately zero volts the positive input DC potential is about 1.3 volts above the negative supply, necessitating capacitive coupling into the input.)



Differential Input Biasing



Single Ended Input Biasing

VODC ≈ 0 VOLTS
VINDC ≈ VEE + 1.2 VOLTS

FIGURE 2.6.7 Split Supply Operation

2

2.6.5 Non-Inverting AC Amplifier

Perhaps the most common application of the LM381 is as a flat gain, non-inverting AC amplifier operating from a single supply. Such a configuration is shown in Figure 2.6.8. Resistors R_4 and R_5 provide the necessary biasing and establish the DC gain, A_{VDC}, per Equation (2.6.8).

$$A_{VDC} = 1 + \frac{R_4}{R_5} \qquad (2.6.8)$$

AC gain is set by resistor R_6 with low frequency roll-off at f_o being determined by capacitor C_2.

$$A_{VAC} = 1 + \frac{R_4}{R_6} \; (R_6 \ll R_5) \qquad (2.6.9)$$

$$C_2 = \frac{1}{2\pi f_o R_6} \; (C_c R_L \gg C_2 R_6) \qquad (2.6.10)$$

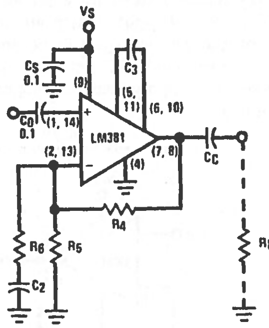


FIGURE 2.6.8 Non-inverting AC Amplifier

The small-signal bandwidth of the LM381 is nominally 20MHz, making the preamp suitable for wide-band instrumentation applications. However, in narrow-band applications it is desirable to limit the amplifier bandwidth and thus eliminate high frequency noise. Capacitor C_3 accomplishes this by shunting the internal pole-splitting capacitor (C_1), limiting the bandwidth of the amplifier. Thus, the high frequency −3dB corner is set by C_3 according to Equation (2.6.11).

$$C_3 = \frac{1}{2\pi f_3 2 r_e A_{VAC}} - 4 \times 10^{-12} \qquad (2.6.11)$$

where: f_3 = high frequency −3dB corner

r_e = first stage small-signal emitter resistance $\approx 1.3\text{k}\Omega$

A_{VAC} = mid-band gain in V/V

Capacitor C_o acts as an input AC coupling capacitor to block DC potentials in both directions and can equal 0.1µF (or larger). Output coupling capacitor C_c is determined by the load resistance and low frequency corner f per Equation (2.6.12).

$$C_c = \frac{1}{2\pi f R_L} \qquad (2.6.12)$$

Note: To avoid affecting f_o, $f \ll f_o$. For example, f = $0.25 f_o$ will cause a 0.25dB drop at f_o.

Since the LM381 is a high gain amplifier, proper power supply decoupling is required. For most applications a 0.1µF ceramic capacitor (C_s) with short leads and located close (within one inch) to the integrated circuit is sufficient. When used non-inverting, the maximum input voltage of 300mV_{RMS} ($850\text{mV}_{p\text{-}p}$) must be observed to maintain linear operation and avoid excessive distortion. Such is not the case when used inverting.

2.6.6 Inverting AC Amplifier

The inverting configuration (2.6.9) is very useful since it retains the excellent low noise characteristics without the limit on input voltage and has the additional advantage of being inherently unity gain stable. This is achieved by the voltage divider action of R_6 and R_5 on the input voltage. For normal values of R_4 and R_5 (with typical supply voltages) the gain of the amplifier itself, i.e., the voltage gain relative to pins 2 or 13 rather than the input, is always around ten — which is stable. The real importance is that while the addition of C_3 will guarantee unity gain stability (and roll-off high frequencies), it does so at the expense of slew rate.

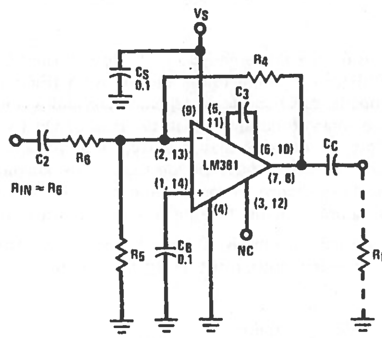


FIGURE 2.6.9 Inverting AC Amplifier

Using Figure 2.6.9 without C_3 at any gain retains the full slew rate of 4.7V/µs. The new gain equations follow:

$$A_{VDC} = -\frac{R_4}{R_5} \qquad (2.6.13)$$

$$A_{VAC} = -\frac{R_4}{R_6} \qquad (2.6.14)$$

Capacitor C_2 is still found from Equation (2.6.10), and C_c and C_s are as before. Capacitor C_B is added to provide AC decoupling of the positive input and can be made equal to 0.1µF. Observe that pins 3 and 12 are not used, since the inverting configuration is not normally used with single ended input biasing techniques.

2.7 LM381A DUAL PREAMPLIFIER FOR ULTRA-LOW NOISE APPLICATIONS

2.7.1 Introduction

The LM381A is a dual preamplifier expressly designed to meet the requirements of amplifying low level signals in noise critical applications. Such applications include hydrophones, scientific and instrumentation recorders, low level wideband gain blocks, tape recorders, studio sound equipment, etc.

The LM381A can be externally biased for optimum noise performance in ultra-low noise applications. When this is done the LM381A provides a wideband, high gain amplifier with excellent noise performance.

The amplifier can be operated in either the differential or single ended input configuration. However, for optimum noise performance, the input must be operated single ended, since both transistors contribute noise in a differential stage, degrading input noise by the factor $\sqrt{2}$. (See Section 2.3) A second consideration is the design of the input bias circuitry. Both the load and biasing elements must be resistive, since active components would each contribute additional noise equal to that of the input device.

2.8 LM387/387A LOW NOISE MINIDIP DUAL PREAMPLIFIER

2.8.1 Introduction

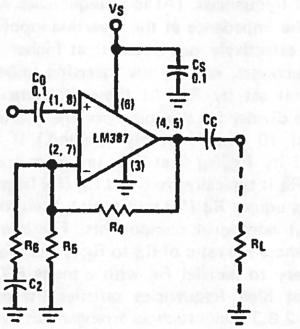
The LM387 is a low cost, dual preamplifier supplied in the popular 8 lead minidip package. The internal circuitry is identical to the LM381 and has comparable performance. By omitting the external compensation and single ended biasing pins it has been possible to package this dual amplifier into the 8 pin minidip, making for very little board space requirement. Like the LM381, this preamplifier is 100% noise tested and guaranteed, when purchased through authorized distributors. Total equivalent input noise is typically $0.65\mu V_{RMS}$ ($R_s = 600\Omega$, 100 Hz-10 kHz) and supply rejection ratio is typically 110 dB (f = 1 kHz). All other parameters are identical to the LM381. Biasing, compensation and split-supply operation are as previously explained.

2.8.2 Non-Inverting AC Amplifier

For low level signal applications requiring optimum noise performance the non-inverting configuration remains the most popular. The LM387 used as a non-inverting AC amplifier is configured similar to the LM381 and has the same design equations. Figure 2.8.1 shows the circuit with the equations duplicated for convenience.

2.8.3 Inverting AC Amplifier

For high level signals (greater than 300 mV), the inverting configuration may be used to overcome the positive input overload limit. Voltage gains of less than 20 dB are possible with the inverting configuration since the DC biasing resistor R_5 acts to voltage divide the incoming signal as previously described for the LM381. Design equations are the same as for the LM381 and are duplicated along with the inverting circuit in Figure 2.8.2.



$$R_4 = \left(\frac{V_s}{2.6} - 1\right) R_5$$

R_5 = 240kΩ MAXIMUM

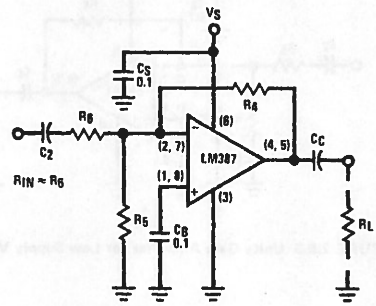
$$A_{VAC} = 1 + \frac{R_4}{R_6} \quad (R_5 >> R_6)$$

$$C_2 = \frac{1}{2\pi f_0 R_6}$$

$$C_c = \frac{1}{2\pi f R_L}$$

f_0 = LOW FREQUENCY −3dB CORNER $(f << f_0)$

FIGURE 2.8.1 LM387 Non-inverting AC Amplifier



$$R_4 = \left(\frac{V_s}{2.6} - 1\right) R_5$$

R_5 = 240kΩ MAXIMUM

$$A_{VAC} = -\frac{R_4}{R_6}$$

$$C_2 = \frac{1}{2\pi f_0 R_6}$$

$$C_c >> \frac{1}{2\pi f_0 R_L}$$

f_0 = LOW FREQUENCY −3dB CORNER

FIGURE 2.8.2 LM387 Inverting AC Amplifier

2.8.4 Unity Gain Inverting Amplifier

The requirement for unity gain stability is that the gain of the amplifier from pin 2 (or 7) to pin 4 (or 5) must be at least ten at all frequencies. This gain is the ratio of the feedback resistor R_4 divided by the total net impedance seen by the inverting input with respect to ground. The assumption is made that the driving, or source, impedance is small and may be neglected. In Figure 2.8.2 the net impedance looking back from the inverting input is $R_5 \| R_6$, at high frequencies. (At low frequencies where loop gain is large the impedance at the inverting input is very small and R_5 is effectively not present; at higher frequencies loop gain decreases, causing the inverting impedance to rise to the limit set by R_5. At these frequencies R_5 acts as a voltage divider for the input voltage guaranteeing amplifier gain of 10 when properly selected.) If the ratio of R_4 divided by $R_5 \| R_6$ is at least ten, then stability is assured. Since R_4 is typically ten times R_5 (for large supply voltages) and R_6 equals R_4 (for unity gain), then the circuit is stable without additional components. For low voltage applications where the ratio of R_4 to R_5 is less than ten, it becomes necessary to parallel R_5 with a series R-C network so the ratio at high frequencies satisfies the gain requirement. Figure 2.8.3 shows such an arrangement with the constraints on R_7 being given by Equations (2.8.1)-(2.8.3).

$$|A_V|(\text{pin 2 to 4}) = \frac{R_4}{R_5 \| R_6 \| R_7} \geqslant 10 \qquad (2.8.1)$$

$$R_Y = R_5 \| R_6 \qquad (2.8.2)$$

$$R_7 \leqslant \frac{R_Y R_4}{10 R_Y - R_4} \qquad (2.8.3)$$

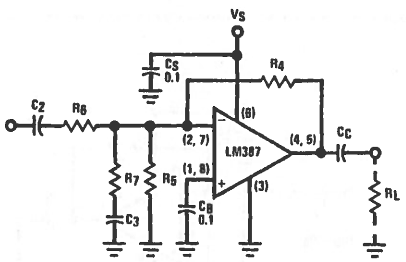


FIGURE 2.8.3 Unity Gain Amplifier for Low Supply Voltage

Example 2.8.1

Design a low noise unity gain inverting amplifier to operate from $V_S = 12V$, with low frequency capabilities to 20 Hz, input impedance equal to 20 kΩ, and a load impedance of 100 kΩ.

Solution:

1. $R_{in} = R_6 = 20k\Omega$.
2. For unity gain $R_4 = R_6$, $R_4 = 20k$.
3. From Figure 2.8.2:

$$R_4 = \left(\frac{V_s}{2.6} - 1\right) R_5 = \left(\frac{12}{2.6} - 1\right) R_5$$

$$R_4 = 3.62\ R_5$$

Therefore:

$$R_5 = \frac{R_4}{3.62} = \frac{20k}{3.62} = 5{,}525\Omega$$

Use $R_5 = 5.6k$.

4. From Equation (2.8.2):

$$R_Y = R_5 \| R_6 = \frac{5.6k \times 20k}{5.6k + 20k} = 4375$$

5. From Equation (2.8.3):

$$R_7 \leqslant \frac{R_Y R_4}{10 R_Y - R_4} = \frac{4375 \times 20 \times 10^3}{10 \times 4375 - (20 \times 10^3)} = 3684$$

Use $R_7 = 3.6k$

6. For $f_o = 20\text{Hz}$,

$$C_2 = \frac{1}{2\pi f_o R_6} = \frac{1}{2\pi \times 20 \times 20k} = 3.98 \times 10^{-7}$$

Use $C_2 = 0.5\mu F$.

For $f_{-3dB} = 20\text{Hz}$, the low frequency corner given by C_c and R_L must be at least a factor of 4 lower, i.e., $f < 5\text{Hz}$.

$$C_c = \frac{1}{2\pi f R_L} = \frac{1}{2\pi \times 5 \times 100k} = 3.18 \times 10^{-7}$$

Use $C_c = 0.33\mu F$

7. The selection of C_3 is somewhat arbitrary, as its effect is only necessary at high frequencies. A convenient frequency for calculation purposes is 20 kHz.

$$C_3 = \frac{1}{2\pi(20\text{kHz}) R_7} = \frac{1}{2\pi \times 20k \times 3.6k} = 2.21 \times 10^{-9}$$

Use $C_c = 0.0022\mu F$

2.8.5 Application to Feedback Tone Controls

One of the most common audio circuits requiring unity gain stability is active tone controls. Complete design details are given in Section 2.14. An example of modified Baxandall tone controls using an LM387 appears as Figure 2.14.17 and should be consulted as an application of the stabilizing methods discussed in Section 2.8.4.

2.9 LM382 LOW NOISE DUAL PREAMPLIFIER WITH RESISTOR MATRIX

2.9.1 Introduction

The LM382 is a dual preamplifier patterned after the LM381 low noise circuitry but with the addition of an internal resistor matrix. The resistor matrix allows the user to select a variety of closed loop gain options and frequency response characteristics such as flat-band, NAB (tape), or RIAA (phonograph) equalization. The LM382 possesses all of the features of the LM381 with two exceptions: no single ended input biasing option and no external pins for adding additional compensation capacitance. The internal resistors provide for biasing of the negative input automatically, so no external resistors are necessary and use of the LM382 creates the lowest parts count possible for standard designs. Originally developed for the automotive tape player market with a nominal supply voltage of +12V, the output is self queuing to about +6V (regardless of applied voltage — but this can be defeated, as will be discussed later). A diagram of the LM382 showing the resistor matrix appears as Figure 2.9.1.

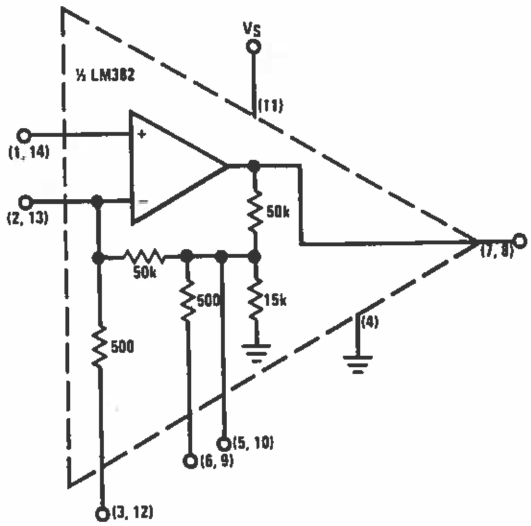
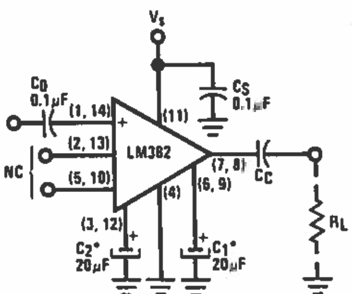


FIGURE 2.9.1 LM382 Resistor Matrix

2.9.2 Non-Inverting AC Amplifier

The fixed-gain flat-response configuration of the LM382 (Figure 2.9.2) shows that with just two or three capacitors a complete high gain, low noise preamplifier is created.



GAIN	*REQUIRED CAPACITORS
40dB	C_1 ONLY
55dB	C_2 ONLY
80dB	C_1 & C_2

FIGURE 2.9.2 LM382 as Fixed Gain-Flat Response Non-inverting Amplifier

To understand how the gains of Figure 2.9.2 are calculated it is necessary to redraw each case with the capacitors short-circuited and include only the relevant portion of the resistor network per Figure 2.9.1. The redrawn 40dB gain configuration (C_1 only) appears as Figure 2.9.3.

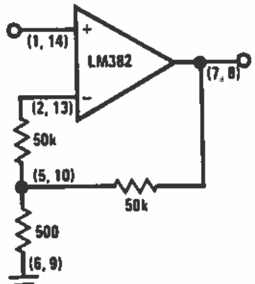


FIGURE 2.9.3 Equivalent Circuit for 40 dB Gain (C_1 Only)

Since bias currents are small and may be ignored in gain calculations, the 50k input resistor does not affect gain. Therefore, the gain is given by:

$$A_{V1} = 1 + \frac{50k}{500} = 101 \approx 40\text{dB}$$

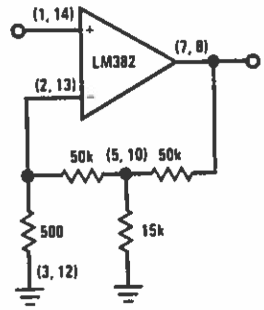


FIGURE 2.9.4 Equivalent Circuit for 55 dB Gain (C_2 Only)

With C_2 only, the redrawn equivalent circuit looks like Figure 2.9.4. Since the feedback network is wye-connected, it is easiest to perform a wye-delta transformation (see Appendix A3) in order to find an effective feedback resistor so the gain may be calculated. A complete transformation produces three equivalent resistors, two of which may be ignored. These are the ones that connect from the ends of each 50kΩ resistor to ground; one acts as a load on the amplifier and doesn't enter into the gain calculations, and the other parallels 500Ω and is large enough to have no effect. The remaining transformed resistor connects directly from the output to the input and is the equivalent feedback resistor, R_f. Its value is found from:

$$R_f \text{ (equivalent)} = 50k + 50k + \frac{(50k)^2}{15k} = 267k$$

The gain is now simply

$$A_{V2} = 1 + \frac{267k}{500} = 535 \approx 55\text{dB}$$

Adding both C_1 and C_2 gives the equivalent circuit of Figure 2.9.5.

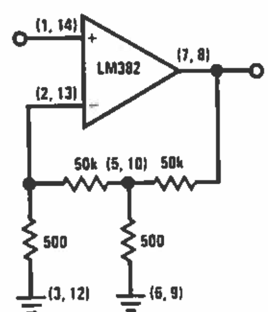


FIGURE 2.9.5 Equivalent Circuit for 80 dB Gain (C_1 and C_2)

Treating Figure 2.9.5 similarly to Figure 2.9.4, an equivalent feedback resistor is calculated:

$$R_f \text{ (equivalent)} = 50k + 50k + \frac{(50k)^2}{500} = 5.1 \text{ Meg}$$

Therefore, the gain is:

$$A_{V12} = 1 + \frac{5.1 \text{ Meg}}{500} = 10201 \approx 80\text{dB}.$$

2.9.3 Adjustable Gain for Non-Inverting Case

As can be learned from the preceding paragraphs, there are many combinations of ways to configure the resistor matrix. By adding a resistor in series with the capacitors it is possible to vary the gain. Care must be taken in attempting low gains (< 20dB), as the LM382 is not unity gain stable and should not be operated below gains of 20dB. (Under certain specialized applications unity gain is possible, as will be demonstrated later.) A general circuit allowing adjustable gain and requiring only one capacitor appears as Figure 2.9.6.

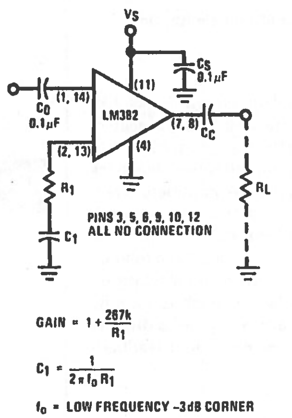


FIGURE 2.9.6 Adjustable Gain Non-Inverting Amplifier

Referring to Figure 2.9.1, it is seen that the R_1-C_1 combination is used instead of the internal 500Ω resistor and that the remaining pins are left unconnected. The equivalent resistance of the 50k-50k-15k wye feedback network was found previously to equal 267kΩ, so the gain is now given by Equation (2.9.1).

$$\text{Gain} = 1 + \frac{267k}{R_1} \qquad (2.9.1)$$

And C_1 is found from Equation (2.9.2):

$$C_1 = \frac{1}{2\pi f_o R_1} \qquad (2.9.2)$$

where: f_o = low frequency −3dB corner.

2.9.4 Internal Bias Override

As mentioned in the introduction, it is possible to override the internal bias resistor which causes the output quiescent point to sit at +6V regardless of applied voltage. This is done by adding a resistor at pin 5 (or 10) which parallels the internal 15kΩ resistor and defeats its effect (Figure 2.9.7).

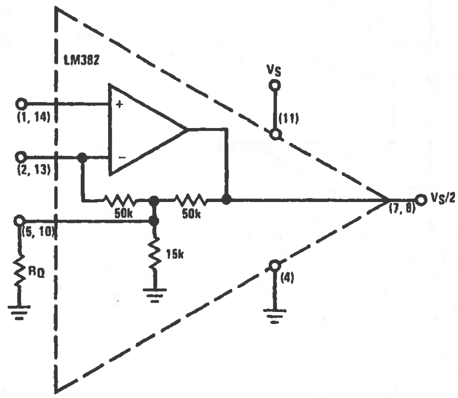


FIGURE 2.9.7 Internal Bias Override Resistor

Since the positive input is biased internally to a potential of +1.3V (see circuit description for LM381), it is necessary that the DC potential at the negative input equal +1.3V also. Because bias current is small (0.5μA), the voltage drop across the 50k resistor may be ignored, which says there is +1.3V across RQ. The current developed by this potential across RQ is drawn from the output stage, through the 50k resistor, through RQ and to ground. The subsequent voltage drop across the 50k resistor is additive to the +1.3V and determines the output DC level. Stated mathematically,

$$\frac{V_s}{2} = \left(\frac{50k}{RX}\right)1.3V + 1.3V \qquad (2.9.3)$$

where: $RX = RQ \| 15k$

From Equation (2.9.3) the relationships of RX and RQ may be expressed.

$$RX = \frac{50k}{\frac{V_s}{2.6} - 1} \qquad (2.9.4)$$

$$RQ = \frac{RX\,(15k)}{15k - RX} \qquad (2.9.5)$$

Example 2.9.1

Select RQ such that the output of a LM382 will center at $12V_{DC}$ when operated from a supply of $V_s = 24V_{DC}$.

Solution

1. Calculate RX from Equation (2.9.4).

$$RX = \frac{50 \times 10^3}{\frac{24}{2.6} - 1} = 6075\Omega$$

2. Calculate RQ from Equation (2.9.5).

$$RQ = \frac{(6075)(15 \times 10^3)}{(15 \times 10^3) - 6075} = 10210\Omega$$

Use RQ = 10kΩ.

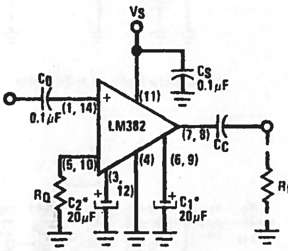
Since RQ parallels the 15k resistor, then the AC gains due to the addition of capacitor C_1 or C_2 (or both) (as given in Figure 2.9.2) are changed. The new gain equations become a function of RQ and are given as Equations (2.9.6)-(2.9.8) and refer to Figure 2.9.8.

$$C_1 \text{ Only:} \quad \text{Gain} \approx 1 + \frac{50k}{RQ \| 500} \qquad (2.9.6)$$

$$C_2 \text{ Only:} \quad \text{Gain} = 201 + \frac{5 \times 10^6}{RX} \qquad (2.9.7)$$

$$C_1 \,\&\, C_2: \quad \text{Gain} \approx 201 + \frac{5 \times 10^6}{RQ \| 500} \qquad (2.9.8)$$

where: RX and RQ are given by Equations (2.9.4) and (2.9.5).



* – IF REQUIRED
PINS 2 & 13 NO CONNECTION

FIGURE 2.9.8 Fixed Gain Amplifier with Internal Bias Override

Continuing the previous example to find the effect of RQ on the gain yields:

3. C_1 Only: $\quad \text{Gain} = 1 + \dfrac{50k}{10k\|500} = 53.6\text{dB}$

4. C_2 Only: $\quad \text{Gain} = 201 + \dfrac{5 \times 10^6}{6075} = 60.2\text{dB}$

5. C_1 & C_2: $\quad \text{Gain} = 201 + \dfrac{5 \times 10^6}{10k\|500} = 80.6\text{dB}$

2.9.5 Inverting AC Amplifier

Examination of the resistor matrix (Figure 2.9.1) reveals that an inverting AC amplifier can be created with just one resistor (Figure 2.9.9).

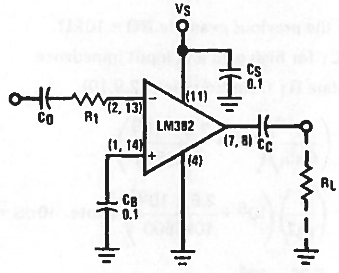
The gain is found by calculating the equivalent feedback resistance as before, and appears in Figure 2.9.9. Higher gains are possible (while retaining large input resistance = R_1) by adding capacitor C_1 as shown in Figure 2.9.10.

The internal bias override technique discussed for the non-inverting configuration may be applied to the inverting case as well. The required value of RQ is calculated from Equations (2.9.4) and (2.9.5) and affects the gain relation shown in Figures 2.9.9 and 2.9.10. The new gain equations are:

$$\text{Without } C_1: \quad \text{Gain} = \left(-\frac{1}{R_1}\right)\left(10^5 + \frac{2.5 \times 10^9}{RQ \| 15k}\right) \qquad (2.9.9)$$

$$\text{With } C_1: \quad \text{Gain} = \left(-\frac{1}{R_1}\right)\left(10^5 + \frac{2.5 \times 10^9}{RQ \| 500}\right) \qquad (2.9.10)$$

and the circuit is shown in Figure 2.9.11.



$\text{GAIN} = -\dfrac{267k}{R_1}$ (> 20dB FOR STABILITY)

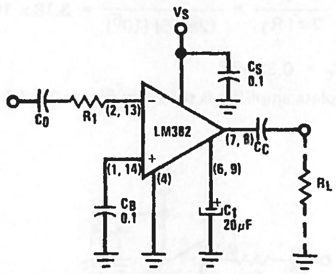
$C_o = \dfrac{1}{2\pi f_o R_1}$

f_o = LOW FREQUENCY -3dB CORNER

INPUT IMPEDANCE = R_1

PINS 3, 5, 6, 9, 10, 12 NOT USED

FIGURE 2.9.9 LM382 as Inverting AC Amplifier



$\text{GAIN} = -\dfrac{5.1 \times 10^6}{R_1}$

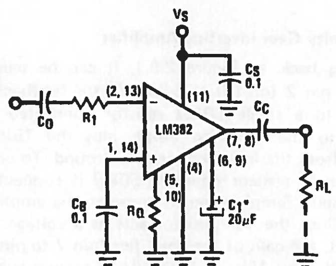
$C_o = \dfrac{1}{2\pi f_o R_1}$

f_o = LOW FREQUENCY -3dB CORNER ($C_c R_L >> C_o R_1$)

INPUT IMPEDANCE = R_1

PINS 3, 5, 10, 12 NOT USED

FIGURE 2.9.10 High Gain Inverting AC Amplifier



* – IF REQUIRED
PINS 3 & 12 NOT USED

FIGURE 2.9.11 Inverting Amplifier with Internal Bias Override

Example 2.9.2

Design an inverting amplifier to operate from a supply of $V_s = 24V_{DC}$, with output quiescent point equal to $12V_{DC}$, gain equal to 40dB, input impedance greater than 10kΩ, low frequency performance flat to 20Hz, and a load impedance equal to 100kΩ.

1. From the previous example RQ = 10kΩ.
2. Add C_1 for high gain and input impedance.
3. Calculate R_1 from Equation (2.9.10).

$$R_1 = \left(\frac{1}{\text{Gain}}\right)\left(10^5 + \frac{2.5 \times 10^9}{RQ \| 500}\right)$$

$$R_1 = \left(\frac{1}{10^2}\right)\left(10^5 + \frac{2.5 \times 10^9}{10k \| 500}\right) \quad \text{(Note: 40dB} = 10^2 \text{V/V)}$$

$$R_1 = 5.35 \times 10^4$$

Use R_1 = 56kΩ.

4. Calculate C_o from equation shown in Figure 2.9.9.

$$C_o = \frac{1}{2\pi f_o R_1} = \frac{1}{(2\pi)(20)(56k)} = 1.42 \times 10^{-7}$$

Use C_o = 0.15μF.

5. Calculate C_c from Equation (2.6.12).

$$C_c = \frac{1}{2\pi f R_1} = \frac{1}{(2\pi)(5)(10^5)} = 3.18 \times 10^{-7}$$

Use C_c = 0.33μF

The complete amplifier is shown in Figure 2.9.12.

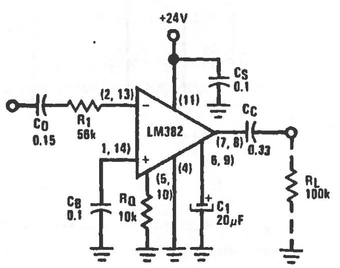
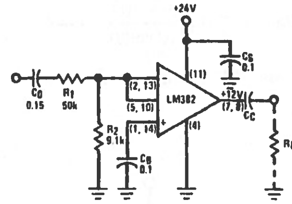


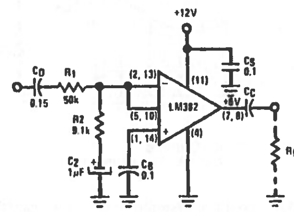
FIGURE 2.9.12 Inverting Amplifier with Gain = 40 dB and V_S = +24 V

2.9.6 Unity Gain Inverting Amplifier

Referring back to Figure 2.9.1, it can be seen that by shorting pin 2 (or 13) to 5 (or 10) the feedback network reduces to a single 50kΩ resistor connected from the output to the inverting input, plus the 15kΩ biasing resistor from the inverting input to ground. To create unity gain then, a resistor equal to 50kΩ is connected to the minus input. Simple enough; however, the amplifier is not stable. Since the 15k resistor acts as a voltage divider to the input, the gain of the amplifier (pin 7 to pin 2) is only 50k divided by 15k, or 3.33V/V. Minimum required gain for stability is 10V/V, so it becomes necessary to shunt the 15k resistor with a new resistor such that the parallel combination equals 5kΩ. This may be done AC or DC, depending upon supply voltage. If done DC (tied from pin 2 (or 13) directly to ground), then it becomes RQ (from Figure 2.9.7) and affects the output DC level. Placing a capacitor in series with this resistor makes it effective only for AC voltages and does not change the output level. The required resistor equals 9.1kΩ, which is close enough to the required RQ for V_s = 24V. Two examples of unity gain amplifiers appear as Figure 2.9.13 and should satisfy the majority of applications.



(a) Supply Voltage = 24 Volts



(b) Supply Voltage = 12 Volts

FIGURE 2.9.13 Unity Gain Inverting Amplifier

2.9.7 Remarks

The above application hints are not meant to be all-inclusive, but rather are offered as an aid to LM382 users to familiarize them with its many possibilities. Once understood, the internal resistor matrix allows for many possible configurations, only a few of which have been described in this section.

2.10 LM1303 STEREO PREAMPLIFIER

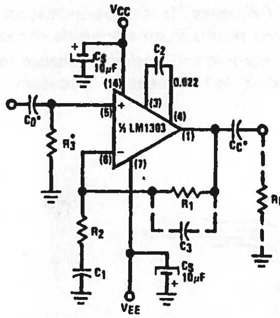
2.10.1 Introduction

The LM1303 is a dual preamplifier designed to be operated from split supplies ranging from ±4.5V up to ±15V. It has "op amp" type inputs allowing large input signals with low distortion performance. The wideband noise performance is superior to traditional operational amplifiers, being typically $0.9\mu V_{RMS}$ (10kHz bandwidth). Compensation is done externally and offers the user a variety of choices, since three compensation points are brought out for each amplifier. The LM1303 is pin-for-pin compatible with "739" type dual preamplifiers and in most applications serves as a direct replacement.

2.10.2 Non-Inverting AC Amplifier

The LM1303 used as a non-inverting amplifier (Figure 2.10.1) with split supplies allows for economical direct-coupled designs if the DC levels between stages are maintained at zero volts. Gain and C_1 equations are shown in the figure. Resistor R_3 is made equal to R_1 and provides DC bias currents to the positive input. Compensation capacitor C_2 is equal to $0.022\mu F$ and guarantees unity gain stability with a slew rate of approximately $1V/\mu s$. Higher slew rates are possible when higher gains are used by reducing C_2 proportionally to the increase in gain, e.g., with a gain of ten, C_2 can equal $0.0022\mu F$, increasing the slew rate to around $10V/\mu s$. Some layouts may dictate the addition of C_3 for added stability. It should be picked according to equation (2.10.1) where f_H is the high frequency $-3dB$ corner.

$$C_3 = \frac{1}{2\pi f_H R_1} \qquad (2.10.1)$$



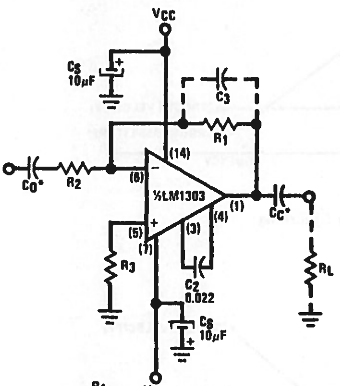
$$A_{VAC} = 1 + \frac{R_1}{R_2}$$

$$C_1 = \frac{1}{2\pi f_o R_2} \qquad C_c R_L >> C_1 R_2 \quad C_o R_3 >> C_1 R_2$$

f_o = LOW FREQUENCY $-3dB$ CORNER

* – MAY BE OMITTED FOR DIRECT-COUPLED DESIGNS.

FIGURE 2.10.1 LM1303 Non-inverting AC Amplifier



$$A_{VAC} = -\frac{R_1}{R_2}$$

$$C_o = \frac{1}{2\pi f_o R_2} \quad (C_c R_L >> C_o R_2)$$

f_o = LOW FREQUENCY $-3dB$ CORNER

* – MAY BE OMITTED FOR DIRECT-COUPLED DESIGNS.

FIGURE 2.10.2 LM1303 Inverting AC Amplifier

2.10.3 Inverting AC Amplifier

For applications requiring inverting operation, Figure 2.10.2 should be used. Capacitors C_2 and C_3 have the same considerations as the non-inverting case. Resistor R_3 is made equal to R_1 again, minimizing offsets and providing bias current. The same slew rate-gain stability trade-offs are possible as before.

2.11 PHONO PREAMPLIFIERS AND RIAA EQUALIZATION

2.11.1 Introduction

Phono preamplifiers differ from other preamplifiers only in their frequency response, which is tailored in a special manner to compensate, or equalize, for the recorded characteristic. If a fixed amplitude input signal is used to record a phonograph disc, while the frequency of the signal is varied from 20 Hz to 20 kHz, the playback response curve of Figure 2.11.1 will result. Figure 2.11.1 shows a plot of phono cartridge output amplitude versus frequency, indicating a severe alteration to the applied fixed amplitude signal. *Playback equalization* corrects for this alteration and recreates the applied flat amplitude frequency response. To understand why Figure 2.11.1 appears as it does, an explanation of the recording process is necessary.

2.11.2 Recording Process and RIAA

The grooves in a stereo phonograph disc are cut by a chisel shaped cutting stylus driven by two vibrating systems arranged at right angles to each other (Figure 2.11.2). The cutting stylus vibrates mechanically from side to side in accordance with the signal impressed on the cutter. This is termed a "lateral cut" as opposed to the older method of "vertical cut." The resultant movement of the groove back and forth about its center is known as groove modulation. The amplitude of this modulation cannot exceed a fixed amount or "cutover" occurs. (Cutover, or overmodulation, describes the breaking through the wall of one groove into the wall of the previous groove.) The ratio of the maximum groove signal amplitude possible before cutover, to the effective groove noise amplitude caused by the surface of the disc material, determines the dynamic range of a record (typically 58 dB). The latter requirement results from the grainy characteristic of the disc surface acting as a noise generator. (The cutting stylus is heated in recording to impart a smooth side wall to minimize the noise.) Of interest in phono preamp design is that the record noise performance tends to be ten times worse than that of the preamp, with typical wideband levels equal to $10\mu V$.

Amplitude and frequency characterize an audio signal. Both must be recorded and recovered accurately for high quality music reproduction. Audio amplitude information translates to groove modulation amplitude, while the frequency of the audio signal appears as the rate of change of the groove modulations. Sounds simple enough, but Figure 2.11.1 should, therefore, be a horizontal straight line centered on 0dB, since it represents a fixed amplitude input signal. The trouble results from the characteristics of the cutting head. Without the negative feedback coils (Figure 2.11.2) the velocity frequency response has a resonant peak at 700 Hz due to its construction. Adding the feedback coils produces a velocity output independent of frequency; therefore, the cutting head is known as a constant velocity device (Figure 2.11.2a).

Figure 2.11.1 appears as it does because the cutting amplifier is pre-equalized to provide the recording character-

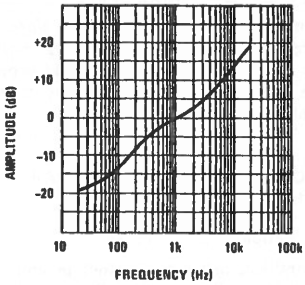


FIGURE 2.11.1 Typical Phono Playback Characteristic for a Fixed Amplitude Recorded Signal

istic shown. Two reasons account for the shape: first, low frequency attenuaticn prevents cutover; second, high frequency boosting improves signal-to-noise ratio. The unanswered question is why is all this necessary?

The not-so-simple answer begins with the driving coils of the cutting head. Being primarily inductive, their impedance characteristic is frequency dependent. If a fixed amplitude input signal translates to a fixed voltage used to drive the coils (called "constant velocity") then the resulting current, i.e., magnetic field, hence amplitude of vibration, becomes frequency dependent (Figure 2.11.2a); if a fixed amplitude input signal translates to a fixed current, i.e., fixed amplitude of vibration, used to drive the coils (called "constant amplitude) then the resulting voltage, i.e., cutting velocity, becomes frequency dependent (Figure 2.11.2b). With respect to frequency, for a given input amplitude the cutting head has only one degree of freedom: vibrating *rate* (constant velocity = voltage drive) or vibrating *distance* (constant amplitude = current drive).

The terms constant velocity and constant amplitude create confusion until it is understood that they have meaning only for a *fixed amplitude input signal*, and are used strictly to describe the resultant behavior of the cutting head as a *function of frequency*. It is to be understood that changing the *input* level results in an *amplitude* change for constant amplitude recording and a *velocity* change for constant velocity recording *independent* of frequency. For example,

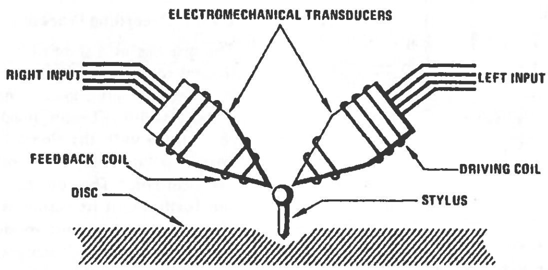


FIGURE 2.11.2 Stereo Cutting Head

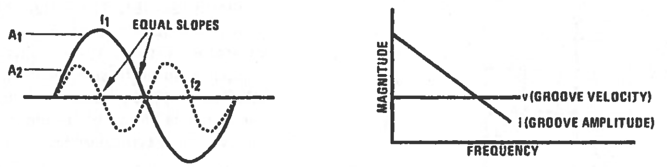


FIGURE 2.11.2A Constant Velocity Recording

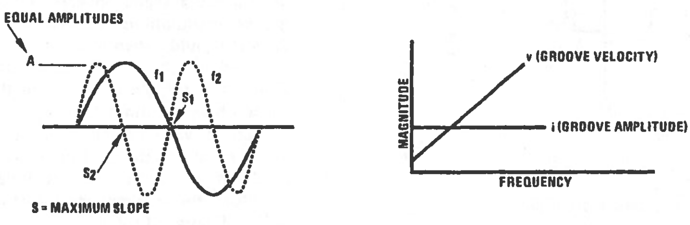


FIGURE 2.11.2B Constant Amplitude Recording

if an input level of 10mV results in 0.1 mil amplitude change for constant amplitude recording and a velocity of 5cm/s for constant velocity recording, then a change of input level to 20mV would result in 0.2 mil and 10cm/sec respectively – independent of frequency.

Each of these techniques when used to drive the vibrating mechanism suffers from dynamic range problems. Figures 2.11.2a and 2.11.2b diagram each case for two frequencies an octave apart. The discussion that follows assumes a fixed amplitude input signal and considers only the effect of frequency change on the cutting mechanism.

Constant velocity recording (Figure 2.11.2a) displays two readily observable characteristics. The amplitude varies inversely with frequency and the maximum slope is constant with frequency. The second characteristic is ideal since magnetic pickups (the most common type) are constant velocity devices. They consist of an active generator such as a magnetic element moving in a coil (or vice versa) with the output being proportional to the speed of movement through the magnetic field, i.e., proportional to groove velocity. However, the variable amplitude creates serious problems at both frequency extremes. For the ten octaves existing between 20Hz and 20kHz, the variation in amplitude is 1024 to 1! If 1kHz is taken as a reference point to establish nominal cutter amplitude modulation, then at low frequencies the amplitudes are so great that cutover occurs. At high frequencies the amplitude becomes so small that acceptable signal-to-noise ratios are not possible – indeed, if any displacement exists at all. So much for constant velocity.

Looking at Figure 2.11.2b, two new observations are seen with regard to constant amplitude. Amplitude is constant with frequency (which corrects most of the ills of constant velocity), but the maximum slope varies directly with frequency, i.e., groove velocity is directly proportional to frequency. So now velocity varies 1024 to 1 over the audio band – swell! Recall that magnetic cartridges are constant velocity devices, not constant amplitude, so the output will rise at the rate of +6dB/octave. (6dB increase equals twice the amplitude.) To equalize such a system would require 60dB of headroom in the preamp – not too practical. The solution is to try to get the best of both systems, which results in a modified constant amplitude curve where the midband region is allowed to operate constant velocity.

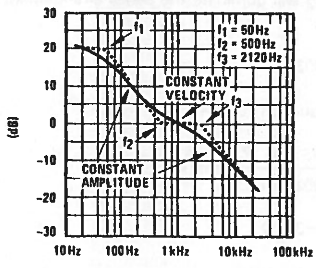


FIGURE 2.11.3 RIAA Playback Equalization

The required RIAA (Record Industry Association of America) playback equalization curve (Figure 2.11.3) shows the idealized case dotted and the actual realization drawn solid. Three frequencies are noted as standard design reference points and are sometimes referred to as time constants. This is a carryover from the practice of specifying corner frequencies by the equivalent RC circuit (t = RC) that realized the response. Conversion is done simply with the expression $t = 1/2\pi f$ and results in time constants of 3180μs for f_1, 318μs for f_2, and 75μs for f_3. Frequency f_2 is referred to as the *turnover* frequency since this is the point where the system changes from constant amplitude to constant velocity. (Likewise, f_3 is another turnover frequency.) Table 2.11.1 is included as a convenience in checking phono preamp RIAA response.

TABLE 2.11.1 RIAA Standard Response

Hz	dB	Hz	dB
20	+19.3	800	+0.7
30	+18.6	1k	0.0*
40	+17.8	1.5k	−1.4
50	+17.0	2k	−2.6
60	+16.1	3k	−4.8
80	+14.5	4k	−6.6
100	+13.1	5k	−8.2
150	+10.3	6k	−9.6
200	+8.2	8k	−11.9
300	+5.5	10k	−13.7
400	+3.8	15k	−17.2
500	+2.6	20k	−19.6

* Reference frequency.

2.11.3 Ceramic and Crystal Cartridges

Before getting into the details of designing RIAA feedback networks for magnetic phono cartridges, a few words about crystal and ceramic cartridges are appropriate. In contra-distinction to the constant velocity magnetic pickups, ceramic pickups are constant amplitude devices and therefore do not require equalization, since their output is inherently flat. Referring to Figure 2.11.3 indicates that the last sentence is not entirely true. Since the region between f_2 and f_3 is constant velocity, the output of a ceramic device will drop 12dB between 500Hz and 2000Hz. While this appears to be a serious problem, in reality it is not. This is true due to the inherently poor frequency response of ceramic and restriction of its use to lo-fi and mid-fi market places. Since the output levels are so large (100mV-2V), a preamp is not necessary for ceramic pickups; the output is fed directly to the power amplifier via passive tone (if used) and volume controls.

2.11.4 LM387 or LM381 Phono Preamp

Magnetic cartridges have very low output levels and require low noise devices to amplify their signals without appreciably degrading the system noise performance. Nevertheless, note that usually the noise of the cartridge and loading resistor is comparable to the active device and should be included in the calculations (see Appendix A5).

Typical cartridge output levels are given in Table 2.11.2.

Output voltage is specified for a given modulation velocity. The magnetic pickup is a velocity device, therefore output is proportional to velocity. For example, a cartridge producing 5mV at 5cm/s will produce 1mV at 1cm/s and is specified as having a sensitivity of 1mV/cm/s.

In order to transform cartridge sensitivity into useful preamp design information, we need to know typical and maximum modulation velocity limits of stereo records.

TABLE 2.11.2

Manufacturer	Model	Output at 5cm/sec
Empire Scientific	999	5mV
	888	8mV
Shure	V-15	3.5mV
	M91	5mV
Pickering	V-15 AT3	5mV

The RIAA recording characteristic establishes a maximum recording velocity of 25cm/s in the range of 800 to 2500Hz. Typically, good quality records are recorded at a velocity of 3 to 5cm/s.

Figure 2.11.3 shows the RIAA playback equalization. To obtain this, the desired transfer function of the preamplifier is given by:

$$\frac{V_{OUT}}{V_{IN}} = \frac{A(s+2\pi\cdot 500)}{(s+2\pi\cdot 50)(s+2\pi\cdot 2120)} \quad (2.11.1)$$

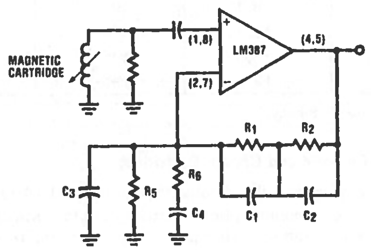


FIGURE 2.11.4 RIAA Phono Preamp

From Figure 2.11.4:

$$\frac{V_{OUT}}{V_{IN}} = \frac{K\left[s+\dfrac{R_1+R_2}{(C_1+C_2)R_1R_2}\right]}{R_6\left(s+\dfrac{1}{C_1R_1}\right)\left(s+\dfrac{1}{C_2R_2}\right)}+1 \quad (2.11.2)$$

Equating coefficients of (2.11.1) and (2.11.2),

$$R_1C_1 = \frac{1}{2\pi\cdot 50} = 3180\mu s \quad (2.11.3)$$

$$R_2C_2 = \frac{1}{2\pi\cdot 2120} = 75\mu s \quad (2.11.4)$$

$$\frac{R_1R_2(C_1+C_2)}{R_1+R_2} = \frac{1}{2\pi\cdot 500} = 318\mu s \quad (2.11.5)$$

Substituting (2.11.3) and (2.11.4) in (2.11.5):

$$R_1 = 11.78R_2 \quad (2.11.6)$$

$$0\text{dB reference gain} = \frac{z+R_6}{R_6} \quad (2.11.7)$$

where: $z=\left(R_1 \parallel \frac{1}{2\pi fC_1}\right)+\left(R_2 \parallel \frac{1}{2\pi fC_2}\right)$

Resistor R_5 together with R_1 and R_2 sets the DC bias (Section 2.6) and C_3 stabilizes the amplifier by rolling off the feedback at higher frequencies since the LM387 is not compensated for unity gain.

Example 2.11.1

Design a phonograph preamp operating from a 24V supply, with a cartridge of 0.5mV/cm/s sensitivity, to drive a power amplifier with an input overload limit of $1.25V_{RMS}$.

Solution

1. The maximum cartridge output of 25cm/s is (0.5mV/cm/s) × (25cm/sec) = 12.5mV. The required mid-band gain is:

$$\frac{1.25V_{RMS}}{12.5mV_{RMS}} = 100$$

2. Before selecting R_6 to give a gain of 40dB at 1kHz, we must determine the complex impedance of the R_1R_2, C_1C_2 network at 1kHz. Ideally this should be such that R_6 is relatively low to minimize any noise contributions from the feedback network.

3. If we assume the amplifier output must be able to drive the feedback equalization network to the rated output at 20kHz, the slew rate required is:

$$\text{S.R.} = 2\pi E p f, \text{ where } E_p = 1.25 \times \sqrt{2}$$
$$= 2\pi \times 1.77 \times 20 \times 10^3$$
$$= 0.22V/\mu s$$

Using 1V/μs as a safety margin and noting that the output sink current of the LM387 is 2mA, the capacitance of the feedback network should be:

$$\leq \frac{2\times 10^{-3}}{1\times 10^{-6}}$$

$$\leq 0.002\mu F$$

Since C_2 will dominate the series arrangement of C_1 and C_2, put:

$C_2 = 0.0027\mu F$

4. From Equation (2.11.4):

$$R_2 = \frac{75\times 10^{-6}}{0.0027\times 10^{-6}} = 28k\Omega$$

Put $R_2 = 30k\Omega$

5. Equation (2.11.6):

$R_1 = 11.78R_2$

$= 11.78 \times 30 \times 10^3 = 353k\Omega$

Put $R_1 = 360k\Omega$

6. Equation (2.11.3):

$$C_1 = \frac{3180 \times 10^{-6}}{360 \times 10^3} = 0.0088\,\mu F$$

Put $C_1 = 0.01\,\mu F$

7. At 1 kHz the feedback network impedance (z) = 37.6k 49°. Equation (2.11.7):

$$0\,dB \text{ reference gain} = 100 = \frac{37.6 \times 10^3}{R_6} + 1$$

$$\therefore R_6 = \frac{37.6 \times 10^3}{99} = 379\,\Omega$$

Put $R_6 = 390\,\Omega$

8. From Equation (2.6.4):

$$\left(\frac{V_{CC}}{2.6} - 1\right) R_5 = R_1 + R_2$$

$$\therefore R_5 = \frac{390 \times 10^3}{8.23} = 47\,k\Omega$$

Note: This value of R_5 will center the output at the mid supply point. However, for symmetrical clipping it is worth noting that the LM387 can swing to within 0.3V of ground and 1.7V of V_{CC}. To put the output midway between these points ($11.2V_{DC}$ with $V_{CC} = 24V$), put $R_5 = 56\,k\Omega$.

9. From Equation (2.6.10):

$$C_4 = \frac{1}{2\pi f_o R_6}$$

$$= \frac{1}{2\pi \cdot 10 \cdot 390}$$

$$= 40.8 \times 10^{-6}$$

Put $C_4 = 47\,\mu F$

The completed design is shown in Figure 2.11.5 where a 47 kΩ input resistor has been included to provide the RIAA standard cartridge load.

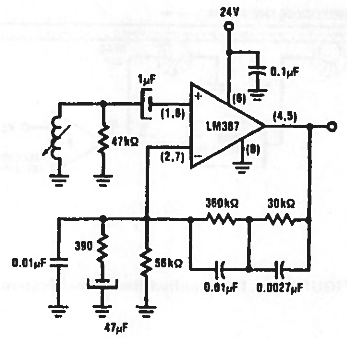


FIGURE 2.11.5 LM387 Phono Preamp (RIAA)

The LM381 integrated circuit may be substituted for the LM387 in Figure 2.11.5 by making the appropriate pin number changes.

2.11.5 LM382 Phono Preamp

By making use of the internal resistor matrix, a minimum parts count low noise phono preamp is possible using the LM382 (Figure 2.11.6). The circuit has been optimized for a supply voltage equal to 12-14 V. The midband 0 dB reference gain equals 46 dB (200 V/V) and cannot easily be altered. For designs requiring either gain or supply voltage changes, the required extra parts make selection of a LM381 or LM387 more appropriate.

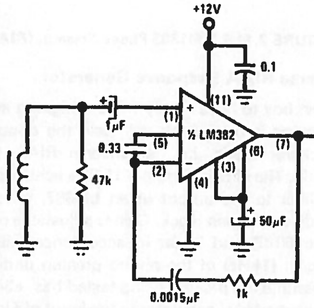


FIGURE 2.11.6 LM382 Phono Preamp. (RIAA)

2.11.6 LM1303 Phono Preamp

The LM1303 allows a convenient low noise phono preamp design when operating from split supplies. The circuit appears as Figure 2.11.7. For trimming purposes and/or gain changes the relevant formulas follow:

$$0\,dB \text{ Ref Gain} = 1 + \frac{R_2}{R_3} \tag{2.11.5}$$

$$f_1 = \frac{1}{2\pi R_1 C_1} \tag{2.11.6}$$

$$f_2 \approx \frac{1}{2\pi R_2 C_1} \tag{2.11.7}$$

$$f_3 = \frac{1}{2\pi R_2 C_2} \tag{2.11.8}$$

As shown in Figure 2.11.7, the 0 dB reference gain (1 kHz) equals about 34 dB and the feedback values have been altered slightly to minimize pole-zero interactions.

2

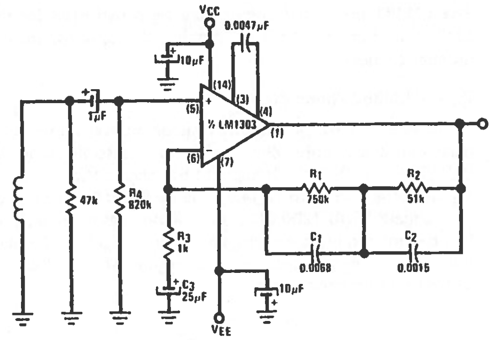


FIGURE 2.11.7 LM1303 Phono Preamp. (RIAA)

2.11.7 Inverse RIAA Response Generator

A useful test box to have handy while designing and building phono preamps is one which will yield the opposite of the playback characteristic, i.e., an inverse RIAA (or record) characteristic. The circuit (Figure 2.11.9) is achieved by adding a passive filter to the output of an LM387, used as a flat-response adjustable gain block. Gain is adjustable over a range of 24dB to 60dB and is set in accordance with the 0dB reference gain (1kHz) of the phono preamp under test. For example, assume the preamp being tested has +34dB gain at 1kHz. Connect a 1kHz generator to the input of Figure 2.11.9. The passive filter has a loss of −40dB at 1kHz, which is corrected by the LM387 gain, so if a 1kHz test output level of 1V is desired from a generator input level of 10mV, then the gain of the LM387 is set at +46dB (+46dB − 40dB + 34dB = 40dB (×100); 10mV × 100 = 1V). Break frequencies of the filter are determined by Equations (2.11.9)-(2.11.11).

$$f_1 = 50\text{Hz} = \frac{1}{2\pi R_9 C_4} \quad (2.11.9)$$

$$f_2 = 500\text{Hz} \approx \frac{1}{2\pi R_{10} C_4} \quad (2.11.10)$$

$$f_3 = 2120\text{Hz} = \frac{1}{2\pi R_{10} C_5} \quad (2.11.11)$$

The R_7-C_3 network is necessary to reduce the amount of feedback for AC and is effective for all frequencies beyond 20Hz. With the values shown the inverse RIAA curve falls within 0.75dB of Table 2.11.1.

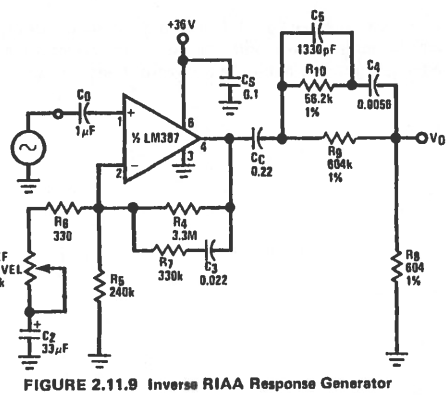


FIGURE 2.11.9 Inverse RIAA Response Generator

2.12 TAPE PREAMPLIFIERS

2.12.1 Introduction

A simplified diagram of a tape recording system is shown in Figure 2.12.1. The tape itself consists of a plastic backing coated with a ferromagnetic material. Both the record and erase heads are essentially inductors with circular metal cores having a narrow gap at the point of contact with the tape. The tape coating then forms a low reluctance path to complete the magnetic circuit. As the tape moves across the record head gap, the magnetic field at the trailing edge of the gap leaves the tape coating permanently magnetized with a remanent flux level (Φ_R) proportional to the signal current in the record head windings.

The bias and erase currents (I_B and I_E) are constant amplitude and frequency waveforms (between 50kHz and 200kHz) generated by the bias oscillator. In the erase head, the amplitude of the waveform (from 30 Volts to 150 Volts typically) will determine the degree to which previously recorded signals are "erased" from the tape — in a good machine this will be from 60dB to 75dB below the normal recording level. This same waveform, reduced in amplitude to between 5 to 25 times the maximum recording signal level, is used in the record head to determine the "operating point" of the magnetic recording process. Distortion, maximum output level and sensitivity are strong functions of the bias level. To gain some insight into the need for bias, let us take a closer look at the recording process.

Figure 2.12.2(a) shows the permanent magnetization Br (or remanent flux) of a short section of magnetic tape, obtained by applying a magnetizing field H produced by a dc current in the record head winding. This curve is clearly non-linear and if an ac signal current was used in the head winding a highly distorted recording would be made. One solution would be to apply a steady dc bias to the record head along with the ac signal so that the tape was always magnetized in a linear region of the curve (between points A and B for example). This method, called *dc bias*, uses only one part of the curve and reduces the distortion but has a very poor S/N ratio. An improvement may be obtained by pre-magnetizing the tape to saturation and using a dc bias on the record head to bring the magnetization back to zero. Even so, S/N ratios above 30dB are not easy to achieve.

For high S/N ratios and low distortion, another method called *ac bias* is used.

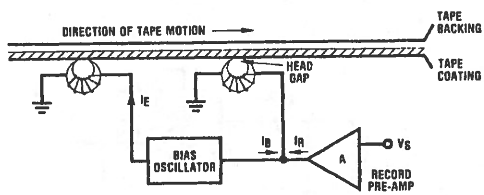
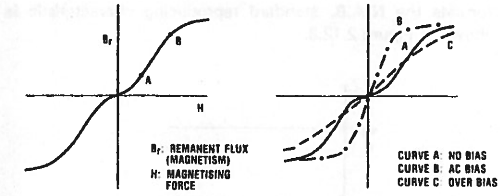


FIGURE 2.12.1 Simplified Recording System



(a)Br-H Curve for Recording Tape (b)Br-H Curve with AC Bias

FIGURE 2.12.2

Figure 2.12.2(b) shows the remanent flux characteristic when a high level ac magnetic field is applied along with the signal. The sensitivity of the tape (curve B) has increased and the magnetization is a linear function of the signal over a much wider range. Note however, that if the bias signal is increased even more (curve C), the tape sensitivity falls off and the non-linearity increases again. The choice of "best" bias current level will depend on a number of factors including the characteristics of the tape and the record/playback heads. Also the ac bias waveform must be free of even order harmonics as these would add an effective dc component to the bias causing distortion for large signal swings and degrading the S/N ratio.

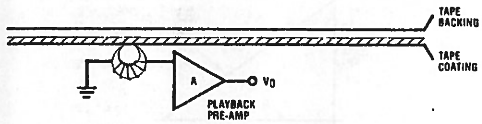


FIGURE 2.12.3 Simplified Playback System

Figure 2.12.3 shows a simplified diagram of a tape playback system. In a two head system (the majority of cassette recorders) the playback head also functions as the record head with appropriate switching. A three head system (record/-playback/erase) allows monitoring of the actual recorded signal and the playback head gap can be optimized to improve its frequency response.

Magnetic tape is recorded "constant current" — i.e., constant recording current with frequency, implying a constant recorded magnetic flux level for a given signal amplitude at all frequencies. Since the heads can be regarded as primarily inductive, the impedance of a playback head rises at a 6dB/octave rate with respect to increasing frequency. Therefore the signal voltage from the playback head to the playback preamplifier does not have a flat frequency response, but instead shows a steadily increasing level with increasing signal frequency (Figure 2.12.4).

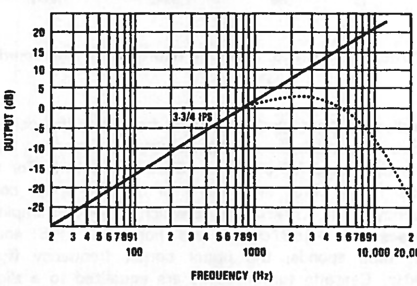


FIGURE 2.12.4 Playback Head Voltage Output vs. Frequency

For practical heads at high frequencies there is an abrupt change in response resulting in a severe decrease in amplitude with a continuing increase in frequency (dashed line on Figure 2.12.4). There are several reasons for this phenomenon — all different and unrelated, but each contributing to the loss of high frequency response. The first area of degradation is due to the effects of the decreasing recorded wavelengths of the higher frequencies.

$$\text{wavelength} = \lambda = \frac{\text{Tape speed (IPS)}}{\text{Frequency (Hz)}} \tag{2.12.1}$$

Two factors are important in minimizing recorded wavelength problems: recording tape speed (Figure 2.12.5) and *playback head* gap width (Figure 2.12.6).

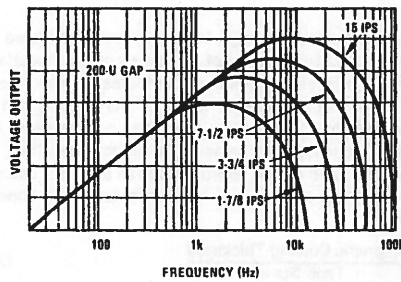


FIGURE 2.12.5 Effect of Tape Speed on Response

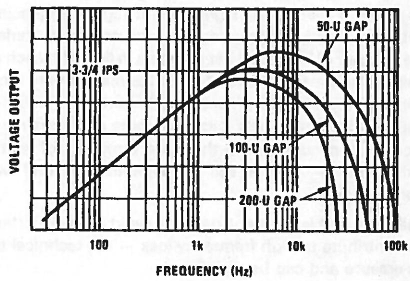


FIGURE 2.12.6 Effect of Head Gap on Response

The first of these is accounted for by the fact that for a given number of flux lines per unit cross-sectional area of the tape (corresponding to a given magnetizing force), higher tape speeds increase the total available flux in the head. For the playback head, when the gap length equals the recorded wavelength (100-U = 100 micro-inches = 0.0001 inches), no output signal is possible since both edges of the gap are at equal magnetic potentials. The gap loss for any given playback head gap and recorded wavelength can be calculated from Equation (2.12.2).

$$\text{Gap Loss (dB)} = 20 \log_{10} \frac{\sin \pi R}{\pi R} \tag{2.12.2}$$

$$\text{where: } R = \frac{\text{Gap Width}}{\text{Wavelength}}$$

Table 2.12.1 gives the calculated gap losses for typical gap widths at 1-7/8 I.P.S. and 3-3/4 I.P.S. tape speeds.

		Gap Loss with Signal Freqency (dB)				
Tape Speed (IPS)	Gap Width Micro-inches	1kHz	2kHz	4kHz	8kHz	16kHz
1-7/8	50 – U	−0.01	−0.04	−0.16	−0.66	−2.78
	100 – U	−0.04	−0.16	−0.66	−2.78	−15.61
3-3/4	100 – U	−0.01	−0.09	−0.16	−0.66	−2.78
	160 – U	−0.03	−0.10	−0.42	−1.73	−8.14

TABLE 2.12.1 Playback Head Gap Loss.

Other areas of serious high frequency loss are related to the thickness and formulation of the tape coating material. The thickness of the tape coating contributes to high frequency loss since only the surface layers of the coating contribute measurably to the recording of shorter wavelengths. As the signal frequency increases this effect becomes more pronounced and can be approximated as a −6dB/octave roll-off with a corner frequency equivalent to a time constant T given by:

$$T = \frac{\text{Magnetic Coating Thickness}}{\text{Tape Speed}} \qquad (2.12.3)$$

The particular coating formulation used affects the high frequency response because as the magnetic flux variations increase in intensity, a point is reached at which the tape saturates and higher flux levels cannot produce a corresponding higher permanent magnetization of the tape. This effect is particularly significant at higher freqencies and can be explained by regarding the tape coating material as a large number of individual bar magnets in line with each other. At higher frequencies more of these bar magnets are recorded per inch of tape: thus each one grows shorter. As the effective length of the bar magnets decrease, more and more magnetic cancellation occurs due to the close proximity of north and south poles — hence self demagnetization and weaker recorded signals.

Finally the ac bias current used to avoid tape distortion will also contribute to high frequency loss — the technical term is bias erasure and can be significant.

2.12.2 Frequency Equalization

If the tape/head system were "ideal", the application of an unequalized signal current I_R to the recording head would result in a recorded flux Φ_R on the tape exactly proportional to the input signal current and independent of frequency — Figure 2.12.7(a). Similarly the recorded flux on the tape would produce a playback current I_p also independent of frequency. The voltage on the playback head terminals will be proportional to the rate of change of flux (Figure 2.12.7(b). To compensate for this 6dB/octave rise in amplitude with increasing frequency, the playback preamplifier is equalized for the response shown in Figure 2.12.7(c).

Because of the miscellaneous losses in a real tape recording system further frequency equalization is necessary in both the record and playback preamplifiers. Additionally, despite variations in tape formulations and thicknesses, there are internationally recognized frequency equalization standards (similar to the R.I.A.A. equalization standard for phonograph discs — see Section 2.11.2). For open-reel and cartridge tape formats the N.A.B. standard reproducing characteristic is shown in Figure 2.12.8.

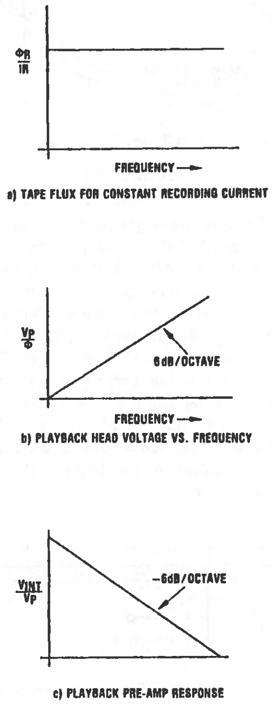


FIGURE 2.12.7 "Ideal" Record/Play System

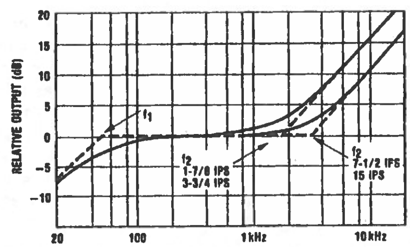


FIGURE 2.12.8 N.A.B. Standard Reproducing Characteristic

Regardless of tape speed the lower frequency (f_1) corner is 50Hz, below which the amplifier output (for constant flux in the head) should fall off at a −6dB/octave rate. For tape speeds 1-7/8 I.P.S. and 3-3/4 I.P.S., the upper corner frequency (f_2) is 1.77kHz, above which the output amplitude increases at a +6dB/octave rate. For 7-1/2 I.P.S. and 15 I.P.S. tape speeds, the upper corner frequency (f_2) is 3.18kHz. Cassette format tapes are equalized to a slightly different standard discussed in a later section.

Ignoring tape/head losses for the moment, if we take the N.A.B. standard response and add the integrating function necessary to compensate the $d\Phi/dt$ characteristic of the playback head, we arrive at the overall playback preamplifier frequency response shown in Figure 2.12.9 (the 0dB reference amplitude is defined at the upper corner frequency f_2).

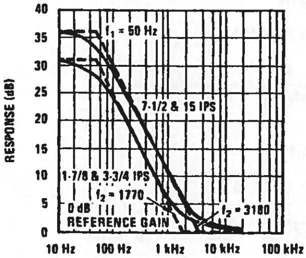


FIGURE 2.12.9 N.A.B. Playback Equalization Including Integration

In the same way the N.A.B. record characteristic will be the complement of the N.A.B. playback characteristic of Figure 2.12.8. This is shown in Figure 2.12.10 where the record current is boosted at +6dB/octave below 50Hz and cut at 6dB/octave above either 1.77kHz or 3.18kHz, depending on the tape speed.

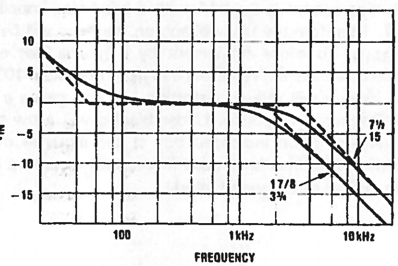


FIGURE 2.12.10 N.A.B. Record Equalization

Both the record and the playback preamplifier responses can be modified to accommodate the losses previously described, but in order to ensure compatibility of tapes recorded to N.A.B. standards on one machine with playback on a different machine, the necessary equalization for losses is obtained in a specific way. In the record preamplifier compensation is made for record head losses and the high frequency losses caused by a particular tape formulation and coating thickness, as well as providing the inverse of the N.A.B. playback characteristic and a current drive to the record head. The playback preamplifier is used to provide the standard playback characteristic, to compensate for playback head gap losses and to integrate the playback head voltage.

2.12.3 LM381 OR LM387 Tape Record Preamp

The frequency response of the record preamp is the complement of the N.A.B. playback equalization including record head and tape loss compensation. A practical design method is to equalize the playback preamplifier first for a flat response with a standard equalized reference tape of the type to be used. Then the record preamplifier is equalized using the same tape formulation and appropriate bias level adjustments for an overall flat response between record and playback. However, to illustrate where loss compensation occurs we will first design a record preamplifier for a 3-3/4 I.P.S. tape speed.

Figure 2.12.11 shows a typical response curve for an eight track record/play head at 3-3/4 I.P.S., obtained with an input signal level −12dB below tape saturation, peak biased at 1kHz and with no record and playback equalization. For a head gap width of 100−U, Table 2.12.1 shows that the playback head gap loss at 16Hz is still under −3dB. Using Equation (2.12.2) for a 440 microinch thick tape coating at 3-3/4 I.P.S.;

$$T = \frac{440 \times 10^{-6}}{3.75} = 117\mu\text{Seconds}$$

or f−3dB = 1.36kHz.

From this, we would expect the tape thickness loss to predominate, causing a −6dB/octave fall in frequency response above 1.4kHz. This is confirmed by Figure 2.12.11 where the overall response rises at 6dB/octave to 1.36kHz (f_2)

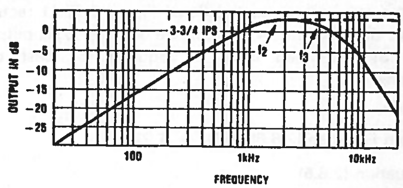
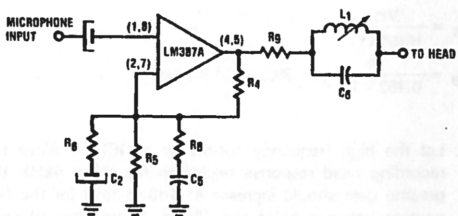


FIGURE 2.12.11 Record/Playback Head Frequency Response

and is then flattened out until head and tape high frequency losses cause the output to fall off (f_3). Note that the −3dB corner frequency of 1.36kHz corresponds almost exactly to the corner frequency (f_2) required by the N.A.B. standard for 3-3/4 I.P.S. tape speed, which calls for a −6dB/octave response (Figure 2.12.10). As a result, out preamplifier will not require any equalization for the N.A.B. standard but will require compensation above 4kHz for the head and other tape losses. This can be obtained with a 6dB/octave boost from 4kHz up to the upper desired frequency limit for the system. The response required can be met with the circuit shown in Figure 2.12.12.



Resistors R_4 and R_5 set the dc bias and resistor R_6 and capacitor C_2 set the mid band gain as before (see Section 2.6). Capacitor C_5 sets the +3dB corner frequency f_3 at which the preamplifier compensates for the head losses.

$$f_3 = \frac{1}{2\pi C_5 R_6} \tag{2.12.4}$$

The preamp gain increases at +6dB/octave above f_3 until the desired high freqency cut-off is reached (f_4)

$$R_8 = \frac{1}{2\pi f_4 C_5} \tag{2.12.5}$$

Resistor R_9 is chosen to provide the proper head recording current

$$R_9 = \frac{V_O}{I_{R(MAX)}} \tag{2.12.6}$$

L_1 and C_6 form a parallel resonant trap at the bias frequency to present a high impedance to the record bias waveform and prevent intermodulation distortion.

Example 2.12.1

A recorder having a 24V power supply uses recording heads with the response characteristic of Figure 2.12.11 requiring 38µA ac drive current. A microphone of 10mV peak output is used. Single ended input is required for best noise performance.

Solution.

1. From Equation (2.6.5) let R_5 = 1.2kΩ.
2. Equation (2.6.6)

$$R_4 = \left(\frac{V_{CC}}{1.3} - 1\right) R_5$$

$$R_4 = \left(\frac{24}{1.3} - 1\right) 1200$$

$$R_4 = 2.09 \times 10^4 \cong 22\text{k}\Omega$$

3. The maximum output of the LM381 is (V_{CC} − 2V) p-p. With a 24V power supply, the maximum output is 22V (p-p) or 7.8V_{RMS}. Therefore an output swing of 6V_{RMS} is reasonable. For the specified head with a proper recording bias level a record current I_R of 38µA gives a recorded level −12dB below saturation. The preamp output should be able to deliver about four times this current without distortion. Therefore $I_{R(MAX)}$ = 0.152mA.

From Equation (2.12.6)

$$R_9 = \frac{V_O}{I_{R(MAX)}}$$

$$R_9 = \frac{6}{0.152 \times 10^{-3}} = 39.5 \times 10^3 \cong 39\text{k}\Omega$$

4. Let the high frequency cut-off be at 16kHz. Since the recording head response begins to fall off at 4kHz, the preamp gain should increase at 6dB/octave for the two octaves between 4kHz and 16kHz. If we allow 6V_{RMS} output voltage swing, then the peak gain $= \frac{6}{10 \times 10^{-3}}$ = 600 or 55.6dB

The midband gain is 12dB below this or 43.6dB (151V/V)

5. From Equation (2.6.9) the midband gain is

$$\frac{R_4 + R_6}{R_6} = 151$$

$$R_6 = \frac{R_4}{150} = \frac{22 \times 10^3}{150} = 146.7$$

$$R_6 = 150\Omega$$

6. Equation (2.6.10)

$$C_2 = \frac{1}{2\pi f_O R_6} = \frac{1}{6.28 \times 30 \times 150}$$

$$C_2 \cong 33\mu F$$

7. Equation (2.12.4)

$$C_5 = \frac{1}{2\pi f_3 R_6} = \frac{1}{6.28 \times 4 \times 10^3 \times 150}$$

$$C_5 \cong 0.27\mu F$$

8. Equation (2.12.5)

$$R_8 = \frac{1}{2\pi f_4 C_5} = \frac{1}{6.28 \times 16 \times 10^3 \times 2.7 \times 10^{-7}}$$

$$R_8 = 36.8 \cong 39\Omega$$

The completed circuit is shown in Figure 2.12.13 with the addition of the bias trap L_1C_6. R_{10} and C_9 couple the bias waveform from the bias oscillator to the head with R_{10} being used to adjust the actual bias level. For the specified head, peak bias current is 0.48mA with a head bias impedance of 27kΩ. Therefore the bias voltage on the head will be around 13V_{RMS}. To allow adjustment by R_{10}, the bias oscillator should be able to deliver about 38V_{RMS} or around 100V(p-p). Over biasing will reduce distortion but also cause a drop in high frequency response. Under biasing will allow the high frequencies to be increased but at the expense of higher distortion (the bias level used in cassette recorders is much more critical as discussed later).

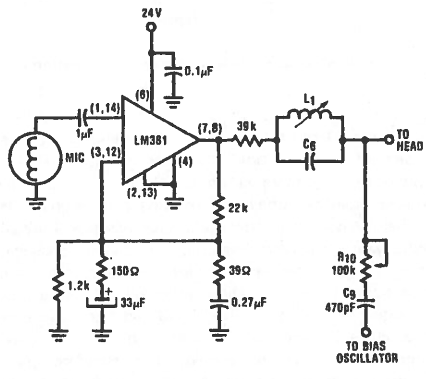


FIGURE 2.12.13 Typical Tape Recording Amplifier

2.12.4 LM387 OR LM381 Tape Playback Preamps

The N.A.B. playback response of Figure 2.12.8 can be obtained with the circuit of Figure 2.12.14. Resistors R_4 and R_5 set the dc bias according to Section 2.6. The reference gain of the preamp, at the upper corner frequency f_2, is set by the ratio

$$\text{0dB reference gain} = \frac{R_7+R_6}{R_6} \qquad (2.12.7)$$

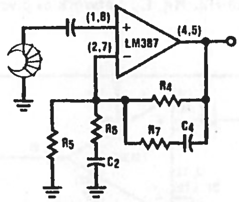


FIGURE 2.12.14 N.A.B. Tape Preamp

The corner frequency f_2 is determined when the impedance of C_4 equals the value of resistor R_7

$$f_2 = \frac{1}{2\pi C_4 R_7} \qquad (2.12.8)$$

Corner frequency f_1 is determined when $X_{C4} = R_4$:

$$\text{i.e. } f_1 = \frac{1}{2\pi C_4 R_4} \qquad (2.12.9)$$

The low frequency −3dB roll off point, f_0, is set where $X_{C2} = R_6$

$$f_0 = \frac{1}{2\pi C_2 R_6} \qquad (2.12.10)$$

Example 2.12.2

Using the R/P head specified in Example 2.12.1, design a N.A.B. equalized preamp using the LM387. At 1kHz, 3-3/4 I.P.S. the head sensitivity is 800μV and the required preamp output is $0.5V_{RMS}$.

Solution

1. From Equation (2.6.3) let $R_5 = 240k\Omega$.
2. Equation (2.6.4):

$$R_4 = \left(\frac{V_{CC}}{2.6} - 1\right) R_5$$

$$R_4 = \left(\frac{24}{2.6} - 1\right) 2.4 \times 10^5$$

$$R_4 = 1.98 \times 10^6 \approx 2.2M\Omega$$

3. For a corner frequency, f_1, equal to 50Hz, Equation (2.12.9) is used.

$$C_4 = \frac{1}{2\pi f_1 R_4} = \frac{1}{6.28 \times 50 \times 2.2 \times 10^6}$$

$$= 1.45 \times 10^{-9} \approx 1500pF$$

4. From Figure 2.12.8, the corner frequency $f_2 = 1770Hz$ at 3-3/4 IPS. Resistor R_7 is found from Equation (2.12.8).

$$R_7 = \frac{1}{2\pi f_2 C_4}$$

$$R_7 = \frac{1}{6.28 \times 1770 \times 1.5 \times 10^{-9}} = 59.9 \times 10^3$$

$$R_7 \approx 62k\Omega$$

5. The required voltage gain at 1kHz is:

$$A_V = \frac{0.5V_{RMS}}{800\mu V_{RMS}} = 6.25 \times 10^2 V/V = 56dB$$

6. From Figure 2.12.9 we see the reference frequency gain, above f_2, is 5dB down from the 1kHz value or 51dB (355V/V).

From Equation (2.12.7):

$$\text{0dB Ref Gain} = \frac{R_7 + R_6}{R_6} = 355$$

$$R_6 = \frac{R_7}{355-1} = \frac{62k}{354} = 175$$

$$R_6 \approx 180\Omega$$

7. For low frequency corner $f_o = 40Hz$, Equation (2.12.10):

$$C_2 = \frac{1}{2\pi f_o R_6} = \frac{1}{6.28 \times 40 \times 180} = 2.21 \times 10^{-5}$$

$$C_2 \approx 20\mu F$$

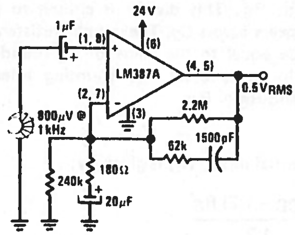
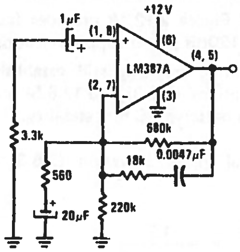


FIGURE 2.12.15 Typical Tape Playback Amplifier

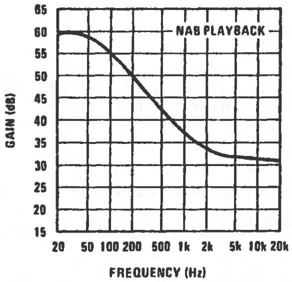
An example of a LM387A tape playback preamp designed for 12 volt operation is shown in Figure 2.12.16 along with its frequency response.



(a) NAB Tape Circuit

FIGURE 2.12.16 (a) LM387 Tape Preamp

2



(b) Frequency Response of NAB Circuit
FIGURE 2.12.16 (b) LM387 Tape Preamp

2.12.5 Fast Turn-On NAB Tape Playback Preamp

The circuit shown in Figure 2.12.15 requires approximately 2.5 seconds to turn on for the gain and supply voltage chosen in the example. Turn-on time can closely be approximated by:

$$t_{ON} \approx -R_4 C_2 \ln\left(1 - \frac{1.2}{V_{CC}}\right) \qquad (2.12.11)$$

As seen by Equation (2.12.11), increasing the supply voltage decreases turn-on time. Decreasing the amplifier gain also decreases turn-on time by reducing the R_4C_2 product.

Where the turn-on time of the circuit of Figure 2.12.14 is too long, the time may be shortened by using the circuit of Figure 2.12.17. The addition of resistor R_D forms a voltage divider with R_6'. This divider is chosen so that zero DC voltage appears across C_2. The parallel resistance of R_6' and R_D is made equal to the value of R_6 found by Equation (2.12.7). In most cases the shunting effect of R_D is negligible and $R_6' \approx R_6$.

For differential input, R_D is given by:

$$R_D = \frac{(V_{CC} - 1.2)\, R_6'}{1.2} \qquad (2.12.12)$$

For single ended input:

$$R_D = \frac{(V_{CC} - 0.6)\, R_6'}{0.6} \qquad (2.12.13)$$

In cases where power supply ripple is excessive, the circuit of Figure 2.12.17 cannot be used since the ripple is coupled into the input of the preamplifier through the divider.

The circuit of Figure 2.12.18 provides fast turn-on while preserving the 120dB power supply rejection.

The DC operating point is still established by R_4/R_5. However, Equations (2.6.3) and (2.6.5) are modified by a factor of 10 to preserve DC bias stability.

For differential input, Equation (2.6.3) is modified as:

$$R_5 = \frac{2V_{BE}}{100\, I_{Q2}} = \frac{1.2}{50 \times 10^{-6}} \qquad (2.6.3a)$$

$= 24\,k\Omega$ maximum

For single ended input:

$$R_5 = \frac{V_{BE}}{50\, I_{FB}} = \frac{0.6}{50 \times 10^{-4}} \qquad (2.6.5a)$$

$= 120\,\Omega$ maximum

Equations (2.12.7), (2.12.8), and (2.12.10) describe the high frequency gain and corner frequencies f_2 and f_o as before.

Frequency f_1 now occurs where X_{C4} equals the composite impedance of the R_4, R_6, C_2 network as given by Equation (2.12.14).

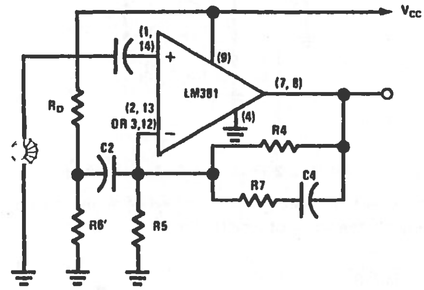


FIGURE 2.12.17 Fast Turn-On NAB Tape Preamp

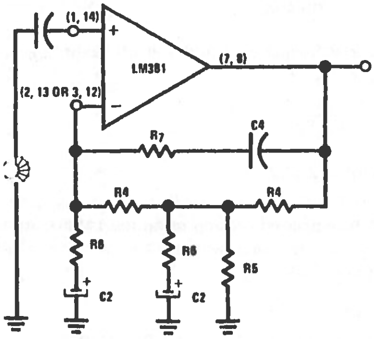


FIGURE 2.12.18 Two-Pole Fast Turn-On NAB Tape Preamp

$$C_4 = \frac{1}{2\pi f_1 R_6 \left[\left(\frac{R_4 + R_6}{R_6}\right)^2 - 1\right]} \qquad (2.12.14)$$

The turn-on time becomes:

$$t_{ON} \approx -2\sqrt{R_4 C_2}\, \ln\left(1 - \frac{1.2}{V_{CC}}\right) \qquad (2.12.15)$$

Example 2.12.3

Design an NAB equalized preamp with the fast turn-on circuit of Figure 2.18.18 for the same requirements as given in Example 2.12.2.

Solution

1. From Equation (2.6.3a) let $R_5 = 24\,k\Omega$.

2. Equation (2.6.4):

$$R_4 = \left(\frac{V_{CC}}{2.6} - 1\right) R_5$$

$$= \left(\frac{24}{2.6} - 1\right) 24 \times 10^3 = 1.98 \times 10^5$$

$$R_4 \approx 220\text{k}\Omega$$

3. From Example 2.12.2, the reference frequency gain, above f_2, is 51 dB or 355 V/V.

Equation (2.12.7):

$$\frac{R_7 + R_6}{R_6} = 355$$

4. The corner frequency f_2 is 1770 Hz for 3-3/4 IPS.

Equation (2.12.8):

$$C_4 = \frac{1}{2\pi f_2 R_7}$$

5. The corner frequency f_1 is 50 Hz and is given by Equation (2.12.14).

$$C_4 = \frac{1}{2\pi f_1 R_6 \left[\left(\frac{R_4 + R_6}{R_6}\right)^2 - 1\right]}$$

6. Solving Equations (2.12.7), (2.12.8), and (2.12.14) simultaneously gives:

$$R_6 = \frac{R_4 \left[f_1 + \sqrt{f_1{}^2 + f_1 f_2 (\text{Ref Gain})}\right]}{f_2 (\text{Ref Gain})} \quad (2.12.15)$$

$$R_6 = \frac{2.2 \times 10^5 \left(50 + \sqrt{2500 + 50 \times 1770 \times 355}\right)}{1770 \times 355}$$

$$= 1.98 \times 10^3 \approx 2\text{k}\Omega$$

7. From Equation (2.12.7):

$$R_7 = 354 R_6 = 708 \times 10^3$$

$$R_7 \approx 680\text{k}\Omega$$

8. Equation (2.12.8):

$$C_4 = \frac{1}{2\pi f_2 R_7} = \frac{1}{6.28 \times 1770 \times 680 \times 10^3}$$

$$C_4 = 1.32 \times 10^{-10} \approx 120\text{pF}$$

9. Equation (2.12.10):

$$C_2 = \frac{1}{2\pi f_o R_6} = \frac{1}{6.28 \times 40 \times 2 \times 10^3}$$

$$C_2 = 1.99 \times 10^{-6} \approx 2\mu\text{F}$$

This circuit is shown in Figure 2.12.19 and requires only 0.1 seconds to turn on. Note, however, that the non-inverting input has to charge the head coupling capacitance to 1.2V through an internal 250kΩ resistor and will increase the turn-on time slightly.

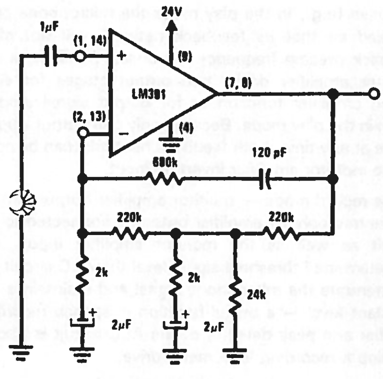


FIGURE 2.12.19

2.12.6 LM382 Tape Playback Preamp

With just one capacitor in addition to the gain setting capacitors, it is possible to design a complete low noise, NAB equalized tape playback preamp (Figure 2.12.20). The circuit is optimized for automotive use, i.e., V_s = 10-15V. The wideband 0dB reference gain is equal to 46dB (200V/V) and is not easily altered. For designs requiring either gain or supply voltage changes the required extra parts make selection of a LM387 a more appropriate choice.

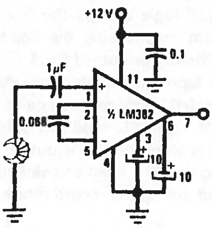


FIGURE 2.12.20 LM382 Tape Preamp (NAB, 1-7/8 & 3-3/4 IPS)

2.12.7 LM1303 Tape Playback Preamp

For split supply applications, the LM1303 may be used as a tape preamp as shown in Figure 2.12.21. Design equations are given below for trimming or alteration purposes. (Frequency points refer to Figure 2.12.9.)

$$\text{0dB Ref Gain} = 1 + \frac{R_2}{R_3} \quad (2.12.16)$$

$$f_1 = \frac{1}{2\pi R_1 C_1} \quad (2.16.17)$$

$$f_2 = \frac{1}{2\pi R_2 C_1} \quad (2.12.18)$$

response (e.g., in the play mode the microphone preamp is disabled so that its feedback network will not affect the playback preamp frequency response). Similarly, a common monitor amplifier drives two output stages for either the record amplifier function or for output signal amplification when in the play mode. Because only one output stage can be active at any time, both feedback networks can be connected to the monitor amplifier inverting input.

In the record mode — monitor amplifier output pin 10 active — the microphone amplifier output is connected to the ALC circuit as well as the monitor amplifier input. Above a predetermined threshold signal level the ALC circuit operates to attenuate the microphone signal and maintain a relatively constant level — a useful function in speech recorders. The rectifier and peak detector of the ALC circuit is also used to develop a recording level meter drive.

2.12.10 LM1818 Microphone and Playback Preamplifiers

Both the microphone and playback amplifiers are similar in design (Figure 2.12.28). The non-inverting inputs, pins 16 and 17, are biased at $1.2V_{DC}$ through 50kΩ resistors. Normally these resistors would also source current (24μA) at turn-on for the input capacitor from the tape head. This capacitor is usually selected to be large in value to give a low impedance at low frequencies which will minimize the amplifier input noise current degrading the system noise figure. To prevent long turn-on times, an internal circuit will source 200μA to pin 17 at turn-on, enabling capacitors around 10μF to be used. At the same time, the R/P logic clamps the head to prevent this charge current from magnetizing the head at turn-on. The amplifiers have collector currents of 50μA — optimized for low noise with typical tape head source impedances — and are internally compensated for closed loop gains greater than 5. In the "Record" mode, Q_8 is saturated and Q_4 is "off" so that the microphone signal is amplified to the output stage Q_9. Pin 2 is held low, clamping the R/P head and sinking the bias current flowing in the head during the record mode.

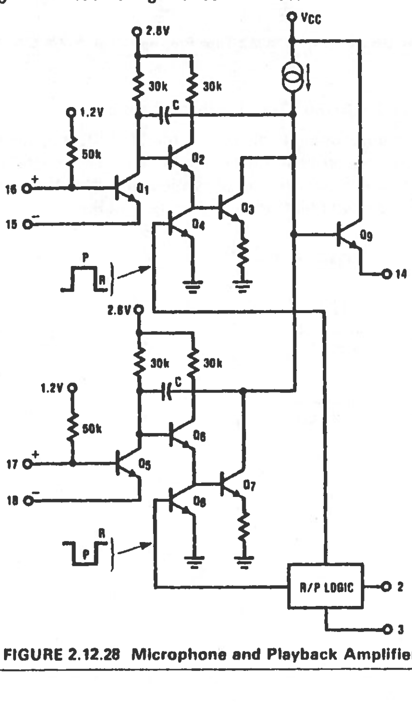


FIGURE 2.12.28 Microphone and Playback Amplifiers

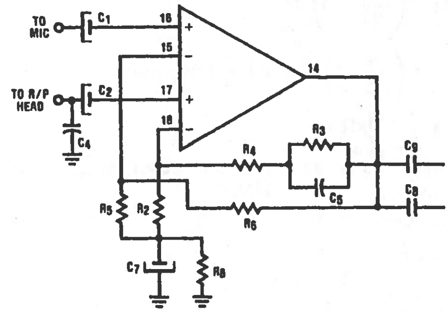


FIGURE 2.12.29 Microphone and Playback Preamplifier Feedback Networks

In the circuit of Figure 2.12.29, the microphone and playback amplifiers show the same low frequency roll-off capacitor C_7. Both inverting inputs are referenced at $1.2V - 0.7V = 0.5V_{DC}$. The output quiescent point, pin 14, is established by negative feedback through the external divider $(R_6 + R_5)/R_8$. For bias stability the current through R_8 is made ten times the current from the inverting inputs (pin 15 or pin 18). This current is the input stage collector current of 50μA

$$\therefore R_8 = \frac{0.5}{10 \times 50 \times 10^{-6}} = 1\text{k}\Omega \text{ maximum} \qquad (2.12.19)$$

For low values of R_2 and R_5, and large values of R_3 and R_4

$$(R_5 + R_6) = (2V_{DC} - 1)R_8 \qquad (2.12.20)$$

where V_{DC} = pin 14 voltage.

For the playback preamplifier, the midband gain above corner frequency f_2 (Figure 2.12.24).

$$A_{VAC} = 1 + \frac{R_4}{R_2} \qquad (2.12.21)$$

The low frequency corner f_1 (Figure 2.12.24) is determined where the impedance of C_5 equals R_3

$$f_1 = \frac{1}{2\pi C_5 R_3} \qquad (2.12.22)$$

The upper corner frequency f_2 is determined by

$$f_2 = \frac{1}{2\pi R_4 C_5} \qquad (2.12.23)$$

For a tape recorder to accomodate either ferric or CrO_2 tapes, Equation (2.12.23) has two solutions, depending on whether f_2 is 1.33kHz or 2.27kHz. Since C_5 also sets the lower corner frequency of f_1 which is common to either type of tape, R_4 should be switched. This will decrease the midband gain above f_2 [Equation (2.12.21)] for CrO_2 tapes by about 4.7dB.

In the "Record" mode the microphone amplifier is on and the dc output level at pin 14 is given by

$$V_{DC} = \frac{R_5 + R_6 + R_8}{2(R_5 + R_8)} \qquad (2.12.24)$$

The AC voltage gain is given by

$$A_{VAC}(MIC) = 1 + \frac{R_6}{R_5} \qquad (2.12.25)$$

The output amplifiers (Figure 2.12.30) are used to provide a low impedance drive to an audio power amplifier stage when in the "Play" mode, or to provide the necessary equalization for the tape head in the "Record" mode. Again the R/P logic (shown as switches in Figure 2.12.30) decides which stage will be active and therefore which feedback network is operational. The input from the microphone or playback preamplifiers is coupled to the non-inverting input of the differential pair Q_{10}, Q_{11}. The base of Q_{10} is biased externally via R_{10} from the supply voltage divider R_{11} and R_{12}. This divider is normally designed to place the amplifier quiescent outputs at half supply voltage to maximize the output signal swing capability. Both output stages Q_{16} and Q_{18} are Class A amplifiers with active current source loads Q_{15} and Q_{17}. The position of the R/P logic switches determines which current source is active and delivering 700µA.

The gain for the audio output stage (pin 9 output) is set by the ratio of R_{14} and R_{16}

$$\text{Playback gain } V_{AC} = 1 + \frac{R_{16}}{R_{14}} \quad (2.12.26)$$

The low frequency 3dB corner f_0 is given by

$$f_0 = \frac{1}{2\pi C_{12} R_{14}} \quad (2.12.27)$$

When the R/P logic is in the record mode, pin 10 output is active and the midband gain

$$A_{V(REC)} = 1 + \frac{R_{15}}{R_{14}} \quad (2.12.28)$$

The resistor R_{22} provides the proper head recording current

$$R_9 = \frac{V_{O(MAX)}}{I_{R(MAX)}} \quad (2.12.29)$$

For inexpensive, monaural cassette recorders, C_{17} is used to shunt R_{22} to compensate for high frequency losses (Figure 2.12.22)

$$f_3 = \frac{1}{2\pi C_{17} R_{22}} \quad (2.12.30)$$

C_{26} is used to filter the bias waveform at the output from the record amplifier (in place of the more expensive bias trap of Figure 2.12.13).

During the record mode the output from the microphone preamplifier is used to supply a signal to the ALC circuit and meter drive circuit. In certain recording situations — speech for example where the speaker can vary in distance from the microphone — it is convenient to have a circuit (ALC), which continuously and automatically adjusts the overall gain of the recording (microphone) amplifier in order to maintain a proper recording signal level that is neither buried in noise nor high enough to cause tape saturation.

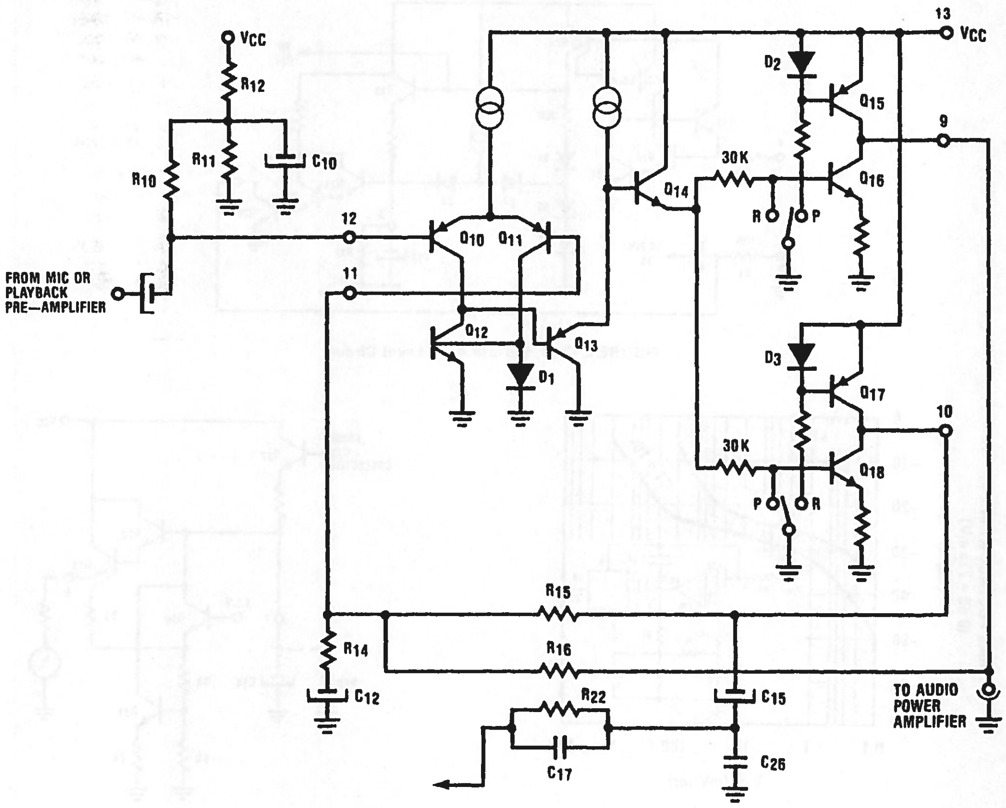


FIGURE 2.12.30 LM1818 Monitor and Output Amplifier With Feedback and Biasing Components

2

Referring to Figure 2.12.31, the input signal is rectified by the action of Q_{19}, Q_{20} and Q_{21}. Q_{20} and Q_{21} are taking most of the current from the 50µA current source so that Q_{19} is "on" just enough to provide base current. When a signal is applied at pin 4, the negative swings will cut-off Q_{20} and Q_{21} allowing Q_{19} to conduct. The current that then flows in the 2kΩ resistor at pin 4 is "mirrored" by Q_{22} and will develop a voltage across the 20kΩ resistor in series with diodes D_4 through D_6. In the absence of signal, a current source biases the filter capacitor C_{13} through diode D_8 to two diode voltage drops above ground. When the signal current in the 20kΩ resistor causes Q_{23} base to reach four diode voltage drops above ground Q_{26} can turn on. Q_{26} operates in a saturated mode and can sink or source current for small positive collector-emitter voltages. By connecting pin 5 to the microphone Q_{26} behaves as a variable resistor working against the 10kΩ resistor (R_1) in series with the microphone to attenuate the microphone signal when it is above the ALC threshold. Typical ALC response curves are shown in Figure 2.12.32.

The same half wave rectifier is used to supply an input to the meter drive circuit, Q_{27} through Q_{21} in Figure 2.12.33. Since Q_{27} base is connected back to the collector of Q_{22}, with no signal present, the meter memory capacitor is held two diode voltage drops above ground — therefore the output (pin 8) starts at ground. For small signals, the memory capacitor C_{14} is charged by Q_{27} and discharged by R_{18} and a constant 50µA discharge current in Q_{30}. This allows fast, accurate response in the lower portion of the meter range. When larger signals cause the output voltage at pin 8 to get above 0.7V, Q_{30} is shut off. This will normally correspond to a recording signal level at 0"VU" and the increased discharge time (since there is no 50µA discharge current and only R_{18} is discharging C_{14}) allows more time for high recording levels to be identified. Should the signal level increase further to around 1.0V at pin 8 (+3"VU" for example), Q_{31} becomes active and rapidly discharges C_{14} to prevent damage to the meter movement.

The meter calibration is performed by setting the series resistor to produce 0"VU" on the meter scale when the voltage at pin 8 is $0.7V_{DC}$. This voltage level is obtained with a $70mV_{RMS}$ signal at pin 4.

Example 2.12.3

Using a combined R/P cassette tape head with a response similar to that shown in Figure 2.12.22, design a portable monaural record-playback system to operate from 6V supplies. The head sensitivity in playback is $0.3mV_{RMS}$ for a 1kHz signal recorded −12dB below tape saturation, obtained with a 60µA record current level. An output of 250mW into an 8Ω speaker is required, with a system bandwidth from 80Hz to 10kHz. A microphone with a 1.8mV output level will be used.

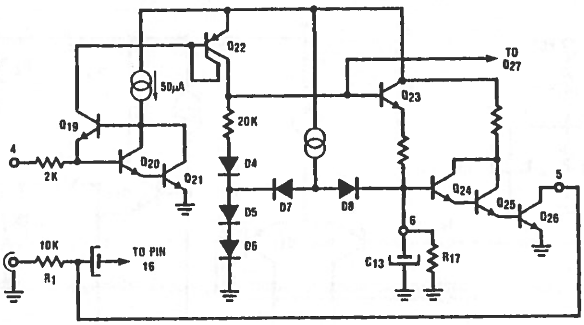


FIGURE 2.12.31 LM1818 Auto Level Circuit

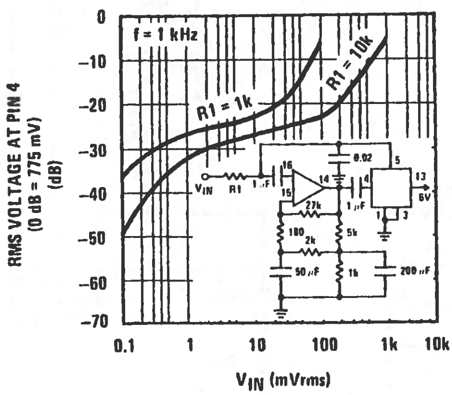


FIGURE 2.12.32 Automatic Level Control (ALC) Response Characteristics

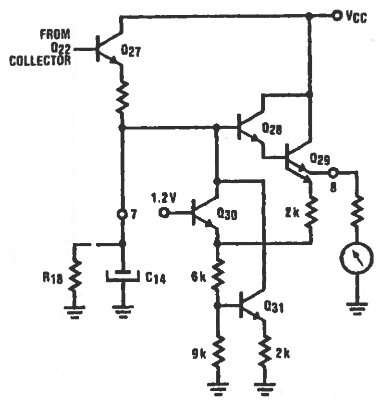


FIGURE 2.12.33 LM1818 Meter Drive Circuit

Solution

We will start by designing the playback system using the circuit configuration of Figure 2.12.29

1. From Equation (2.12.19)

$R_8 = 1k\Omega$

2. To set the playback preamp output (Pin 14) at half the supply voltage

$V_{DC} = \frac{6V}{2} = 3$ Volts

Using Equation (2.12.20)

$(R_5 + R_6) = (2V_{DC} - 1)R_8$
$= 5k\Omega$

3. The value of R_2 should be low to minimize the feedback path noise contribution, but not so low that C_7 is excessively large to maintain the low frequency response. A suitable value is

$R_2 = 180\Omega$

4. $C_7 = \frac{1}{2\pi \times 10 \times 180}$

$= 88.4 \times 10^{-6}$

Put $C_7 = 100\mu F$

5. When the level meter reads 0"VU", the signal level at pin 4 is $70mV_{RMS}$. For a typical cassette recorder, 0"VU" is usually chosen to leave about 3dB to 6dB headroom before tape saturation. For the given head sensitivity at 1kHz, −3dB corresponds to $0.85mV_{RMS}$.

Playback preamplifier gain at 1kHz

$= \frac{70}{0.85} = 82V/V$ or 38.4dB

The midband gain (above the equalization corner frequency f_2) for ferric tapes is 2.4dB below the 1kHz gain — Figure 2.12.24. Therefore the midband gain is 38.4 − 2.4 = 36dB or 63V/V.

Using Equation (2.12.21)

$63 = 1 + \frac{R_4}{R_2}$

$R_4 = 180 \times 62 \approx 12k\Omega$

6. Using Equation (2.12.23)

$f_2 = \frac{1}{2\pi C_5 R_4}$

$C_5 = \frac{1}{2\pi \times 1326 \times 12 \times 10^3}$

$= 1 \times 10^{-8} = 0.01\mu F$

7. Equation (2.12.22)

$f_1 = \frac{1}{2\pi C_5 R_3}$

$R_3 = \frac{1}{2\pi . 50 \times 0.01 \times 10^{-6}}$

$= 318k\Omega$

Put $R_3 = 330k\Omega$

8. To compensate for the playback head gap loss, C_4 is resonated with the playback head inductance to give a 4.5dB boost at 10kHz. The value of C_4 is best determined empirically using a calibrated test tape. Typically,

$C_4 = 470pF$

9. In the record mode, we would like the microphone to produce the same level at pin 4, so that for a given flux level on the tape, the meter will indicate the same reading both in record and playback.

Equation (2.12.25)

$A_{VAC(MIC)} = 1 + \frac{R_6}{R_5}$

$\frac{70}{1.8} = 1 + \frac{R_6}{R_5}$

$R_6 = 39R_5$

From step (2)

$R_5 + R_6 = 5k\Omega$
$\therefore R_5 = 120\Omega$
$R_6 = 4.7k\Omega$

10. The gain required in the monitor amplifier during playback will depend on the sensitivity of the audio power amplifier driving the speaker. Using the LM386 amplifier in the configuration shown in Figure 2.12.34 (see also Section 4.7), an input level of 100mVrms is required to obtain 250mW output power in an 8Ω load. To ensure adequate playback volume at lower recording levels the monitor amplifier should provide this level to the volume control with head output signals — 12dB below 0"VU". Using Equation (2.12.26)

$A_{V(MON)} = 6 = 1 + \frac{R_{16}}{R_{14}}$

$R_{16} = 5R_{14}$

Put $R_{14} = 50k$, $R_{16} = 250k\Omega$

11. For a quiescent output dc voltage of 3V

$R_{11} = R_{12} = 10k\Omega$

Put $R_{10} = 250k\Omega$ to balance input bias currents.

12. In the record mode, the monitor amplifier input from the microphone preamplifier is $70mV_{RMS}$ for 0"VU" on the meter. Since the monitor amplifier output can typically swing $1.65V_{RMS}$ on a 6V supply, Equation (2.12.28) gives the required gain as

$A_{V(REC)} = 1 + \frac{R_{15}}{R_{14}}$

$\frac{1.65}{70 \times 10^{-3}} = 1 + \frac{R_{15}}{R_{14}}$

$R_{15} = 23R_{14}$

$R_{15} = 1m\Omega$

2

13. A record current of 60μA produces a tape flux level −12dB below saturation. Therefore, for our specified 0"VU" level −3dB below saturation

$I_R = 170\mu A$

Equation (2.12.29)

$$R_{22} = \frac{1.65}{170 \times 10^{-6}} = 10k\Omega.$$

14. The high frequency cut-off is 10kHz. The head response begins to fall off at 2.5kHz and is −16dB at 10kHz. Since 4.5dB of this loss is compensated for on playback (see step 8), the remaining −12.5dB loss is compensated for during recording by shunting R_{22} with C_{17}.

Equation (2.12.30)

$$C_{17} = \frac{1}{2\pi . 2.5 \times 10^3 \times 10 \times 10^3}$$
$$= 0.0068\mu F$$

The complete schematic is shown in Figure 2.12.34 with the measured record/play response in Figure 2.12.35. The S/N ratio (CCIR/ARM weighted) is 59.5dB for the electronics, and 54dB with tape. Total current consumption (excluding tape drive and power amp) is 10.7mA in play, increasing to 45mA during record due to the bias oscillator.

No discussion of cassette recorder design would be complete without mention of the noise reduction systems that are a major contribution to the acceptability of the cassette format in higher quality applications. At present the most popular is the complementary noise reduction system developed by Dolby Laboratories — the consumer version of which is the Dolby® B-Type. The National LM1011/1011A is specifically designed to implement the functions of the Dolby B system, and as such is found in many commercially available cassette recorders. However, it should be emphasized that the use of the LM1011/1011A in Dolby systems is by license agreement with Dolby Labs* and LM1011's are available only to licensed manufacturers.

An alternative, non-complementary system suitable for cassette recorders is described in Section 5.8.

* License information available from Dolby Laboratories Licensing Corporation
731 Sansome St., San Fransisco, CA 94111

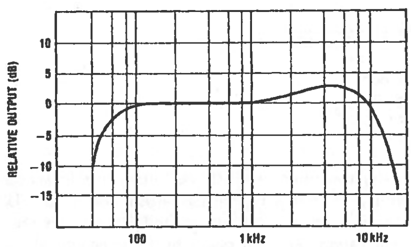


FIGURE 2.12.35 Record/Play Frequency Response

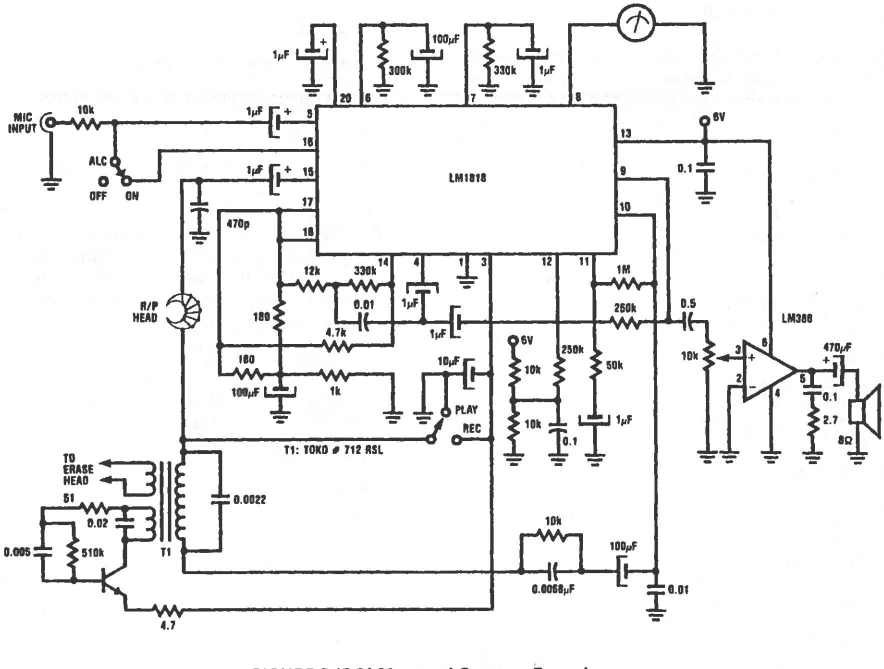


FIGURE 2.12.34 Monaural Cassette Recorder

2.13 MIC PREAMPS

2.13.1 Introduction

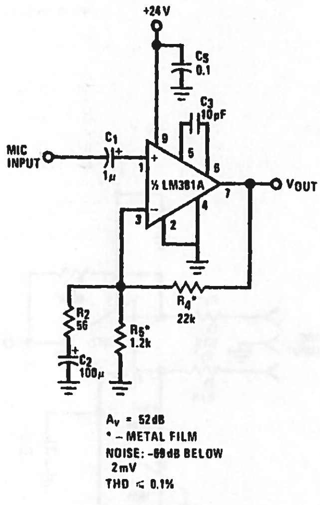
Microphones classify into two groups: high impedance (~20kΩ), high output (~20mV); and low impedance (~200Ω), low output (~2mV). The first category places no special requirements upon the preamp; amplification is done simply and effectively with the standard non-inverting or inverting amplifier configurations. The frequency response is reasonably flat and no equalization is necessary. Hum and noise requirements of the amplifier are minimal due to the large input levels. If everything is so easy, where is the hook? It surfaces with regard to hum and noise pickup of the microphone itself. Being a high impedance source, these mics are very susceptible to stray field pickup (e.g., 60Hz). Their use must be restricted to short distances (typically less than 10 feet of cable length), because of the potential high frequency roll-off caused by cable capacitance.

Low impedance microphones also have a flat frequency response, requiring no special equalization in the preamp section. Their low output levels do, however, impose rather stringent noise requirements upon the preamp. For a signal-to-noise ratio of 65dB with a 2mV input signal, the total equivalent input noise (EIN) of the preamp must be 1.12μV (10-10kHz). National's line of low noise dual preamps with their guaranteed EIN of ≤ 0.7μV (LM381A) and ≤ 0.9μV (LM387A) make excellent mic preamps, giving at least 67dB S/N (LM387A) performance (re: 2mV input level).

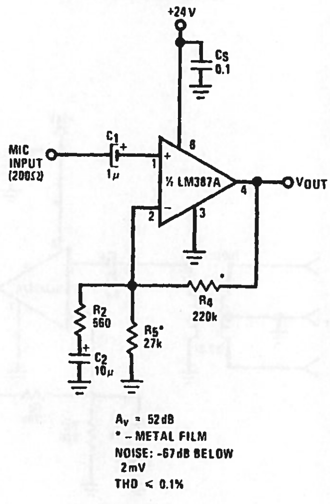
Low impedance mics take two forms: unbalanced two wire output, one of which is ground, and balanced three wire output, two signal and one ground. Balanced mics predominate usage since the three wire system facilitates minimizing hum and noise pickup by using differential input schemes. This takes the form of a transformer with a center-tapped primary (grounded), or use of a differential op amp. More about balanced mics in a moment, but first the simpler unbalanced preamps will be discussed.

2.13.2 Transformerless Unbalanced Designs

Low impedance unbalanced (or single-ended) mics may be amplified with the circuits appearing in Figure 2.13.1. The LM381A (Figure 2.13.1a) biased single-ended makes a simple, quiet preamp with noise performance −69dB below a 2mV input reference point. Resistors R_4 and R_5 provide negative input bias current and establish the DC output level at one-half supply. Gain is set by the ratio of R_4 to R_2, while C_2 establishes the low frequency −3dB corner. High frequency roll-off is done with C_3. Capacitor C_1 is made large to reduce the effects of 1/f noise currents at low



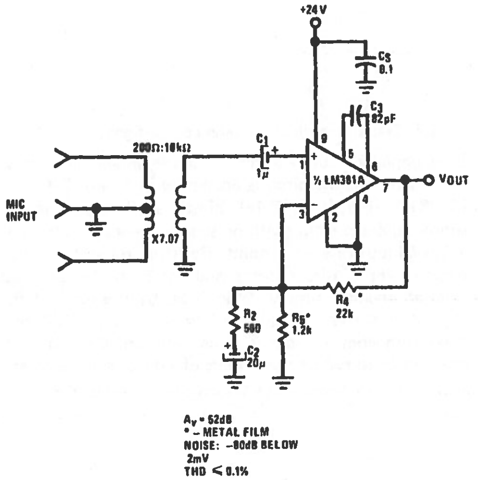
(a) LM381A S. E. Bias



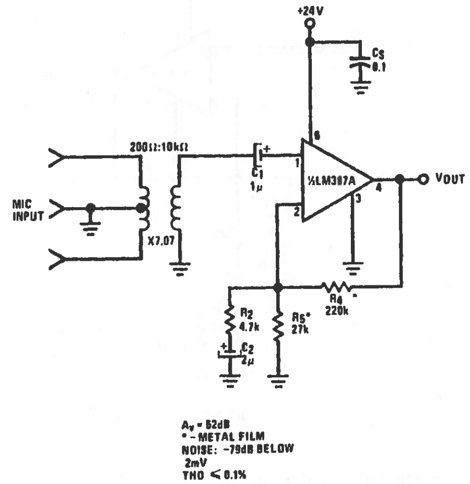
(b) LM387A

FIGURE 2.13.1 Transformerless Mic Preamps for Unbalanced Inputs frequencies. (See Section 2.6 for details on biasing and gain adjust.)

The LM387A (Figure 2.13.1b) offers the advantage of fewer parts and a very compact layout, since it comes in the popular 8-pin minidip package. The noise degradation referenced to the LM381A is only +2dB, making it a desirable alternative for designs where space or cost are dominant factors. Biasing and gain resistors are similar to LM381A. (See Section 2.8 for details.)



(a) LM381A S. E. Bias



(b) LM387A

FIGURE 2.13.2 Transformer-Input Mic Preamps for Balanced Inputs

2.13.3 Transformer-Input Balanced Designs

Balanced microphones are used where hum and noise must be kept at a minimum. This is achieved by using a three wire system — two for signal and a separate wire for ground. Proper grounding of microphones and their interconnecting cables is crucial since all noise and hum frequencies picked up along the way to the preamplifier will be amplified as signal. The rationale behind the twisted-pair concept is that all interference will be induced *equally* into each signal wire and will thus be applied to the preamp *common-mode*, while the actual transmitted signal appears differential. Balanced-input transformers with center-tapped primaries and single-ended secondaries (Figure 2.13.2) dominate balanced mic preamp designs. By grounding the center-tap all common-mode signals are shunted to ground, leaving the differential signal to be transformed across to the secondary winding, where it is converted into a single-ended output. Amplification of the secondary signal is done either with the LM381A (Figure 2.13.2a) or with the LM387A (Figure 2.13.2b). Looking back to Figure 2.13.1 shows the two circuits being the same with the exception of a change in gain to compensate for the added gain of the transformer. The net gain equals 52dB and produces 775mV output for a nominal 2mV input. Selection of the input transformer is fixed by two factors: mic impedance and amplifier optimum source impedance. For the cases shown the required impedance ratio is 200:10k, yielding a voltage gain (and turns ratio) of about seven ($\sqrt{10k/200}$).

Assuming an ideal noiseless transformer gives noise performance –80dB below a 2mV input level. Using a carefully designed transformer with electrostatic shielding, rejection of common-mode signals to 60dB can be expected (which is better than the cable manufacturer can match the twisting of the wires).

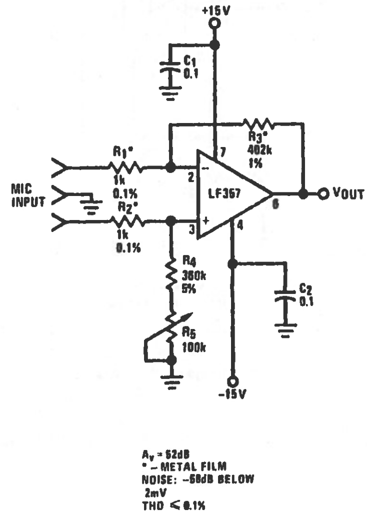


FIGURE 2.13.3 Transformerless Mic Preamp for Balanced Inputs

2.13.4 Transformerless Balanced Designs

Transformer input designs offer the advantage of nearly noise-free gain and do indeed yield the best noise performance for microphone applications; however, when the total performance of the preamplifier is examined, many deficiencies arise. Even the best transformers will introduce certain amounts of harmonic distortion; they are very susceptible to hum pickup; common-mode rejection is not optimum; and not a small problem is the expense of quality input transformers. For these reasons, transformerless designs are desirable. By utilizing the inherent ability of an operational amplifier to amplify differential signals while rejecting common-mode ones, it becomes possible to eliminate the input transformer.

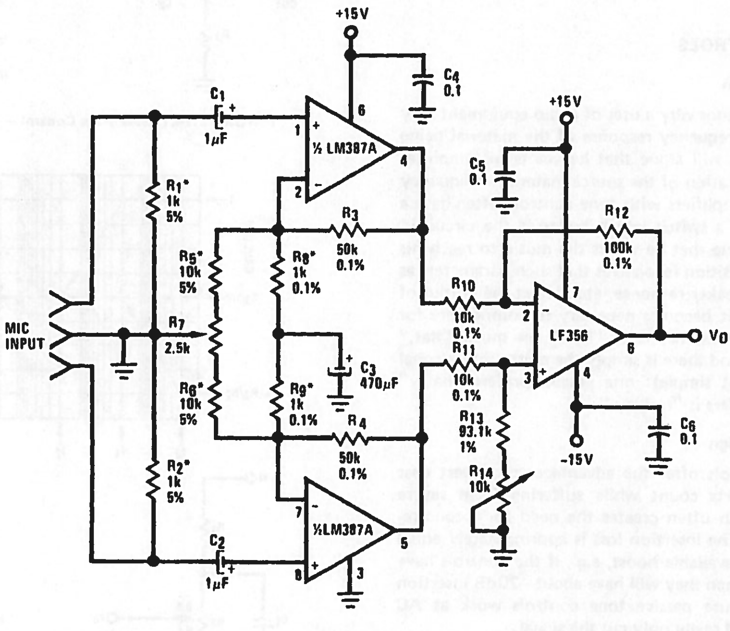
Figure 2.13.3 shows the FET input op amp, LF357 (selected for its high slew rate and CMRR) configured as a difference amplifier. As shown, with $R_1 = R_2$ and $R_3 = R_4 + R_5$ the gain is set by the ratio of R_3 to R_1 (see Appendix A4) and equals 52dB. The LF357 is selected over the quieter LM387A due to its high common-mode rejection capability. The LM387A (or LM381A) requires special circuitry when used with balanced inputs since it was not designed to reject common-mode signals. (A design trade-off was made for lower noise.) See Figure 2.13.4.

Input resistors R_1 and R_2 are made large compared to the source impedance, yet kept as small as possible, to achieve an optimum balance between input loading effects and low noise. Making $R_1 + R_2$ equal to ten times the source impedance is a good compromise value. Matching impedances is *not* conducive to low noise design and should be avoided.[1] The common-mode rejection ratio (CMRR) of the LF357 is 100dB and can be viewed as the "best case" condition, i.e., with a perfect match in resistors, the CMRR will be 100dB. The effect of resistor mismatch on CMRR cannot be overemphasized. The amplifier's ability to reject common-mode assumes that *exactly* the same signal is simultaneously present at both the inverting and non-inverting inputs (pins 2 and 3). Any mismatch between resistors will show up as a *differential* signal present at the input terminals and will be amplified accordingly. By using 0.1% tolerance resistors, and adjusting R_5 for minimum output with a common-mode signal applied, a CMRR near 100dB is possible. Using 1% resistors will degrade CMRR to about 80dB. The LF356 may be substituted for the LF357 if desired with only a degradation in slew rate (12V/µs vs. 50V/µs) and gain bandwidth (5MHz vs. 20MHz).

Due to the thermal noise of the relatively large input resistors the noise performance of the Figure 2.13.3 circuit is poorer than the other circuits, but it offers superior hum rejection relative to Figure 2.13.1 and eliminates the costly transformer of Figure 2.13.2.

2.13.5 Low Noise Transformerless Balanced Designs

A low noise transformerless design can be obtained by using a LM387A in front of the LF356 (or LF357) as shown in Figure 2.13.4. This configuration is known as an instrumentation amplifier after its main usage in balanced bridge instrumentation applications. In this design each half of the LM387A is



A_V = 54dB
* – METAL FILM
ADJ. R_7 FOR $V_{OUT} = 0V_{DC}$
ADJ. R_{14} FOR MAX CMRR
NOISE: –83dB BELOW 2mV INPUT
THD < 0.1%

FIGURE 2.13.4 Low Noise Transformerless Balanced Mic Preamp

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 2kΩ (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, a professional quality transformerless balanced mic preamp can be designed. The circuit is the same as Figure 2.13.4.

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.

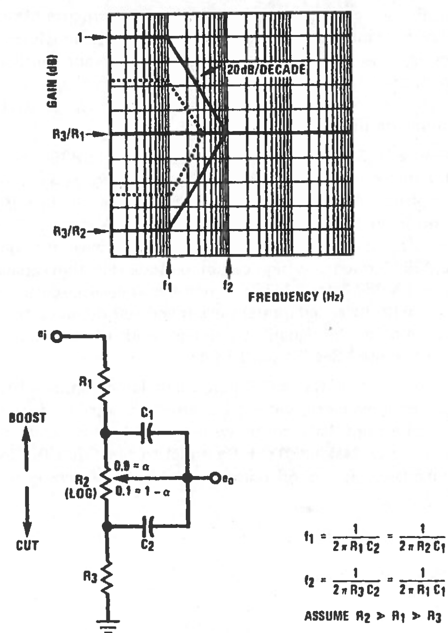


FIGURE 2.14.1 Bass Tone Control — General Circuit

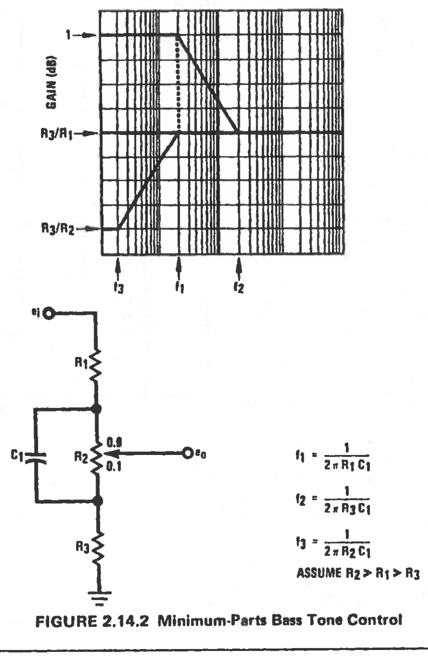


FIGURE 2.14.2 Minimum-Parts Bass Tone Control

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:

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

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 $R_2 \gg R_1 \gg R_3$):

$$C_1 = \frac{1}{2\pi f_2 R_1} \tag{2.14.2}$$

$$C_2 = \frac{1}{2\pi f_2 R_3} \tag{2.14.3}$$

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:

$$f_1 = \frac{1}{2\pi R_1 C_2} = \frac{1}{2\pi R_2 C_1} \tag{2.14.4}$$

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.

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

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:

$$C_1 = \frac{1}{2\pi f_2 R_3} \tag{2.14.6}$$

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 $X_{C1} = R_2$. 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:

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

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:

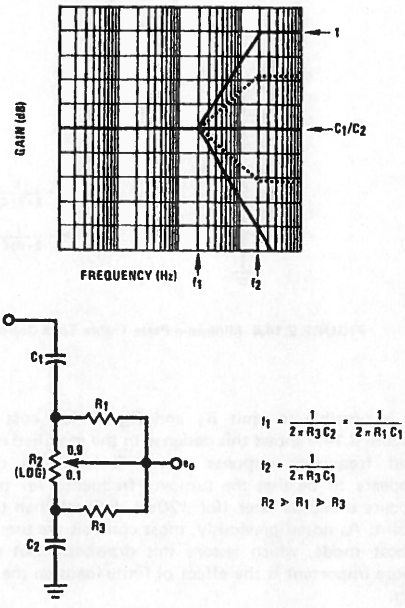


FIGURE 2.14.3 Treble Tone Control – General Circuit

$$C_1 = \frac{1}{2\pi f_1 R_1} \tag{2.14.8}$$

$$C_2 = \frac{1}{2\pi f_1 R_3} \tag{2.14.9}$$

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

$$f_2 = \frac{1}{2\pi R_3 C_1} \tag{2.14.10}$$

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.

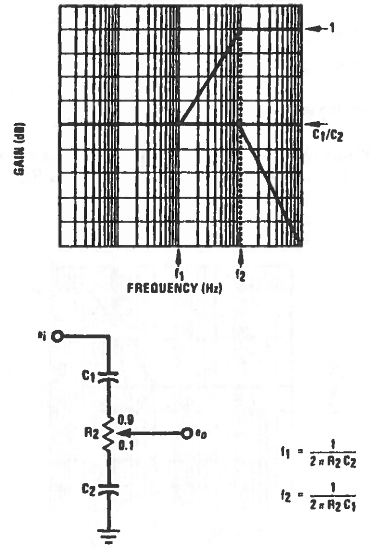


FIGURE 2.14.4 Minimum-Parts Treble Tone Control

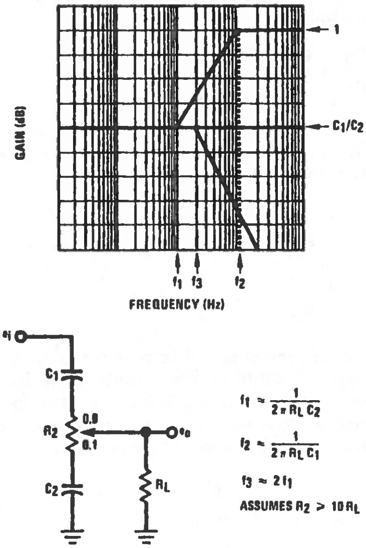


FIGURE 2.14.5 Effect of Loading Treble Tone Control

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 $R_2 \geqslant 10R_L$. 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):

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

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

3. Let R_2 = 100k (audio taper).

4. From Step 2:

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

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

5. From Equation (2.14.2) and Step 2:

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

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):

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

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

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_0 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_0 is selected to agree with system low frequency response:

$$C_0 = \frac{1}{(2\pi)(20Hz)(10k)} = 7.9 \times 10^{-7}$$

Use $C_0 = 1\mu F$.

2.14.5 Use of Passive Tone Controls with LM387 Preamp

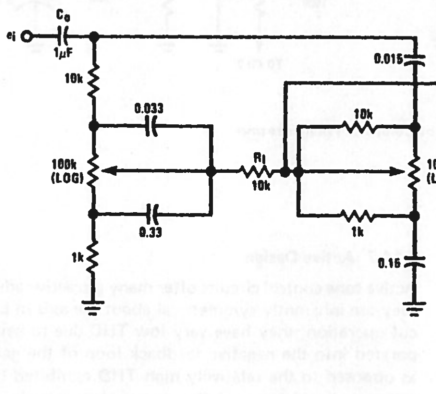
A typical application of passive tone controls (Figure 2.14.7) involves a discrete transistor used following the circuit to further amplify the signal as compensation for the loss through the passive circuitry. While this is an acceptable practice, a more judicious placement of the same transistor results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

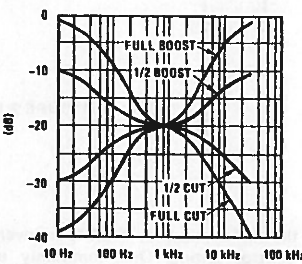
Placing the transistor *ahead* of the LM387 phono or tape preamplifier (Figure 2.14.8) improves the S/N ratio by boosting the signal before equalizing. An improvement of at least 3dB can be expected (analogous to operating a LM381A with single-ended biasing). The transistor selected must be low-noise, 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 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





Bass & Treble Tone Control Response

FIGURE 2.14.6 Complete Passive Bass & Treble Tone Control

2

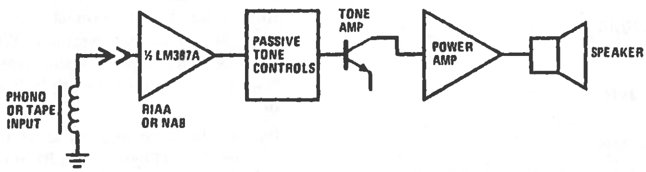


FIGURE 2.14.7 Typical Passive Tone Control Application

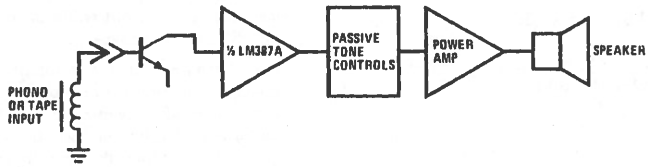


FIGURE 2.14.8 Improved Circuit Using Passive Tone Controls

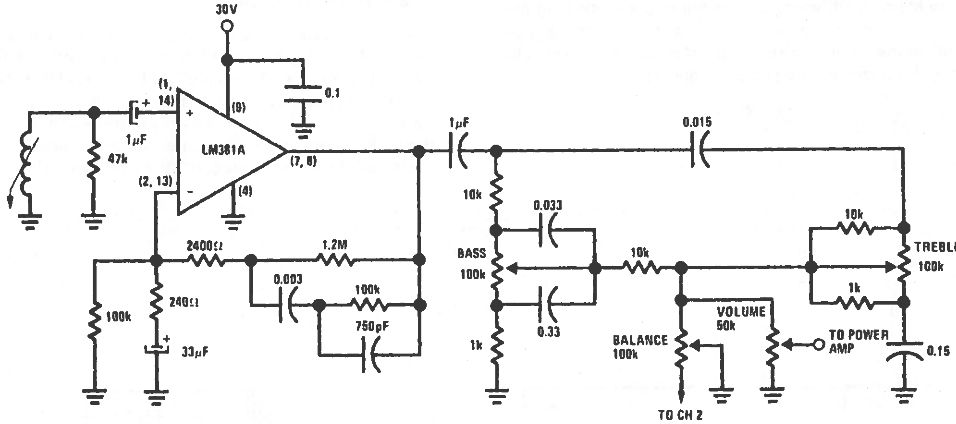


FIGURE 2.14.9 Single Channel of Complete Phono Preamp

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.

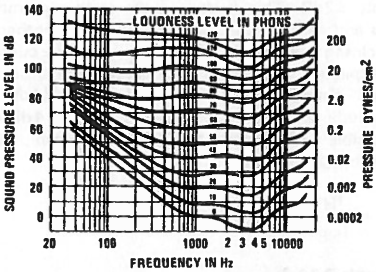


FIGURE 2.14.10 Fletcher-Munson Curves (USA). (Courtesy, Acoustical Society of America)

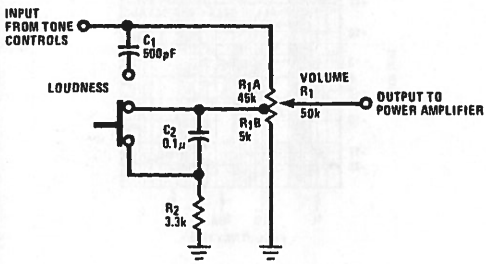


FIGURE 2.14.11 Loudness Control

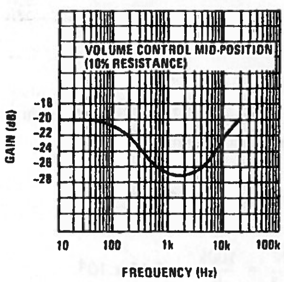


FIGURE 2.14.12 Loudness Control Frequency Response

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.

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

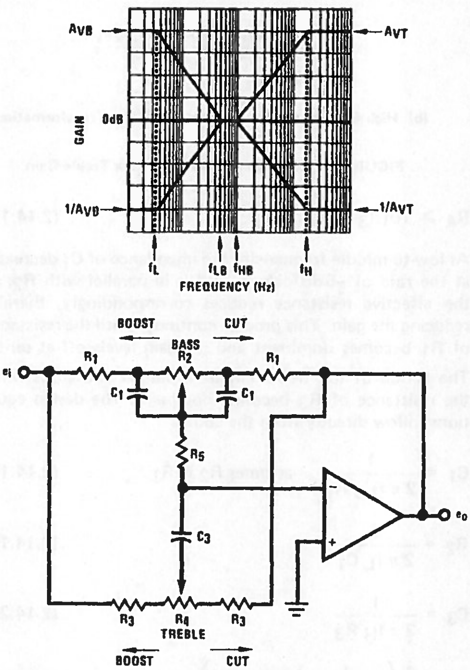
$$\left|\frac{1}{A_{VB}}\right| = \frac{R_1}{R_1 + R_2} \text{ (max bass cut)} \tag{2.14.14}$$

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.

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

$$\left|\frac{1}{A_{VT}}\right| = \frac{R_3}{R_3 + R_1 + 2R_5} \text{ (max treble cut)} \tag{2.14.16}$$

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

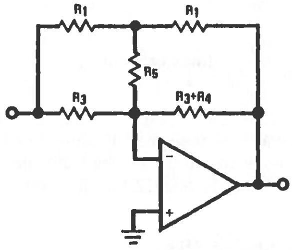
$$f_H = \frac{1}{2\pi R_3 C_3}$$

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

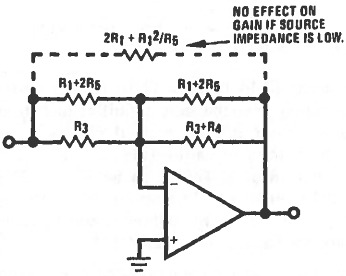
$$A_{VT} = 1 + \frac{R_1 + 2R_5}{R_3}$$

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

$$R_4 \geqslant 10(R_3 + R_1 + 2R_5) \qquad (2.14.17)$$

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.

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

$$R_2 = \frac{1}{2\pi f_L C_1} \qquad (2.14.19)$$

$$C_3 = \frac{1}{2\pi f_H R_3} \qquad (2.14.20)$$

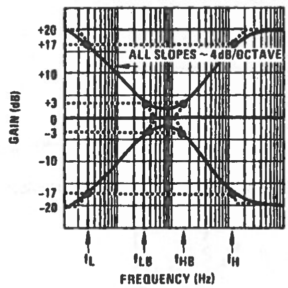
$$R_5 = \frac{1}{2}\left(\frac{1}{2\pi f_{HB} C_3} - R_1 - R_3\right) \qquad (2.14.21)$$

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).

$$\frac{f_{LB}}{f_L} = \frac{f_H}{f_{HB}} \approx 10 \qquad (2.14.22)$$

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):

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

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

$$R_1 = 11\text{k}$$

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

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

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

$$C_1 = 0.05\mu\text{F}$$

TREBLE DESIGN:

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

5. From Equation (2.14.15):

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

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

$$R_3 = 3.6k$$

6. Given f_H = 10kHz and from Equation (2.14.20):

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

$$C_3 = 0.005\mu F$$

7. From Equation (2.14.17):

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

$$\geqslant 10(3.6k + 11k + 22k)$$

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

$$R_4 = 500k$$

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% @ 775mV 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 is used inverting). Coupling capacitors C_4 and C_5 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_0 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

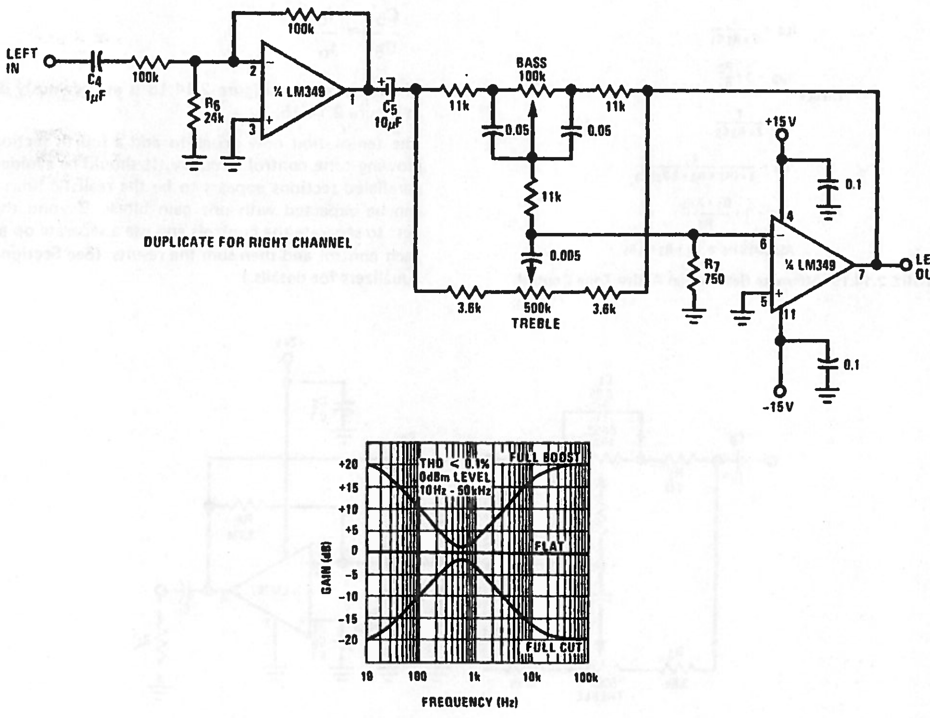


FIGURE 2.14.15 Typical Active Bass & Treble Tone Control with Buffer

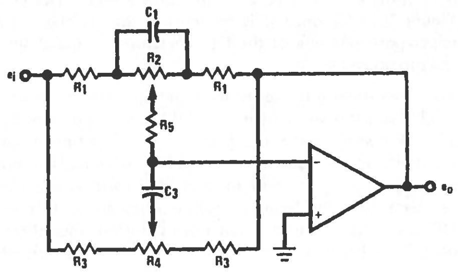
would cause "scratchiness." The R_7-C_3 network creates the input attenuation at high frequencies for stability.

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

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

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

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

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

FIGURE 2.14.16 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.)

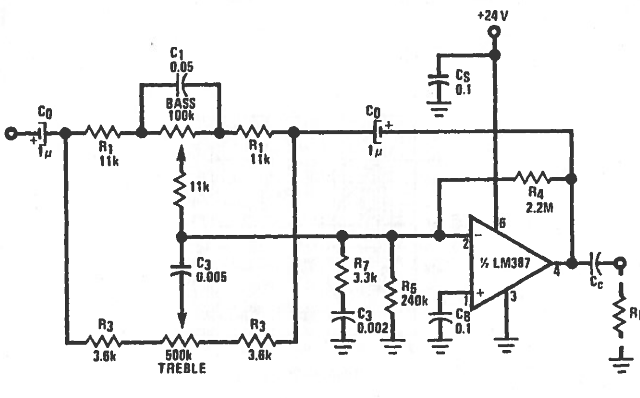


FIGURE 2.14.17 LM387 Feedback Tone Controls

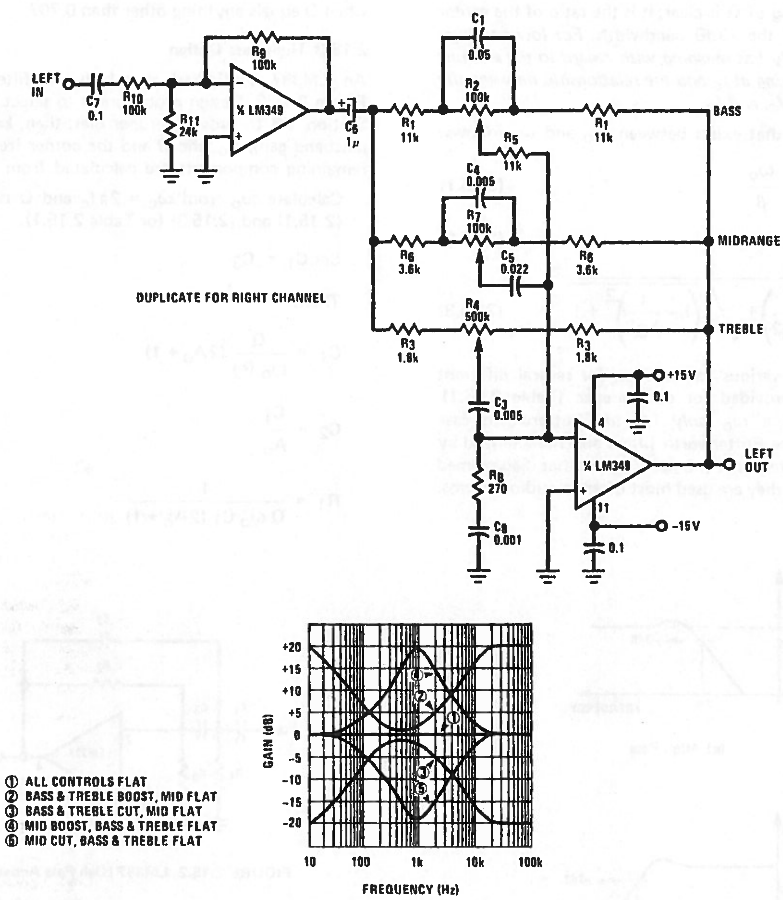


FIGURE 2.14.18 Three Band Active Tone Control (Bass, Midrange & Treble)

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 -3dB 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.*

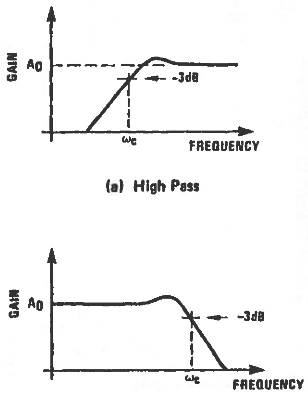
The relationship that exists between ω_o and ω_c follows:

$$\text{High Pass} \quad \omega_c = \frac{\omega_o}{\beta} \tag{2.15.1}$$

$$\text{Low Pass} \quad \omega_c = \beta\omega_o \tag{2.15.2}$$

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

A table showing various values of ω_c 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_c 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:

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

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

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

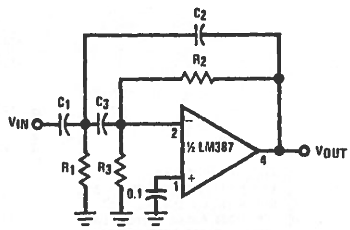
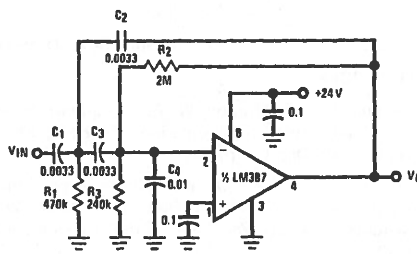


FIGURE 2.15.2 LM387 High Pass Active Filter



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 = 50Hz. Supply V_s = +24V.

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: $\quad C_2 = KC_1 \qquad (2.15.9)$

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 2.15.3, in cascade with a low pass based on Section 2.15.4, 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 $150\mu V$ and is

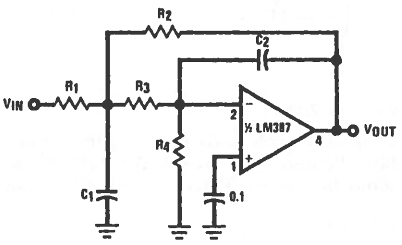


FIGURE 2.15.4 LM387 Low Pass Active Filter

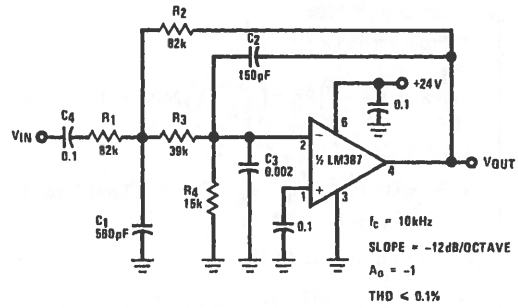


FIGURE 2.15.5 Scratch Filter Using LM387

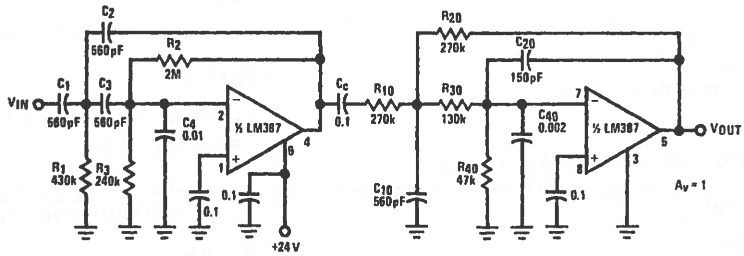


FIGURE 2.15.6 Speech Filter (300Hz-3kHz Bandpass)

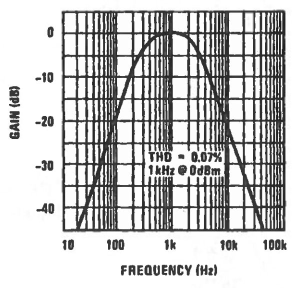


FIGURE 2.15.7 Speech Filter Frequency Response

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 2.15.5. A single amplifier bandpass filter using the LM387 (Figure 2.16.1) is capable of $Q \leqslant 10$ for audio frequency low distortion applications. The wide gain bandwidth (20MHz) and large open loop gain (104dB) allow high frequency, low distortion performance not obtainable 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:

$$R_1 = \frac{R_3}{2A_o} \quad (2.16.1)$$

$$C_1 = \frac{Q}{A_o \omega_o R_1} \quad (2.16.2)$$

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

For checking and trimming, use the following:

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

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

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

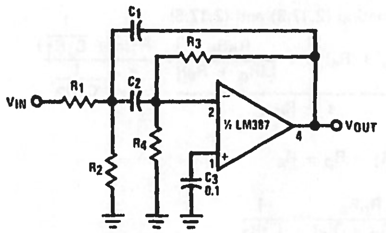


FIGURE 2.16.1 LM387 Bandpass Active Filter

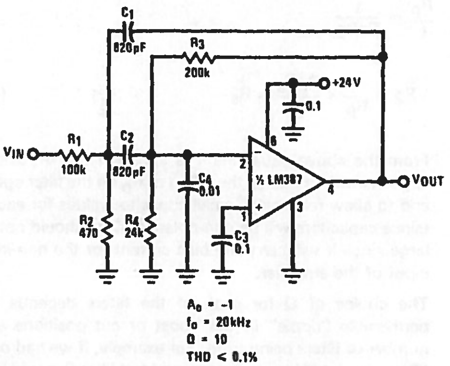


FIGURE 2.16.2 20kHz Bandpass Active Filter

Example 2.16.1

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

Solution

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

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

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

   Use R_3 = 200k

4. From Equation (2.16.1):

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

   R_1 = 100k

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 = 820pF

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 0.1µF for all designs. Input shunting capacitor C_4 is included for stability since the design gain is less than 10.

2

2.17 OCTAVE EQUALIZERS

2.17.1 Ten Band Octave Equalizer

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

A convenient ten band octave equalizer can be constructed based on the filter circuit shown in Figure 2.17.1 where the potentiometer R_2 can control the degree of boost or cut at the resonant frequency set by the series filter of C_2R_S and L, by varying the relative proportions of negative feedback and input signal to the amplifier section.

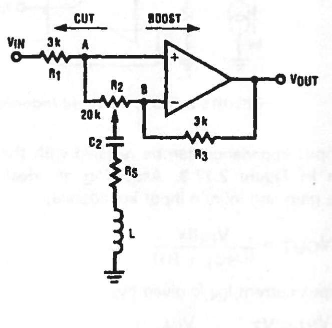


FIGURE 2.17.1 Typical Octave Equalizer Section

Assuming ideal elements, at the resonant frequency with R_2 slider set to the mid position, the amplifier is at unity gain. With the slider of R_2 moved such that C_2 is connected to the junction of R_1 and R_2, the R_sLC_2 network will attenuate the input such that

$$\frac{V_{OUT}}{V_{IN}} = \frac{R_s}{3k + R_s} \qquad (2.17.1)$$

If the slider is set to the other extreme, the gain at the resonant frequency is:

$$\frac{V_{OUT}}{V_{IN}} = \frac{3k + R_s}{R_s} \qquad (2.17.2)$$

In the final design, R_s is approximately 500Ω, giving a boost or attenuation factor of 7 (≅ 17dB). However, other filter sections of the equalizer connected between A and B will reduce this factor to about 12dB.

To avoid trying to obtain ten inductors ranging in value from 3.9H to 7.95mH for the ten octave from 32Hz to 16kHz, a simulated inductor design will be used. Consider the equivalent circuit of an inductor with associated series and parallel resistance as shown in Figure 2.12.2. The input impedance of the network is given by:

$$Z_{IN} = \frac{sLR_P}{(sL + R_p)} + R_s$$

$$= \frac{(R_p + R_s)\left(sL + \dfrac{R_pR_s}{R_p + R_s}\right)}{(sL + R_p)}$$

$$\therefore Z_{IN} = \frac{(R_p + R_s)\left(s + \dfrac{R_pR_s}{L(R_p + R_s)}\right)}{\left(s + R_p/L\right)} \qquad (2.17.3)$$

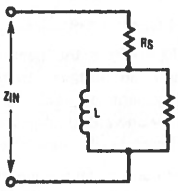


FIGURE 2.17.2 Ideal Inductor with Series and Parallel Resistances

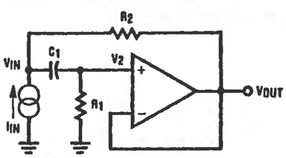


FIGURE 2.17.3 Simulated Inductor

This input impedance can be realised with the active circuit shown in Figure 2.17.3. Assuming an ideal amlifier with infinite gain and infinite input impedance,

$$V_2 = V_{OUT} = \frac{V_{IN}R_1}{(1/sC_1 + R_1)} \qquad (2.17.4)$$

The input current I_{IN} is given by,

$$I_{IN} = \frac{V_{IN} - V_2}{R_2} + \frac{V_{IN}}{(1/sC_1 + R_1)} \qquad (2.17.5)$$

Substituting (2.17.4) into this expression gives:

$$I_{IN} = V_{IN}\left\{\frac{1}{R_2} + \frac{sC_1}{(1 + sC_1R_1)} - \frac{R_1}{R_2\left(R_1 + \dfrac{1}{sC_1}\right)}\right\}$$

$$= V_{IN}\left\{\frac{1 + sC_1R_2}{R_2(sC_1R_1 + 1)}\right\}$$

Since $Z_{IN} = \dfrac{V_{IN}}{I_{IN}}$

$$Z_{IN} = \frac{R_2 + sC_1R_1R_2}{1 + sC_1R_2}$$

$$= \frac{R_1\left(\dfrac{1}{C_1R_1} + s\right)}{\left(\dfrac{1}{C_1R_2} + s\right)} \qquad (2.17.6)$$

Equating (2.17.3) and (2.17.6)

$$\frac{(R_p + R_s)\left[s + \dfrac{R_pR_s}{L(R_p + R_s)}\right]}{s + R_p/L} = \frac{R_1\left(s + \dfrac{1}{C_1R_1}\right)}{s + \dfrac{1}{C_1R_2}}$$

$$\therefore R_1 = R_p + R_s \qquad (2.17.7)$$

$$\frac{R_pR_s}{L(R_p + R_s)} = \frac{1}{C_1R_1}$$

$$\therefore C_1 = \frac{1}{R_pR_s} \qquad (2.17.8)$$

$$\frac{R_p}{L} = \frac{1}{C_1R_2}$$

$$\therefore R_2 = \frac{L}{R_p} \times \frac{R_pR_s}{L} = R_s \qquad (2.17.9)$$

From the above equations it is apparent that R_1 should be large in order to reduce the effect of R_p on the filter operation, and to allow reasonably small capacitor values for each band (since capacitors will be non-polarized). R_1 should not be too large since it will carry the bias current for the non-inverting input of the amplifier.

The choice of Q for each of the filters depends on the permissible "ripple" in the boost or cut positions and the number of filters being used. For example, if we had only two filters separated by one octave, an ideal filter Q would be 1.414 so that the −3dB response frequencies will coincide, giving the same gain as that at the band centers. For the ten band equalizer a Q of 1.7 is better, since several filters will be affecting the gain at a given frequency. This will keep the maximum ripple at full boost or cut to less than ± 2dB.

EXAMPLE 2.17.1

Design a variable (± 12dB) octave equalizer section with a Q of 1.7 and a center frequency of 2kHz.

Solution

1. Select R_1 = 68k
2. From equations (2.17.1) and (2.17.2) R_s = 470
3. $L = \dfrac{QR_s}{2\pi f_o} = \dfrac{QR_2}{2\pi f_o}$

$$\therefore L = \frac{1.7 \times 470}{2\pi \times 2 \times 10^3} = 63.6\text{mH} \qquad (2.17.10)$$

4. From equation (2.17.8)

$$C_1 = \frac{L}{R_p + R_s} = \frac{L}{(R_1 - R_2)R_2}$$

$$= \frac{63.6 \times 10^{-3}}{(68 \times 10^3 - 470)470}$$

$\therefore C_1 = 2000\text{pF}$

5. $$C_2 = \frac{1}{\omega_o^2 L}$$

$$= \frac{1}{(2\pi \times 2 \times 10^3)^2 63.5 \times 10^{-3}}$$

$$\therefore C_2 = 0.1\mu F \qquad (2.17.11)$$

Table 2.17.1 summarizes the component values required for the other sections of the equalizer. The final design appears in Figure 2.17.4 and uses LM348 quad op-amps. Other unity gain stable amplifiers can be used. For example, LF356 will give lower distortion at the higher frequencies. Although linear taper potentiometers can be used, these will result in very rapid action near the full boost or full cut positions. S taper

TABLE 2.17.1

fo(Hz)	C_1	C_2	R_1	R_2
32	0.12μF	4.7μF	75kΩ	560Ω
64	0.056μF	3.3μF	68kΩ	510Ω
125	0.033μF	1.5μF	62kΩ	510Ω
250	0.015μF	0.82μF	68kΩ	470Ω
500	8200pF	0.39μF	62kΩ	470Ω
1k	3900pF	0.22μF	68kΩ	470Ω
2k	2000pF	0.1μF	68kΩ	470Ω
4k	1100pF	0.056μF	62kΩ	470Ω
8k	510pF	0.022μF	68kΩ	510Ω
16k	330pF	0.012μF	51kΩ	510Ω

potentiometers (Allen Bradley #70A1G032 R2035) will give a better response. All the capacitors used for tuning the simulated inductors (C_2) should be non-polarized mylar or polystyrene.

Signal to noise ratio of the equalizer with the controls set "flat" is 73dB referred to a $1V_{RMS}$ input signal. THD is under 0.01% at 20kHz.

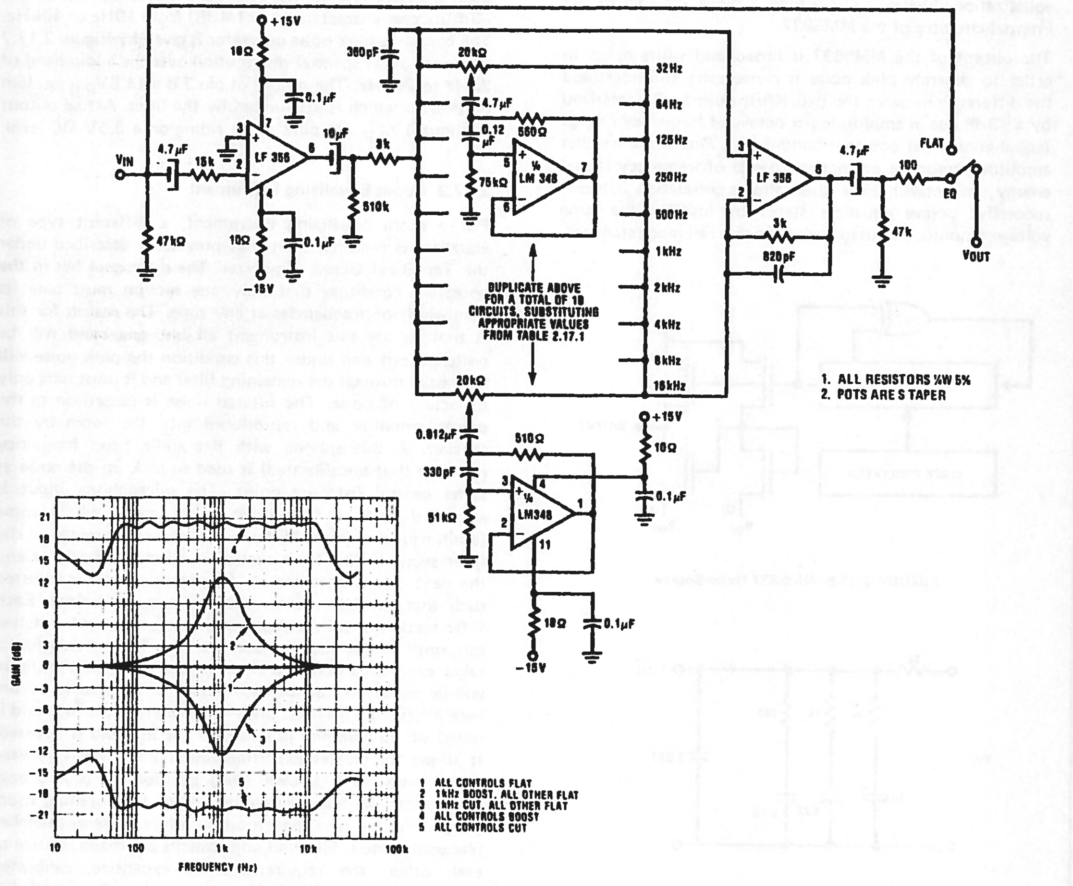


FIGURE 2.17.4 Complete Ten Band Octave Equalizer

2.17.2 Pink Noise Generator

Once an equalizer is incorporated into a music system the question quickly arises as to how best to use it. The most obvious way is as a "super tone control" unit, where control is now extended from the familiar two or three controls to ten controls (or even 30 if 1/3 octave equalizers are used). While this approach is most useful and the results are dramatic in their ability to "liven" up a room, there still remains, with many, the desire to have some controlled manner in which to equalize the listening area without resorting to the use of expensive (and complicated) spectrum or real-time analyzers.

The first step in generating a self-contained room equalizing instrument is to design a pink noise generator to be used as a controlled source of noise across the audio spectrum. With the advent of medium scale integration and MOS digital technology, it is quite easy to create a pink noise generator using only one IC and a few passive components.

The MM5837 digital noise source is an MOS/MSI pseudo-random sequence generator, designed to produce a broad-band white noise signal for audio applications. Unlike traditional semiconductor junction noise sources, the MM5837 provides very uniform noise quality and output amplitude. Originally designed for electronic organ and synthesizer applications, it can be directly applied to room equalization. Figure 2.17.5 shows a block diagram of the internal circuitry of the MM5837.

The output of the MM5837 is broadband white noise. In order to generate pink noise it is necessary to understand the difference between the two. *White noise* is characterized by a +3dB rise in amplitude per octave of frequency change (equal energy per constant bandwidth). *Pink noise* has flat amplitude response per octave change of frequency (equal energy per octave). Pink noise allows correlation between successive octave equalizer stages by insuring the same voltage amplitude is used each time as a reference standard.

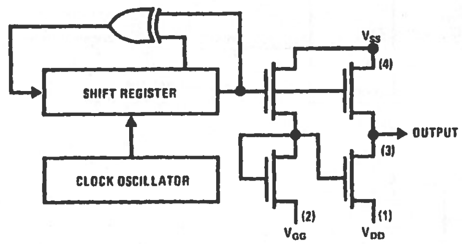


FIGURE 2.17.5 MM5837 Noise Source

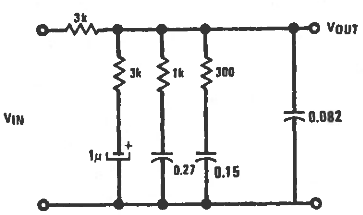


FIGURE 2.17.6 Passive -3dB/Octave Filter

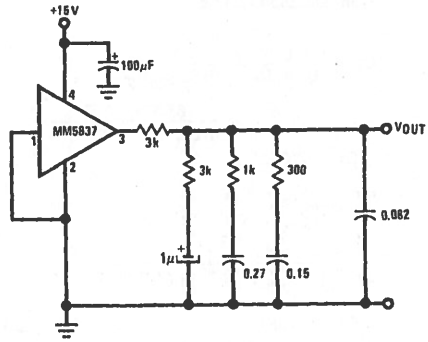
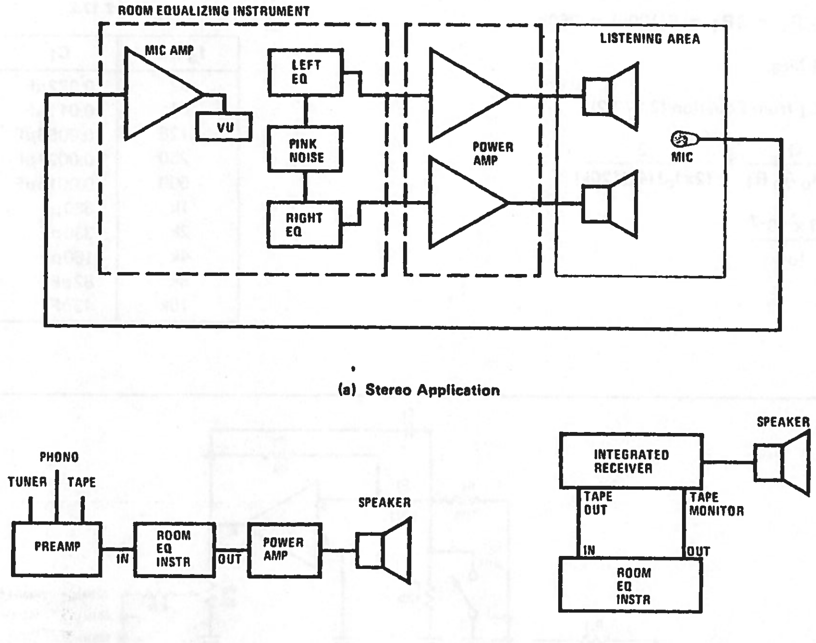


FIGURE 2.17.7 Pink Noise Generator

What is required to produce pink noise from a white noise source is simply a -3dB/octave filter. If capacitive reactance varies at a rate of -6dB/octave then how can a slope of *less* than -6dB/octave be achieved? The answer is by cascading several stages of lag compensation such that the zeros of one stage partially cancel the poles of the next stage, etc. Such a network is shown as Figure 2.17.6 and exhibits a -3dB/octave characteristic (±1/4dB) from 10Hz to 40kHz. The complete pink noise generator is given by Figure 2.17.7 and gives a flat spectral distribution over the audio band of 20Hz to 20kHz. The output at pin 3 is a $11.5V_{p\text{-}p}$ random pulse train which is attenuated by the filter. Actual output is about $1V_{p\text{-}p}$ AC pink noise riding on a 8.5V DC level.

2.17.3 Room Equalizing Instrument

For a room equalizing instrument, a different type of equalizer is required than that previously described under the Ten Band Octave Equalizer. The difference lies in the necessary condition that only one section must pass its bandwidth of frequencies at any time. The reason for this is that to use this instrument all but one band will be switched out and under this condition the pink noise will be passed through the remaining filter and it must pass only its octave of noise. The filtered noise is passed on to the power amplifier and reproduced into the room by the speaker. A microphone with flat audio band frequency response (but uncalibrated) is used to pick up the noise at some central listening point. The microphone input is amplified and used to drive a VU meter where some (arbitrary) level is established via the potentiometer of the filter section. This filter section is then switched out and the next one is switched in. Its potentiometer is adjusted such that the VU meter reads the same as before. Each filter section in turn is switched in, adjusted, and switched out, until all ten octaves have been set. The whole process takes about two minutes. When finished the room response will be equalized flat for each octave of frequencies. From here it becomes personal preference whether the high end is rolled off (a common practice) or the low end is boosted. It allows for greater experimentation since it is very easy to go back to a known (flat) position. It is also easy to correct for new alterations within the listening room (drape changes, new rugs, more furniture, different speaker placement, etc.). Since all adjustments are made relative to each other, the requirement for expensive, calibrated microphones is obviated. Almost any microphone with flat output over frequency will work.



(a) Stereo Application

(b) Adding EQ to Component System

(c) Adding EQ to Receiver System

FIGURE 2.17.8 Typical Equalizing Instrument Application

For stereo applications, a two channel instrument is required as diagrammed in Figure 2.17.8a. Figures 2.17.8b and -c show typical placement of the equalizer unit within existing systems.

While any bandpass filter may be used for the filter sections, the multiple-feedback, infinite-gain configuration of Figure 2.17.9 is chosen for its low sensitivity factors. The design equations appear as follows:

$$R_1 = \frac{Q}{2\pi f_o A_o C_1} \tag{2.17.12}$$

$$R_2 = \frac{Q}{(2Q^2 - A_o)2\pi f_o C_1} = \frac{A_o R_1}{2Q^2 - A_o} \tag{2.17.13}$$

$$R_3 = \frac{Q}{\pi f_o C_1} \tag{2.17.14}$$

$$A_o = \frac{R_3}{2R_1} \tag{2.17.15}$$

$$Q = \pi f_o C_1 R_3 \tag{2.17.16}$$

$$f_o = \frac{1}{2\pi C_1}\sqrt{\frac{R_1 + R_2}{R_1 R_2 R_3}} \tag{2.17.17}$$

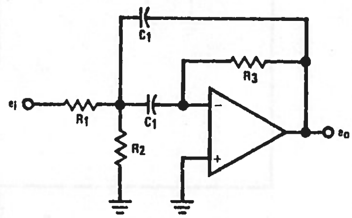


FIGURE 2.17.9 Bandpass Filter Section

Design

1. Select $A_o = 4$ (12dB) and $Q = 2$.
2. Select R_1 for desired input resistance. (Note that net input impedance is $(R_1 + R_2)/10$, since there are 10 sections in parallel.)

   Let $R_1 = 120k$.
3. Calculate R_2 from Equations (2.17.13) and (2.17.12):

$$R_2 = \frac{Q}{(2Q^2 - A_o)2\pi f_o C_1} = \frac{Q}{[2(2)^2 - 4]\,2\pi f_o C_1}$$

$$= \frac{Q}{(4)2\pi f_o C_1} = \frac{Q}{2\pi f_o A_o C_1} = R_1$$

$$R_2 = R_1 = 120k$$

4. Calculate R_3 from Equation (2.17.15).

$$R_3 = 2A_oR_1 = 8R_1 = 8(120k) = 960k$$

Use R_3 = 1 Meg.

5. Calculate C_1 from Equation (2.17.12):

$$C_1 = \frac{Q}{2\pi f_o A_o R_1} = \frac{2}{(2\pi f_o)(4)(120k)}$$

$$C_1 = \frac{6.63 \times 10^{-7}}{f_o}$$

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

TABLE 2.17.2

f_o (Hz)	C_1
32	0.022µF
64	0.011µF
125	0.0056µF
250	0.0027µF
500	0.0015µF
1k	680pF
2k	330pF
4k	160pF
8k	82pF
16k	43pF

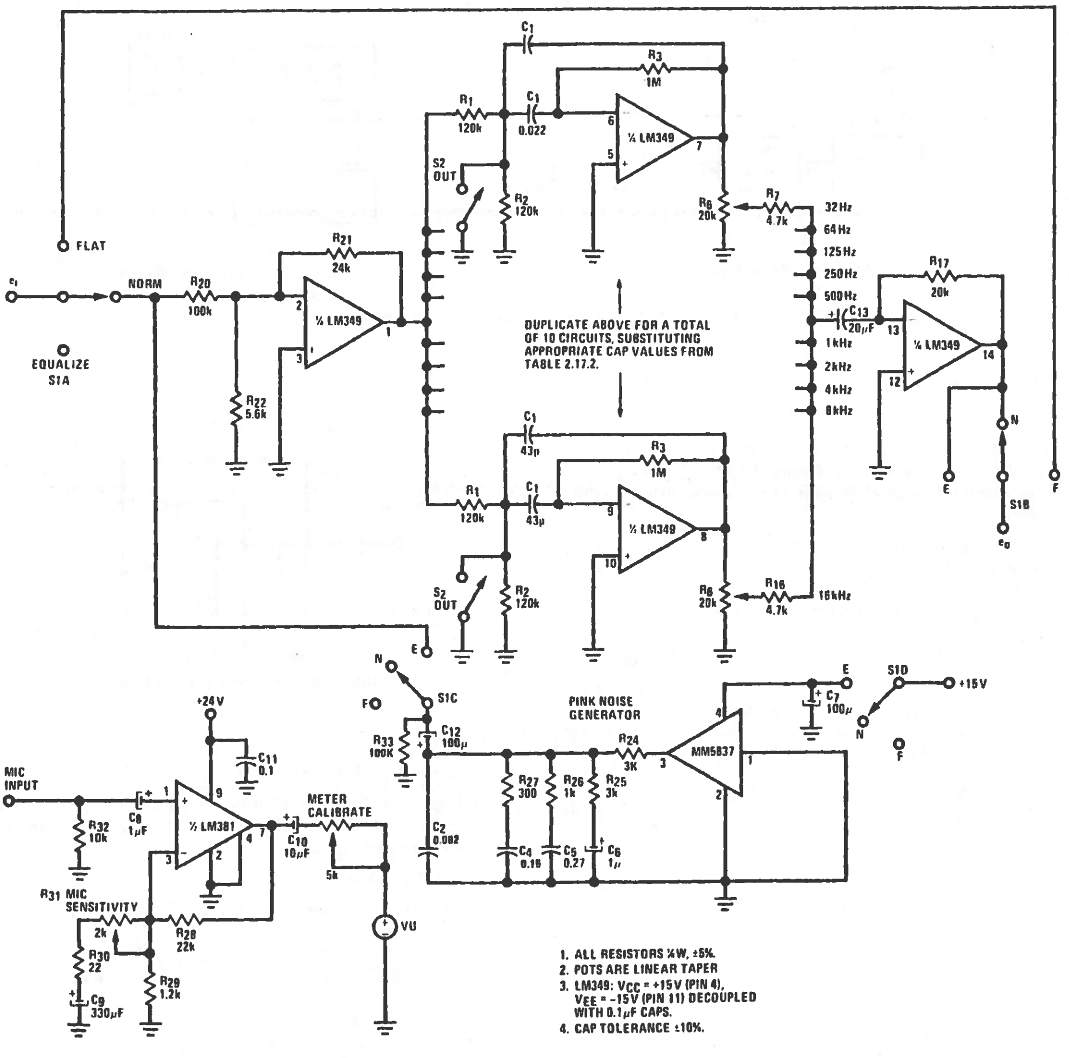


FIGURE 2.17.10 Room Equalizing Instrument

The complete room equalizing instrument appears as Figure 2.17.10. The input buffer and output summer are similar to those that appear in Figure 2.17.2, with some important differences. The input buffer acts as an active attenuator with a gain of 0.25 and the output summer has variable gain as a function of slider position. The purpose of these features is to preserve unity gain through a system that is really "cut-only" (since the gain of each filter section is fixed and the output is dropped across the potentiometers). The result is to create a boost and cut effect about the mid-point of the pot which equals unity gain. To see this, consider just one filter section, and let the input to the system equal 1V. The output of the buffer will be 0.25V and the filter output at the top of potentiometer R_6 will again be 1V (since $A_O = 4$). The gain of the summer is given by $R_{17}/R_7 \approx 4$ when the slider of R_6 is at maximum, so the output will be equal to 4V, or +12dB relative to the input. With the slider at midposition the 4.7k summer input resistor R_7 effectively parallels 1/2 of R_6 for a net resistance from slider to ground of $4.7k \| 10k \approx 3.2k$. The voltage at the top of the pot is attenuated by the voltage divider action of the 10kΩ (top of pot to slider) and the 3.2kΩ (slider to ground). This voltage is approximately equal to 0.25V and is multiplied by 4 by the summer for a final output voltage of 1V, or 0dB relative to the input. With the slider at minimum there is no output from this section, but the action of the "skirts" of the adjacent filters tends to create −12dB cut relative to the input. So the net result is a ±12dB boost and cut effect from a cut only system.

The pink noise generator from Figure 2.17.7 is included as the noise source to each filter section only when switch S1 (3 position, 4 section wafer) is in the "Equalize" position. Power is removed from the pink noise generator during normal operation so that noise is not pumped back onto the supply lines. Switch S2 located on each filter section is used to ground the input during the equalizing process. The LM381 dual low noise preamplifier is used as the microphone amplifier to drive the VU meter. The second channel is added by duplicating all of Figure 2.17.9 with the exception of the pink noise generator which can be shared. Typical frequency response is given by Figure 2.17.11. While the system appears complex, a complete two-channel instrument is made with just 8 ICs (6-LM349, 1-LM381, and 1-MM5837).

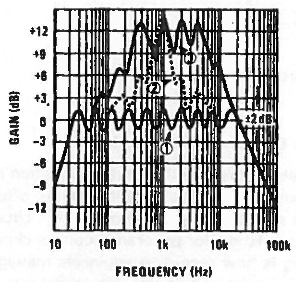


FIGURE 2.17.11 Typical Frequency Response of Room Equalizer

For detailed discussions about room equalization, the interested reader is directed to the references that follow this section.

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2

2.18 MIXERS

2.18.1 Introduction

A microphone mixing console or "mixer" is an accessory item used to combine the outputs of several microphones into one or more common outputs for recording or public address purposes. They range from simple four input-one output, volume-adjust-only units to ultra-sophisticated sixteen channel, multiple output control centers that include elaborate equalization, selective channel reverb, taping facilities, test oscillators, multi-channel panning, automatic mix-down with memory and recall, individual VU meters, digital clocks, and even a built-in captain's chair. While appearing complex and mysterious, mixing consoles are more repetitious than difficult, being constructed from standard building-block modules that are repeated many times.

2.18.2 Six Input-One Output Mixer

A detailed analysis of all aspects of mixer design lies beyond the scope of this book; however, as a means of introduction to the type of design encountered Figure 2.18.1 is included to show the block diagram of a typical six input-one output mixer. Below each block, the section number giving design details is included in parentheses for easy cross reference.

Individual level and tone controls are provided for each input microphone, along with a choice of reverb. All six channels are summed together with the reverb output by the master summing amplifier and passed through the master level control to the octave equalizer. The output of the equalizer section drives the line amplifier, where monitoring is done via a VU meter.

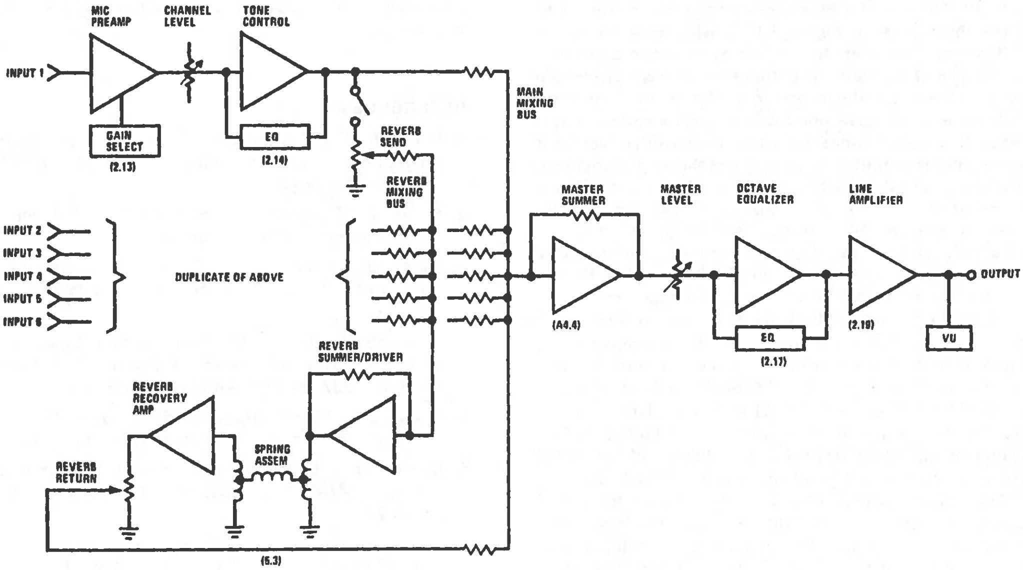


FIGURE 2.18.1 Six Input-One Output Microphone Mixing Console (Design details given in sections shown in parentheses.)

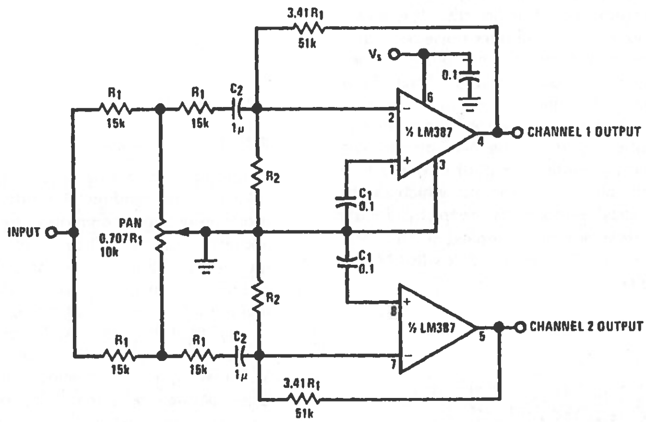


FIGURE 2.18.2 Two Channel Panning Circuit

Expansion of the system to any number of inputs requires only additional input modules, with the limiting constraint being the current driving capability of the summing amplifiers. (The summing amp must be capable of sourcing and sinking the *sum* of all of the input amplifiers driving the summing bus. For example, consider ten amplifiers, each driving a 10kΩ summing input resistor to a maximum level of $5V_{RMS}$. The summing amplifier is therefore required to handle 5mA.) Expanding the number of output channels involves adding additional parallel summing busses and amplifiers, each with separate level, equalizer, and VU capabilities. Other features (test oscillator, pink noise generator, panning, etc.) may be added per channel or per console as required.

2.18.3 Two Channel Panning Circuit

Having the ability to move the apparent position of one microphone's input between two output channels often is required in recording studio mixing consoles. Such a circuit is called a panning circuit (short for panoramic control circuit) or a pan-pot. Panning is how recording engineers manage to pick up your favorite pianist and "float" the sound over to the other side of the stage and back again. The output of a pan circuit is required to have unity gain at each extreme of pot travel (i.e., all input signal delivered to one output channel with the other output channel zero) and −3dB output from each channel with the pan-pot centered. Normally panning requires two oppositely wound controls ganged together; however, the circuit shown in Figure 2.18.2 provides smooth and accurate panning with only one linear pot. With the pot at either extreme the effective negative input resistor equals $3.41R_1$ (see Appendix A3.1) and gain is unity. Centering the pot yields an effective input resistor on each side equal to $4.83R_1$ and both gains are −3dB. The net input impedance as seen by the input equals $0.6R_1$, independent of pan-pot position. Using standard 5% resistor values as shown in Figure 2.18.2, gain accuracies within 0.4dB are possible; replacing R_1 with 1% values (e.g., input resistors equal 14.3kΩ and feedback resistors equal 48.7kΩ) allows gain accuracies of better than 0.1dB. Biasing resistor R_2 is selected per section 2.8 as a function of supply voltage. Capacitor C_1 is used to decouple the positive input, while C_2 is included to prevent shifts in output DC level due to the changing source impedance.

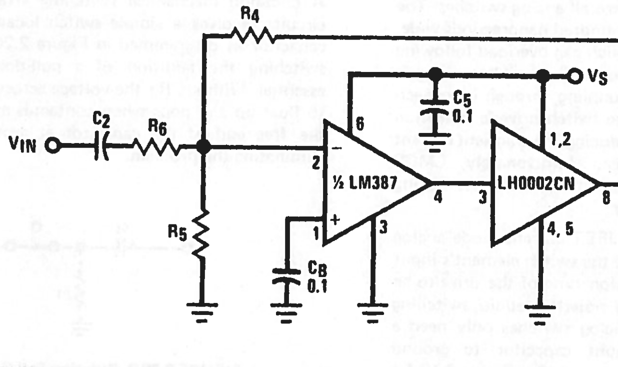


FIGURE 2.19.1 Preamp Current Booster

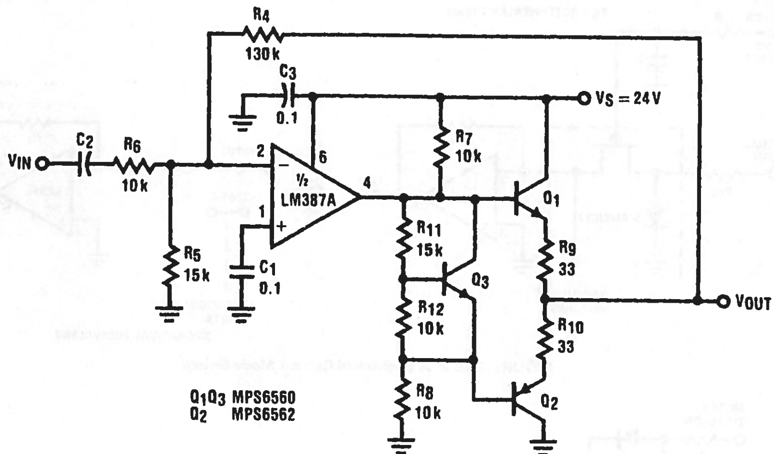


FIGURE 2.19.2 Discrete Current Booster Design

2.19 DRIVING LOW IMPEDANCE LINES

The output current and drive capability of a preamp may be increased for driving low impedance lines by incorporating a LH0002CN current amplifier within the feedback loop (Figure 2.9.1). Biasing and gain equations remain unchanged and are selected per section 2.8. Output current is increased to a maximum of ±100mA, allowing a LM387 to drive a 600Ω line to a full 24dBm when operated from a +36V supply. Insertion of the LH0002C adds less than 10 degrees additional phase shift at 15MHz, thereby not appreciably affecting the stability of the LM387 ($A_V \geq 10$).

Comparable performance can be obtained with the discrete design of Figure 2.19.2 for systems where parts count is not critical. Typical measured characteristics show a bandwidth of 15Hz-250kHz at +10dBm output, with THD below 0.02% up to 20kHz. A maximum output level of +16dBm can be obtained before clipping.

2.20 NOISELESS AUDIO SWITCHING

2.20.1 Active Switching

As prices of mechanical switches continue to increase, solid state switching element costs have decreased to the point where they are now cost effective. By placing the switch on the PC board instead of the front panel, hum pickup and crosstalk are minimized, while at the same time replacing the complex panel switch assemblies.

The CMOS transmission gate is by far the cheapest solid state switching element available today, but it is plagued with spiking when switched, as are all analog switches. The switching spikes are only a few hundred nanoseconds wide, but a few volts in magnitude, which can overload following audio stages, causing audible pops. The switch spiking is caused by the switch's driver coupling through its capacitance to the load. Increasing the switch driver's transition time minimizes the spiking by reducing the transient current through the switch capacitance. Unfortunately, CMOS transmission gates do not have the drivers available, making them less attractive for audio use.

Discrete JFETs and monolithic JFET current mode analog switches such as AM97C11 have the switch element's input available. This allows the transition time of the drive to be tailored to any value, making noiseless audio switching possible. The current mode analog switches only need a simple series resistor and shunt capacitor to ground between the FET switch and the driver. (See Figure 2.10.1.)

Discrete JFETs may be used in place of the quad current mode switch; or, they can be used as voltage mode switches at a savings to the amplifier but at the expense of additional resistors and a diode.

Driver rise times shown in the figures, in the 1-10ms range, will result in coupled voltage spikes of only a few mV when used with the typical impedances found in audio circuits.

2.20.2 Mechanical Switching

A common mechanical switching arrangement for audio circuits involves a simple switch located after a coupling capacitor as diagrammed in Figure 2.20.3. For "pop" free switching the addition of a pull-down resistor, R_1, is essential. Without R_1 the voltage across the capacitor tends to float up and pops when contact is made again; R_1 holds the free end of the capacitor at ground potential, thus eliminating the problem.

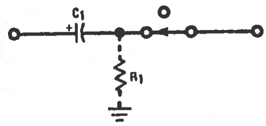


FIGURE 2.20.3 Capacitor Pull-Down Resistor

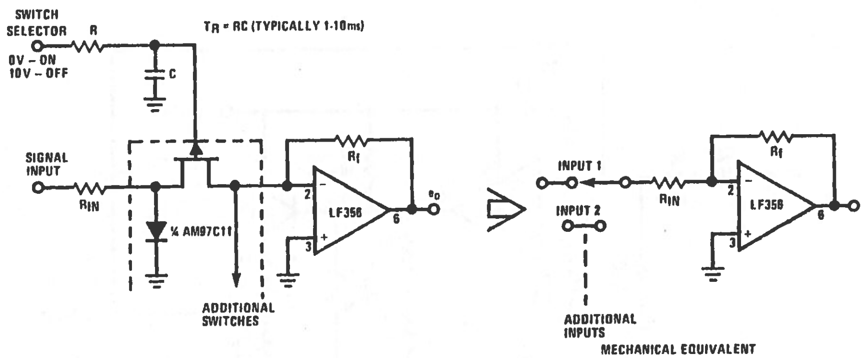


FIGURE 2.20.2 A Deglitched Current Mode Switch

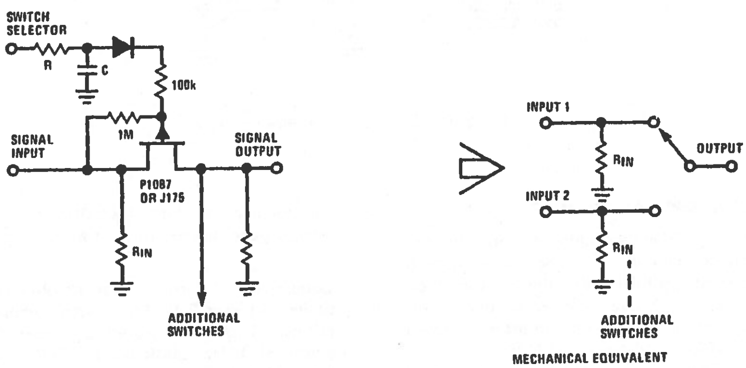
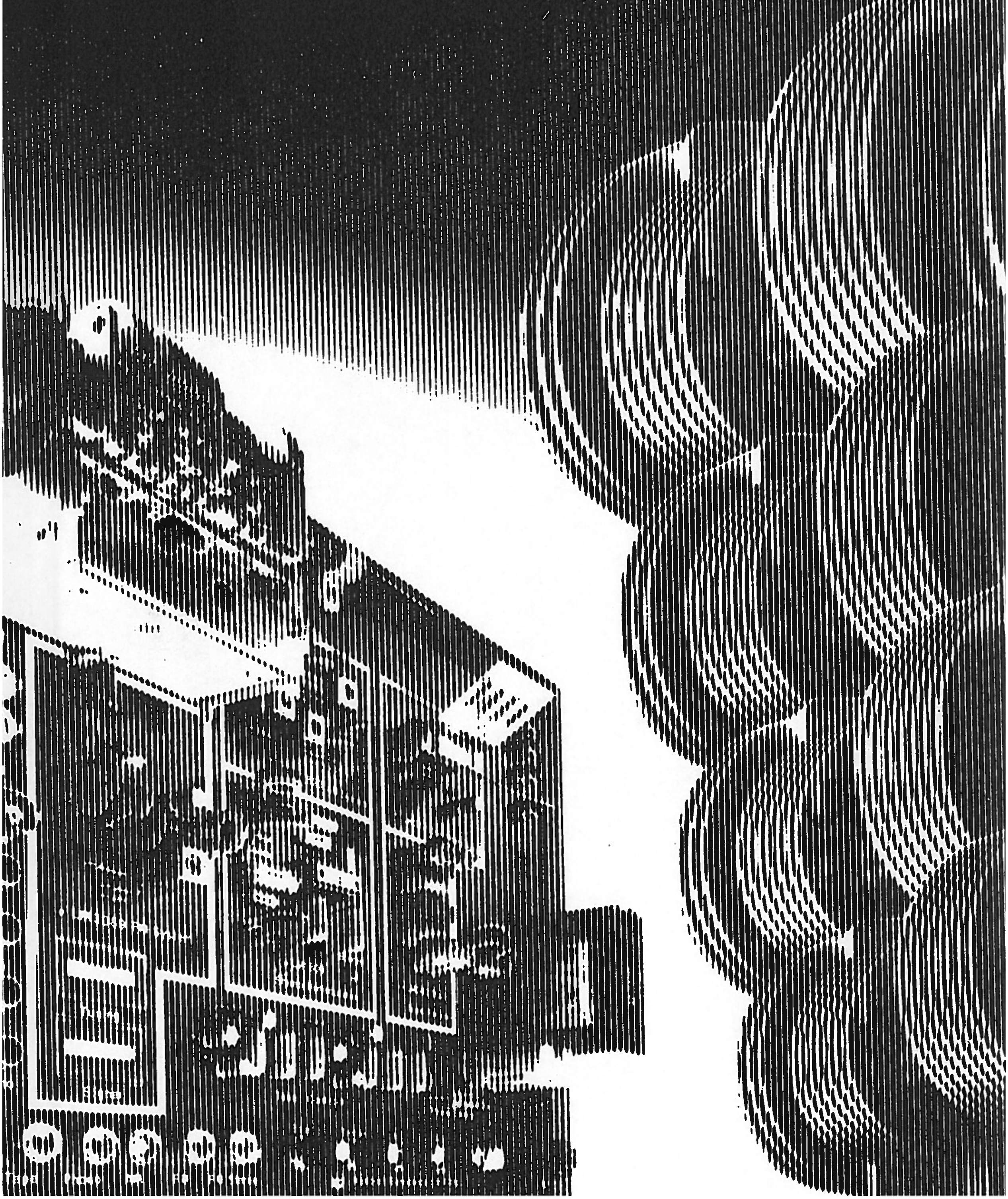


FIGURE 2.20.1 Deglitched Voltage Mode Switch

Section 3.0
AM, FM and FM Stereo



3.0 AM, FM and FM Stereo

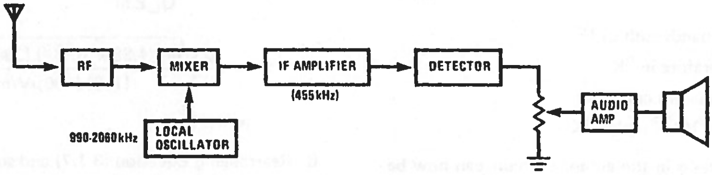


FIGURE 3.1.1 Superheterodyne Radio

3.1 AM RADIO

3.1.1 Introduction

Almost exclusively, the superheterodyne circuit reigns supreme in the design of AM broadcast radio. This circuit, shown in Figure 3.1.1, converts the incoming signal – 535kHz to 1605kHz – to an intermediate frequency, usually 262.5kHz or 455kHz, which is further amplified and detected to produce an audio signal which is further amplified to drive a speaker. Other types of receiver circuits include tuned RF (TRF) and regenerative.

In the tuned RF, the incoming signal is amplified to a relatively high level by a tuned circuit amplifier, and then demodulated.

Controlled positive feedback is used in the regenerative receiver to increase circuit Q and gain with relatively few components to obtain a satisfactory measure of performance at low cost.

Both the TRF and regenerative circuits have been used for AM broadcast, but are generally restricted to low cost toy applications.

3.1.2 Conversion of Antenna Field Strength to Circuit Input Voltage

Looking at Figure 3.1.1, the antenna converts incoming radio signals to electrical energy. Most pocket and table radios use ferrite loop antennas, while automobile radios are designed to work with capacitive whip antennas.

Ferrite Loop Antennas

The equivalent circuit of a ferrite rod antenna appears as Figure 3.1.2. Terms and definitions follow:

L = antenna inductance
C = tuning capacitor plus stray capacitance (20-150pF typ.)
N_o = antenna turns ratio – primary to secondary
R_{IN} = circuit input impedance
R_p = equivalent parallel loss resistance (primarily a function of core material)
R_L = equivalent loading resistance
V_{IN} = volts applied to circuit
V_{ID} = volts induced to antenna
V_T = voltage transferred across tank
Q_u = unloaded Q of antenna coil
Q_L = loaded Q of antenna circuit
H_{eff} = effective height of antenna in meters
E = field strength in volts/meter

Necessary design equations appear below:

$$Q_u = \frac{R_p}{X_L} \quad (3.1.1)$$

$$Q_L = \frac{R_p \| R_L}{X_L} = \frac{R_T}{X_L} \quad (3.1.2)$$

$$R_L = N_o^2 R_{IN} \quad (3.1.3)$$

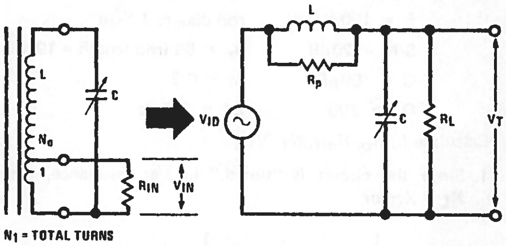


FIGURE 3.1.2 Ferrite Rod Antenna Equivalent Circuit

$$V_T = Q_L V_{ID} \quad (3.1.4)$$

$$V_{ID} = H_{eff} E \quad (3.1.5)$$

$$V_{IN} = \frac{V_T}{N_o} \quad (3.1.6)$$

The effective height of the antenna is a complex function of core and coil geometry, but can be approximated[1] by:

$$H_{eff} \approx \frac{2\pi \mu_r N_1 A}{\lambda} \quad (3.1.7)$$

where: N_1 = total number of turns
μ_r = relative permeability of antenna rod (primarily function of length)
A = cross sectional area of rod
λ = wavelength of received signal
$= \frac{3 \times 10^8 \text{m/sec}}{\text{freq (Hz)}}$

Noise voltage is calculated from the total Thevenin equivalent loading resistance, $R_T = R_p \| R_L$, using Equation (3.1.8):

$$e_n = \sqrt{4 K T \Delta f R_T} \qquad (3.1.8)$$

where: Δf = 3dB bandwidth of IF

T = temperature in °K

K = Boltzmann's constant

$= 1.38 \times 10^{-23}$ joules/°K

The signal-to-noise ratio in the antenna circuit can now be expressed as Equation (3.1.9):

$$S/N = \frac{V_T m}{e_n} = \frac{Q_L H_{eff} E m}{\sqrt{4 K T \Delta f R_T}} \qquad (3.1.9)$$

where: m = index of modulation

Example 3.1.1

Specify the turns ratio N_o, total turns N_1, effective height H_{eff}, and inductance required for an antenna wound onto a rod with the characteristics shown, designed to match an input impedance of 1kΩ. Calculate the circuit input voltage resulting from a field strength of 100µV/m with 20dB S/N in the antenna circuit. Assume a 15-365pF tuning capacitor set at 100pF for an input frequency of 1MHz.

Given:	R_{IN} = 1kΩ	f_o = 1MHz
	E = 100µV/m	rod dia. = 1.5cm
	S/N = 20dB	μ_r = 65 (rod length = 19cm)
	C = 100pF	m = 0.3
	Q_u = 200	Δf = 10kHz

Calculate L, N_o, H_{eff}, N_1, V_{IN}

1. Since the circuit is "tuned," i.e., at resonance, then $X_L = X_C$, or

$$L = \frac{1}{C(2\pi f_o)^2} = \frac{1}{100\text{pF}\,(2\pi \times 1 \times 10^6)^2}$$

$$= 2.53 \times 10^{-4}\ \text{H}$$

$$L \approx 250\mu\text{H}$$

2. From Equation (3.1.1):

$$R_p = Q_u X_L = 200 \times 2\pi \times 1\text{MHz} \times 250\mu\text{H}$$

$$R_p \approx 314\text{k}$$

3. For matched conditions and using Equation (3.1.3):

$$R_p = R_L = N_o^2 R_{IN}$$

$$N_o = \sqrt{\frac{R_p}{R_{IN}}} = \sqrt{\frac{314\text{k}}{1\text{k}}} = 17.7$$

$$N_o \approx 18{:}1$$

4. From Equations (3.1.1) and (3.1.2):

$$Q_L = \frac{R_p \| R_L}{X_L} = \frac{R_p}{2X_L} = \frac{Q_u}{2} \quad \text{since } R_p = R_L$$

$$Q_L = 100$$

5. Rearranging Equation (3.1.9) and solving for required H_{eff}:

$$H_{eff} = \frac{S/N \sqrt{4 K T \Delta f R_T}}{Q_L E m}$$

$$= \frac{10\sqrt{(4)(1.38 \times 10^{-23})(300)(10\text{kHz})(157\text{k})}}{(100)(100\mu\text{V/m})(0.3)}$$

$$= 1.7\text{cm}$$

6. Rearranging Equation (3.1.7) and solving for N_1:

$$N_1 = \frac{H_{eff} \lambda}{2\pi \mu_r A}$$

$$= \frac{(0.017\text{m})(3 \times 10^8\text{m/sec})}{(2\pi)(65)(1 \times 10^6\text{Hz})(\pi)(7.5 \times 10^{-3}\text{m})^2} = 70.7$$

$$N_1 \approx 71 \text{ turns}$$

7. From Equation (3.1.5):

$$V_{ID} = H_{eff} E$$

$$= 0.017\text{m} \times 100\mu\text{V/m}$$

$$V_{ID} = 1.7\mu\text{V}$$

8. Find V_T from Equation (3.1.4):

$$V_T = Q_L V_{ID}$$

$$= 100 \times 1.7\mu\text{V}$$

$$V_T = 170\mu\text{V}$$

9. Using Equation (3.1.6), find V_{IN}:

$$V_{IN} = \frac{V_T}{N_o} = \frac{170\mu\text{V}}{18}$$

$$V_{IN} = 9.4\mu\text{V}$$

Capacitive Automotive Antennas

A capacitive automobile radio antenna can be analyzed in a manner similar to the loop antenna. Figure 3.1.3 shows the equivalent circuit of such an antenna. C_1 is the capacitance of the vertical rod with respect to the horizontal ground plane, while C_2 is the capacitance of the shielded cable connecting the antenna to the radio. In order to obtain a useful signal output, this capacitance is tuned out with an inductor, L. Losses in the inductor and the input resistance of the radio form R_L. The signal appearing at the input stage of the radio is related to field strength:

$$V_T = V_{ID} Q_L \frac{C_1}{C_T} \qquad (3.1.10)$$

where: V_{ID} is defined by Equation (3.1.5)

Q_L is defined by Equation (3.1.2)

$C_T = C_1 + C_2$

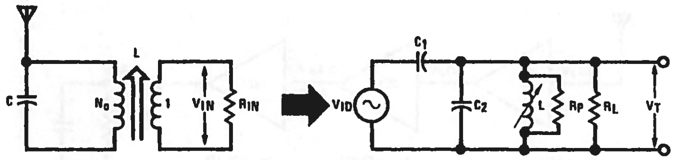


FIGURE 3.1.3 Capacitive Auto Antenna Equivalent Circuit

Similar to the ferrite rod antenna, the signal-to-noise ratio is given by:

$$S/N = \frac{H_{eff}\, E\, m\, Q_L\, (C_1/C_T)}{\sqrt{4\,K\,T\,\Delta f\, Q_L\, X_{CT}}} \quad (3.1.11)$$

The effective height of a capacitive vertical whip antenna can be shown[1] to equal Equation (3.1.12):

$$H_{eff} \approx \frac{h}{2} \quad (3.1.12)$$

where: h = antenna height in meters

Example 3.1.2

For comparison purposes, calculate the circuit input voltage, V_{IN}, for an automotive antenna operating in the same field as the previous example; assume same circuit input impedance of 1kΩ and calculate the resultant S/N. Use the given data for a typical auto radio antenna extended two sections (1 meter).

Given:	$R_{IN} = 1k\Omega$	$\Delta f = 10kHz$
	$E = 100\mu V/m$	$C_1 = 10pF$
	$Q_L = 80$	$C_T = 90pF$
	$f_o = 1MHz$	$m = 0.3$

Calculate S/N, N_o, V_{IN}.

1. Calculate H_{eff} from Equation (3.1.12) and solve for X_{CT}

$$H_{eff} = \frac{h}{2} = 0.5m$$

$$X_{CT} = \frac{1}{2\pi f C_T} = \frac{1}{2\pi \times 1MHz \times 90pF}$$

$$X_{CT} = 1768\Omega$$

2. Rearranging Equation (3.1.11) and solving for S/N:

$$S/N = \frac{H_{eff}\, E\, m \frac{C_1}{C_T} \sqrt{Q_L}}{\sqrt{4\,K\,T\,\Delta F\, X_{CT}}}$$

$$S/N = \frac{(0.5)\,(100\mu V/m)\,(0.3)\,\frac{10pF}{90pF}\,\sqrt{80}}{\sqrt{(4)\,(1.38 \times 10^{-23})\,(300)\,(10k)\,(1768)}}$$

$$S/N = 27.55$$

$$S/N \approx 29dB$$

3. From Equations (3.1.10) and (3.1.5):

$$V_T = H_{eff}\, E\, Q_L \frac{C_1}{C_T}$$

$$= 0.5m \times 100\mu V/m \times 80 \times \frac{10pF}{90pF}$$

$$V_T = 444\mu V$$

4. Since matching requires $R_p = R_L$, and resonance gives $X_{CT} = X_L$, then using Equation (3.1.2):

$$\frac{R_p}{2} = Q_L X_{CT}$$

$$R_p = 2 \times 80 \times \frac{1}{2\pi\,(1MHz)\,(90pF)} = 283k = R_L$$

5. Using Equation (3.1.3):

$$N_o = \sqrt{\frac{R_L}{R_{IN}}} = \sqrt{\frac{283k}{1k}} = 16.8$$

$$N_o \approx 17{:}1$$

6. From Equation (3.1.6):

$$V_{IN} = \frac{V_T}{N_o} = \frac{444\mu V}{17}$$

$$V_{IN} = 26.1\mu V$$

7. From Equation (3.1.1):

$$Q_u = \frac{R_p}{X_{CT}} = 283k \times 2\pi \times 1MHz \times 90pF = 160$$

It is interesting to note that operating in the same field strength, the capacitive antenna will transfer approximately three times as much voltage to the input of the circuit, thus allowing the greater signal-to-noise ratio of 29dB.

REFERENCES

1. Laurent, H. J. and Carvalho, C. A. B., "Ferrite Antennas for AM Broadcast Receivers," Application Note available from Bendix Radio Division of The Bendix Corporation, Baltimore, Maryland.

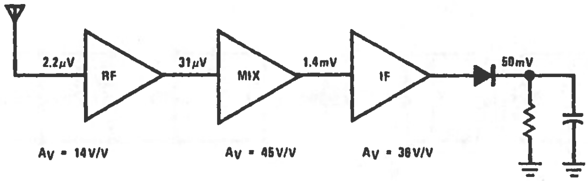


FIGURE 3.1.4 AM Radio Gain Stages

3.1.3 Typical AM Radio Gain Stages

The typical levels of Figure 3.1.4 give some idea of the gain needed in an AM radio. At the IF amplifier output, a diode detector recovers the modulation, and is generally designed to produce approximately $50mV_{RMS}$ of audio with m = 0.3. The gain required is therefore:

$$A_V = \frac{50mV}{2.2\mu V} = 23kV/V \text{ or } 87dB$$

3.2 LM3820 AM RECEIVER SYSTEM

The LM3820 is a 3 stage AM radio IC designed as an improved replacement for the LM1820. It consists of the following functional blocks:

RF Amplifier	IF Amplifier
Oscillator	AGC Detector
Mixer	Regulator

The RF amplifier section (Figure 3.2.1) consists of a cascode amplifier Q_2 and Q_3, whose geometries are specially designed for low noise operation from low source impedances. Q_2 is protected from overloads coupled via capacitive antennae by two back to back diodes. The cascode configuration has very low feedback capacitance to minimize stability problems, and a high output impedance to maximize gain. In addition, bias components (Q_1, etc.) are included. Biased at 5.6mA, the input stage is useful for frequencies in excess of 50mHz. Figure 3.2.2a shows the transconductance as a function of frequency.

Transistors Q_4 and Q_5 make up the local oscillator circuit. Positive feedback from the collector of Q_5 to the base of Q_4 is provided by the resistor divider R_9 and R_8. The oscillator frequency is set with a tuned circuit connected between pin 2 and V_{CC}. Transistors Q_4 and Q_5 are biased at 0.5mA each, so the transconductance of the differential pair is 10mmhos. For oscillation, the impedance at pin 2 must be high enough to provide a voltage gain greater than the loss associated with the resistor divider network R_9, R_8 and the input impedance of Q_4. Values of load impedance greater than 400Ω satisfy this condition, with values of 10kΩ or greater being commonly used.

The differential pair Q_6 and Q_7 serve as a mixer, being driven with current from the oscillator. The input signal, applied to pin 1, is multiplied by the local oscillator frequency to produce a difference frequency at pin 14. This signal, the IF, is filtered and stepped down to match the input impedance of the IF amplifier.

Transistors Q_9 and Q_{10} form the IF amplifier gain stage. Again, a cascode arrangement is used for stability and high gain for a gm of 90mmhos.

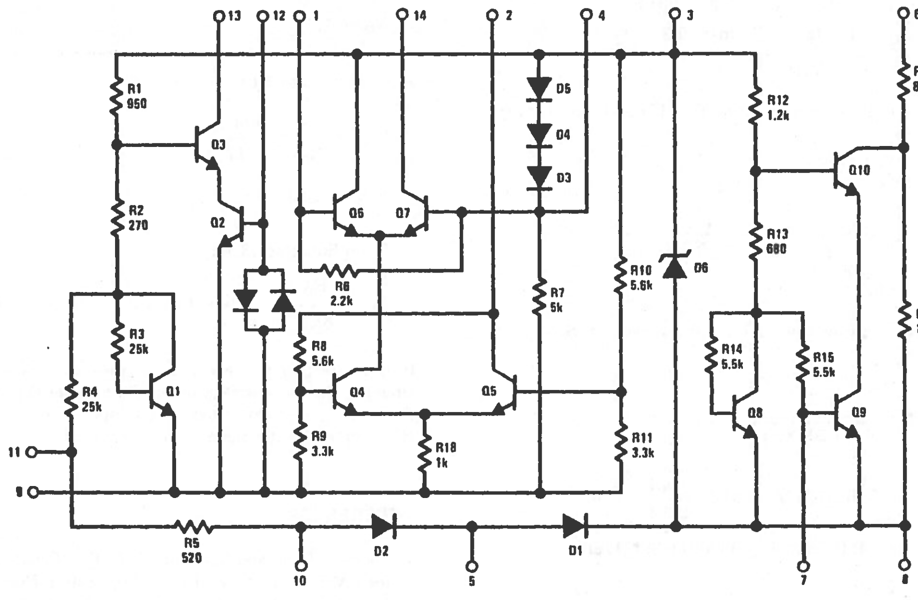
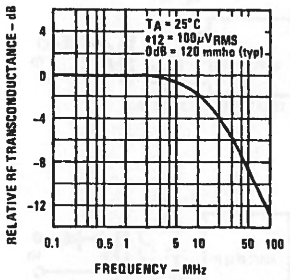
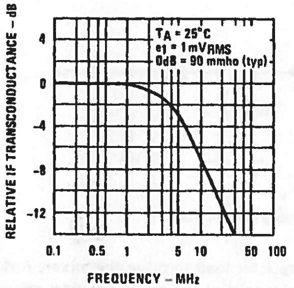


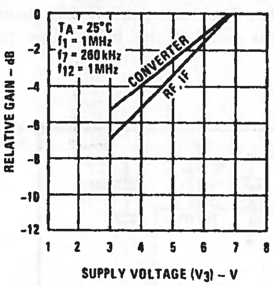
FIGURE 3.2.1 LM3820 Schematic Diagram



(a) RF Transconductance as a Function of Frequency



(b) IF Transconductance as a Function of Frequency



(c) Relative Gain as a Function of Supply Voltage (V_3)

FIGURE 3.2.2 LM3820 Performance Characteristics

An AGC detector is included on the chip. The circuit consists of diodes D_1 and D_2 which function as a peak to peak detector driven with IF signal from the output of the IF amplifier. As the output signal increases, a greater negative voltage is developed on pin 10 which diverts current away from the input transistor Q_2. This current reduction in turn reduces the gain of the input stage, effectively regulating the signal at the IF output.

A zener diode is included on the chip and is connected from V_{CC} to ground to provide regulation of the bias currents on the chip. However, the 3820 functions well at voltages below the zener regulating voltage as shown in Figure 3.2.2c. Table 3.2.1 summarizes circuit parameters.

Basically, two ways exist for using the LM1820 in AM radio applications; these are illustrated in Figure 3.2.3. The mixer-IF-IF configuration (Figure 3.2.3a) results in an economical approach at some performance sacrifice because the mixer contributes excess noise at the antenna input, which reduces sensitivity. Since all gain is taken at the IF frequency, stability problems may be encountered if attention is not paid to layout.

TABLE 3.2.1 Summary of Circuit Parameters

Parameter	RF Section	Mixer	IF
Input Resistance	1k	1.4k	1k
Input Capacitance	80pF	8pF	70pF
Transconductance	120mmhos	2.5mmhos	90mmhos
Input Noise Voltage, 6kHz Bandwidth	0.2μV	0.5μV	

The RF-mixer-IF approach (Figure 3.2.3b) takes advantage of the low noise input stage to provide a high performance receiver for either automobile or high quality portable or table radio applications. Another approach which sacrifices little in performance, yet reduces cost associated with the three gang tuning capacitor, is to substitute a resistor for the tuned circuit load of the RF amplifier. The LM3820 has sufficient gain to allow for the mismatch and still provide good performance.

By appropriate impedance matching between stages, gain in excess of 120dB is possible. This can be seen from Figure 3.2.3c, where the correct interstage matching values for maximum power gain are shown. The gain of the RF section is found from:

$$A_{V1} = \frac{V_1}{V_{IN}} = K_1 \, g_{m1} \, R_{L1} \, N \, K_2$$

where: N = turns ratio = $\sqrt{R_{sec}/R_{pri}}$

K_1 = 6dB loss @ output of RF amplifier due to matching 500k output impedance

K_2 = 6dB loss @ input to mixer due to matching 1.4k input impedance

For the values shown:

$$A_{V1} = \frac{1}{2}\,(120 \times 10^{-3})\,(500k)\,\sqrt{\frac{1.4k}{500k}}\;\frac{1}{2}$$

$$= 793.5 \approx 58\text{dB}$$

Similarly, for the mixer:

$$A_{V2} = \frac{1}{2}\,(2.5 \times 10^{-3})\,(500k)\,\sqrt{\frac{1k}{500k}}\;\frac{1}{2}$$

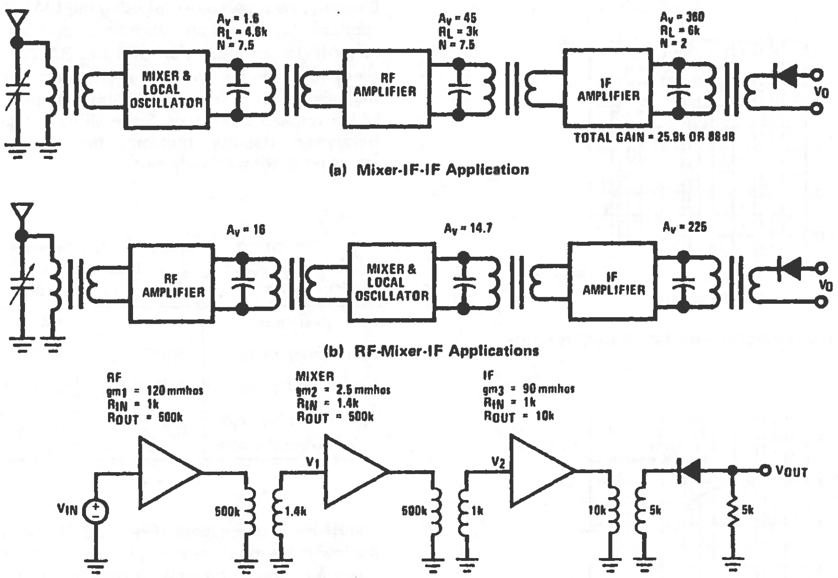
$$= 14 \approx 23\text{dB}$$

And for the IF:

$$A_{V3} = \frac{1}{2}\,(90 \times 10^{-3})\,(10k)\,\sqrt{\frac{5k}{10k}}\;\frac{1}{2}$$

$$= 159 \approx 44\text{dB}$$

Total gain $= 1.8 \times 16^6 \approx 125$dB



(a) Mixer-IF-IF Application

(b) RF-Mixer-IF Applications

(c) Power Matching for Maximum Gain

FIGURE 3.2.3 Circuit Configurations for AM Radios Using the LM3820

This much gain is undesirable from a performance standpoint, since it would result in 1.5V of noise to the diode detector due to the input noise, and it would probably be impossible to stabilize the circuit and prevent oscillation. From a design standpoint, it is desirable to mismatch the RF stage and mixer for less gain.

A capacitor tuned AM radio using the RF-mixer-IF configuration is shown in Figure 3.2.4. The RF amplifier is used with a resistor load to drive the mixer. A double tuned circuit at the output of the mixer provides selectivity, while the remainder of the gain is provided by the IF section, which is matched to the diode through a unity turns ratio transformer. A resistor from the detector to pin 10 bypasses the internal AGC detector in order to increase the recovered audio. The total gain in this design is 57k or 95dB from the base of the input stage to the diode detector.

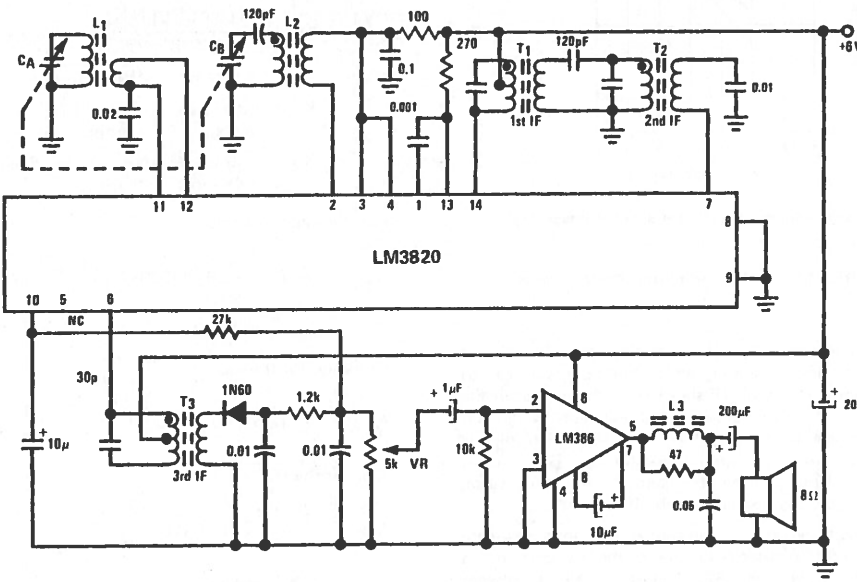


FIGURE 3.2.4 AM Radio Using RF-Mixer-IF

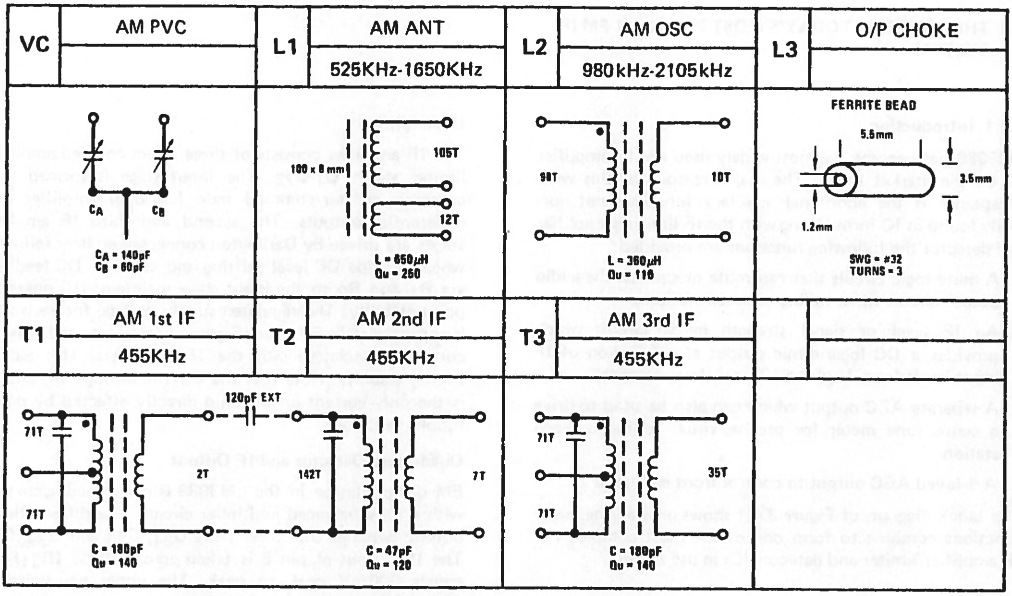


FIGURE 3.2.4 AM Radio Using RF-Mixer-IF continued

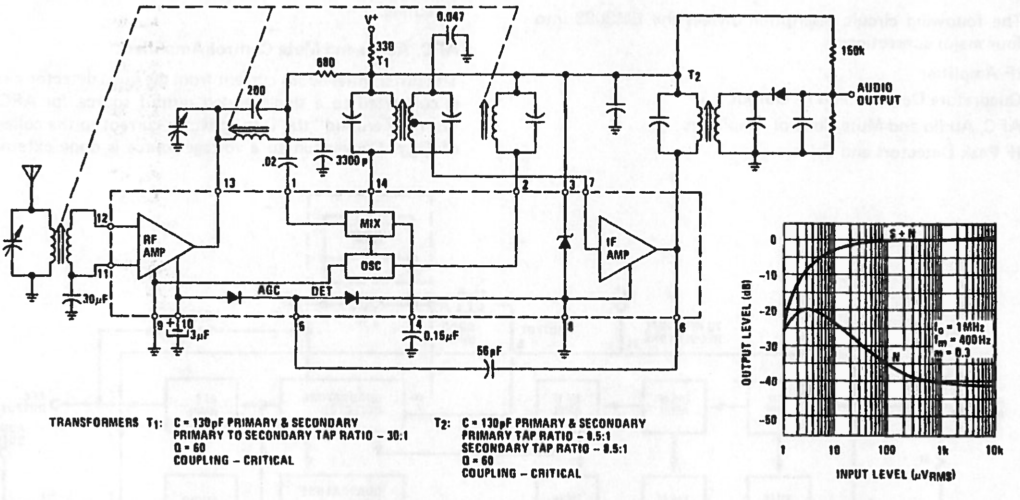


FIGURE 3.2.5 AM Auto Radio

A slug-tuned AM automobile radio design is shown in Figure 3.2.5. Tuning of both the input and the output of the RF amplifier and the mixer is accomplished with variable inductors. Better selectivity is obtained through the use of double tuned interstage transformers. Input circuits are inductively tuned to prevent microphonics and provide a linear tuning motion to facilitate push-button operation.

3.3 FM IF AMPLIFIERS AND DETECTORS

In the consumer field, two areas of application exist for FM IF amplifiers and detectors; in addition, applications exist in commercial two way and marine VHF FM radios:

TABLE 3.3.1 Application for FM-IF Amplifiers

Service	Frequency	Deviation	Input Limiting	Distortion
FM Broadcast	10.7MHz	75kHz	20µV	0.5%
TV Sound	4.5MHz	25kHz	200µV	1.5%
Two-Way Radio	various	5kHz	5µV	5%

The major requirement of an FM IF is good limiting characteristics, i.e., the ability to produce a constant output level to drive a detector regardless of the input signal level. This quality removes noise and amplitude changes that would otherwise be heard in the recovered signal.

3.4 THE LM3089 – TODAY'S MOST POPULAR FM IF SYSTEM

3.4.1 Introduction

LM3089 has become the most widely used FM IF amplifier IC on the market today. The major reason for this wide acceptance is the additional auxiliary functions not normally found in IC form. Along with the IF limiting amplifier and detector the following functions are provided:

1. A mute logic circuit that can mute or squelch the audio output circuit when tuning between stations.
2. An IF level or signal strength meter circuit which provides a DC logarithmic output as a function of IF input levels from 10μV to 100mV (four decades).
3. A separate AFC output which can also be used to drive a center-tune meter for precise visual tuning of each station.
4. A delayed AGC output to control front end gain.

The block diagram of Figure 3.4.1 shows how all the major functions combine to form one of the most complex FM IF amplifier/limiter and detector ICs in use today.

3.4.2 Circuit Description (Figure 3.4.2)

The following circuit description divides the LM3089 into four major subsections:

IF Amplifier
Quadrature Detector and IF Output
AFC, Audio and Mute Control Amplifiers
IF Peak Detectors and Drivers

IF Amplifier

The IF amplifier consists of three direct coupled amplifier-limiter stages Q_1-Q_{22}. The input stage is formed by a common emitter/common base (cascode) amplifier with differential outputs. The second and third IF amplifier stages are driven by Darlington connected emitter followers which provide DC level shifting and isolation. DC feedback via R_1 and R_2 to the input stage maintains DC operating point stability. The regulated supply voltage for each stage is approximately 5V. The IF ground (pin 4) is used only for currents associated with the IF amplifiers. This aids in overall stability. Note that the current through R_9 and Z_1 is the only current on the chip directly affected by power supply variations.

Quadrature Detector and IF Output

FM demodulation in the LM3089 is performed accurately with a fully balanced multiplier circuit. The differential IF output switches the lower pairs Q_{34}, Q_{26} and Q_{39}, Q_{38}. The IF output at pin 8 is taken across 390Ω (R_{31}) and equals 300mV peak to peak. The upper pair-switching (Q_{35}, Q_{23}) leading by 90 degrees is through the externally connected quad coil at pin 9. The 5.6V reference at pin 10 provides the DC bias for the quad detector upper pair switching.

AFC, Audio and Mute Control Amplifiers

The differential audio current from the quad detector circuit is converted to a single ended output source for AFC by "turning around" the Q_{47} collector current to the collector of Q_{57}. Conversion to a voltage source is done externally

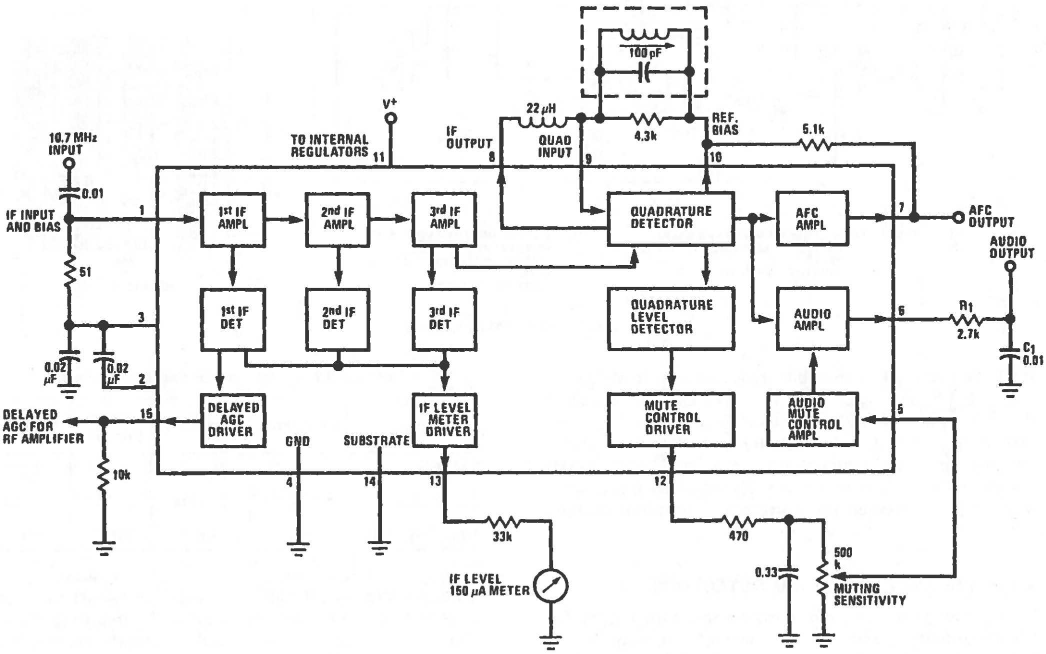


FIGURE 3.4.1 LM3089 Block and Connection Diagram

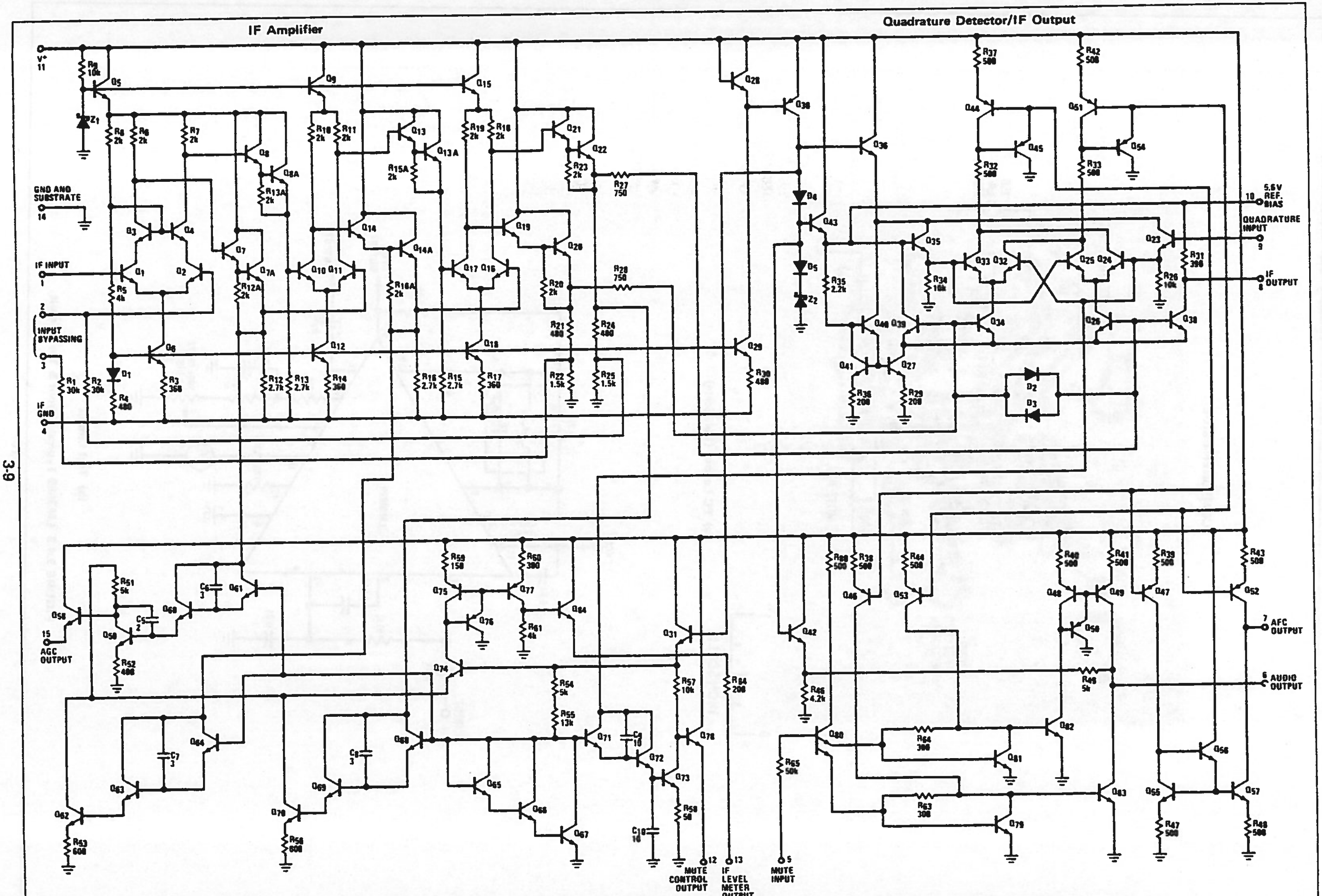
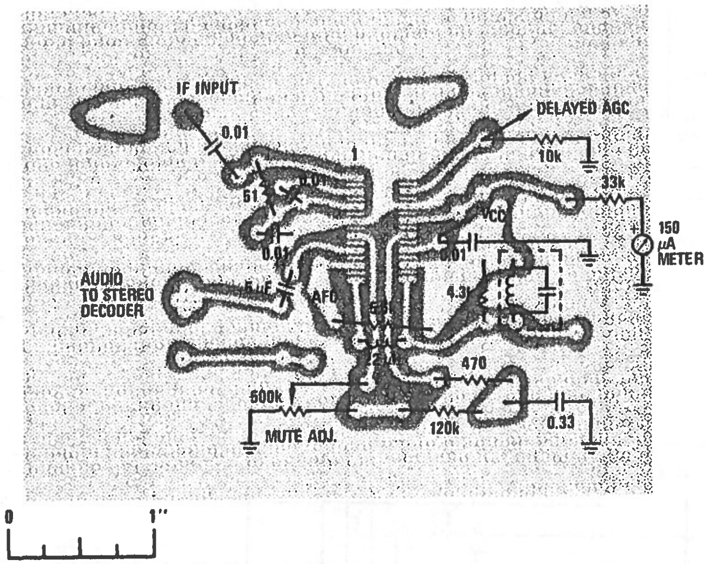
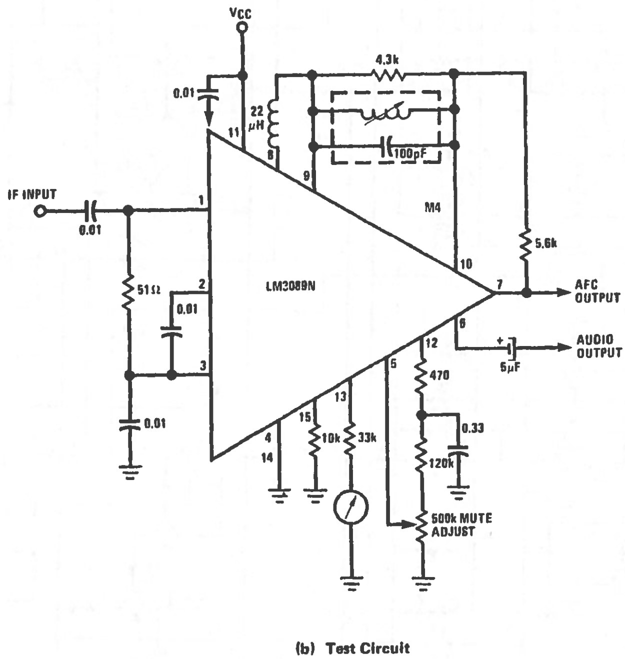


FIGURE 3.4.2 LM3089 Schematic Diagram

COMPONENT SIDE



(a) PC Layout (Full Scale)



(b) Test Circuit

FIGURE 3.4.3 LM3089 Typical Layout & Test Circuit by adding a resistor from pin 7 to pin 10. The audio amplifier stage operates in a similar manner as the AFC amplifier except that two "turn around" stages are used. This configuration allows the inclusion of muting transistor Q_{80}. A current into the base of Q_{80} will cause transistors Q_{79} and Q_{81} to saturate, which turns off the audio amplifier; the gain of the audio stage is set by internal resistor R_{49}. This 5kΩ resistor value is also the output impedance of the audio amplifier. When the LM3089 is used in mono receivers the 75μs de-emphasis (RC time constant) is calculated for a 0.01μF by including R_{49}. ($RC = [R_{49} + R_1][C_1]$, $R_1 = 75\mu s/0.01\mu F - 5k\Omega \approx 2.7k$, Figure 3.7.1.)

IF Peak Detectors and Drivers

Four IF peak or level detectors provide the delayed AGC, IF level and mute control functions. An output from the first IF amplifier drives the delayed AGC peak detector. Since the first IF amplifier is the last IF stage to go into limiting, Q_{60} and Q_{61} convert the first IF output voltage swing to a DC current (for IF input voltages between 10mV and 100mV). This changing current (0.1 to 1mA) is converted to a voltage across R_{51}. Emitter follower Q_{58} buffers this output voltage for pin 15. The top of resistor R_{51} is connected to a common base amplifier Q_{74} along with the output currents from the 2nd and 3rd stage IF peak detectors (which operate for IF input voltages between 10μV and 10mV). The output current from Q_{75} is turned around or mirrored by Q_{75}, Q_{76}, and Q_{77}, cut in half, then converted to a voltage across R_{61}. Emitter follower Q_{84} buffers this voltage for pin 13.

The fourth peak detector "looks" at the IF voltage developed across the quad coil. For levels above about 120mV at pin 9, Q_{73} will saturate and provide no output voltage at pin 12. Because the IF level at pin 9 is constant, as long as the last IF amplifier is in limiting, pin 12 will remain low. Sudden interruptions or loss of the pin 9 IF signal due to noise or detuning of the quad coil will allow the collector of Q_{73} to rise quite rapidly. The voltage at the collector of Q_{73} is buffered by Q_{78} for pin 12.

3.4.3 Stability Considerations

Because the LM3089 has wide bandwidth and high gain (> 80dB at 10.7MHz), external component placement and PC layout are critical. The major consideration is the effect of output to input coupling. The highest IF output signal will be at pins 8 and 9; therefore, the quad coil components should not be placed near the IF input pin 1. By keeping the input impedance low (< 500Ω) the chances of output to input coupling are reduced. Another and perhaps the most insidious form of feedback is via the ground pin connections. As stated earlier the LM3089 has two ground pins; the pin 4 ground should be used only for the IF input decoupling. The pin 4 ground is usually connected to the pin 14 ground by a trace under the IC. Decoupling of V_{CC} (pin 11), AGC driver (pin 15), meter driver (pin 13), mute control (pin 12) and in some cases the 5.6V REF (pin 10) should be done on the ground pin 14 side of the IC. The PC layout of Figure 3.4.3 has been used successfully for input impedances of 500Ω (1kΩ source/1kΩ load).

3.4.4 Selecting Quad Coil Components

The reader can best understand the selection process by example (see Figure 3.7.4):

Given: require quad coil bandwidth equal to 800kHz

$f_o = 10.7MHz$

Q_U (unloaded) $= 75$

Find: L_{CH} and R_{EXT}

Find loaded Q of quad coil for required BW (Q_L)

$$Q_L = \frac{f_o}{BW} = \frac{10.7MHz}{0.8MHz} = 13.38$$

Find total resistance across quad coil for required BW (R_T)

$$R_T = Q_L X_{LI} = 13.38(2\pi f_o L_I) = 1981\Omega$$

Find reactance of coupling choke (XL_{CH})

$$XL_{CH} = \frac{R_T V_8}{V_9} = \frac{1981 \times 0.110}{0.15} = 1453\Omega$$

Find inductance of coupling choke (L_{CH})

$$L_{CH} = \frac{XL_{CH}}{2\pi f} = \frac{1453\Omega}{6.72 \times 10^7} = 22\mu H$$

Find parallel resistance of the unloaded quad coil (R_p)

$$R_p = X_{LI} Q_{UL} = 148\Omega \times 75 = 11.1k\Omega$$

Convert R_{31}, L_{CH} series to parallel resistance (R_{L31})

$$R_{L31} = \frac{(XL_{CH})^2}{R_{31}} + R_{31} = 5803\Omega$$

Find R_{EXT} for $R_T = R_p \| R_{L31} \| R_{EXT}$

$$R_{EXT} = \frac{1}{\frac{1}{R_T} - \frac{1}{R_p} - \frac{1}{R_{L31}}} = 4126$$

Use $R_{EXT} = 4.3k$.

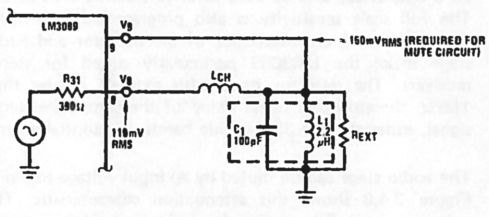


FIGURE 3.4.4 Quad Coil Equivalent Circuit

3.4.5 Typical Application of the LM3089

The circuit in Figure 3.4.5 illustrates the simplicity in designing an FM IF. The ceramic filters used in this application have become very popular in the last few years because of their small physical size and low cost. The filters eliminate all but one IF alignment step. The filters are terminated at the LM3089 input with 330Ω. Disc ceramic type capacitors with typical values of 0.01 to 0.02μF should be used for IF decoupling at pins 2 and 3.

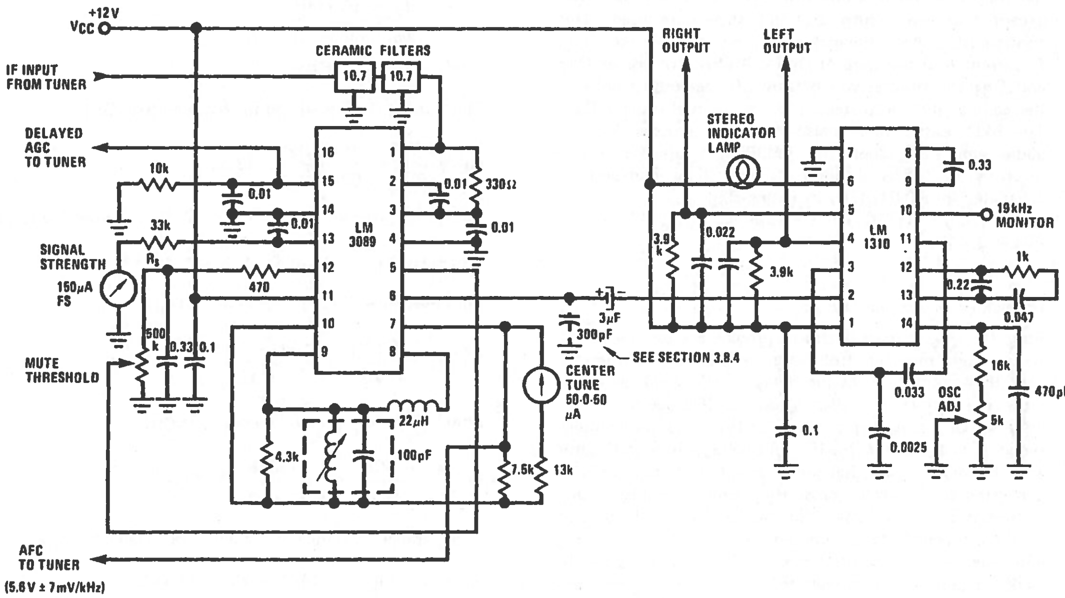


FIGURE 3.4.5 Typical Application of the LM3089

The AFC output at pin 7 can serve a dual purpose. In Figure 3.4.6 AFC sensitivity, expressed as mV/kHz, is programmed externally with a resistor from pin 7 to pin 10. A voltage reference other than pin 10 may be used as long as the pin 7 voltage stays less than 2V from the supply and greater than 2V from ground. The voltage change for a 5kΩ resistor will be ≈ 7.5mV/kHz or ≈ 1.5μA/kHz. The AFC output can also be used to drive a center tune meter. The full scale sensitivity is also programmed externally. The wide band characteristics of the detector and audio stage make the LM3089 particularly suited for stereo receivers. The detector bandwidth extends greater than 1MHz, therefore the phase delay of the composite stereo signal, especially the 38kHz side bands, is essentially zero.

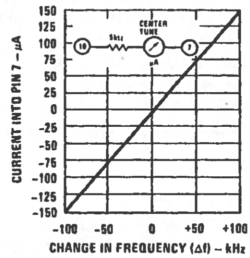


FIGURE 3.4.6 AFC (Pin 7) Characteristics vs. IF Input Frequency Change

The audio stage can be muted by an input voltage to pin 5. Figure 3.4.8 shows this attenuation characteristic. The voltage for pin 5 is derived from the mute logic detector pin 12. Figure 3.4.7 shows how the pin 12 voltage rises when the IF input is below 100μV. The 470Ω resistor and 0.33μF capacitor filter out noise spikes and allow a smooth mute transition. The pot is used to set or disable the mute threshold. When the pot is set for maximum mute sensitivity some competitors' versions of the LM3089 would cause a latch-up condition, which results in pin 12 staying high for all IF input levels. National's LM3089 has been designed such that this latch-up condition cannot occur.

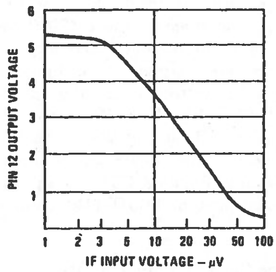


FIGURE 3.4.7 Mute Control Output (Pin 12) vs. IF Input Signal

The signal strength meter is driven by a voltage source at pin 13 (Figure 3.4.9). The value of the series resistor is determined by the meter used:

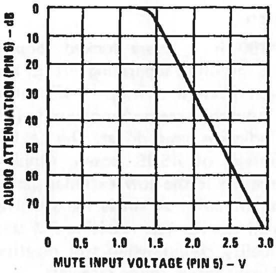


FIGURE 3.4.8 Typical Audio Attenuation (Pin 6) vs. Mute Input Voltage (Pin 5)

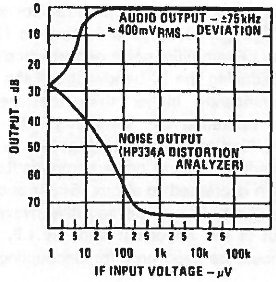


FIGURE 3.4.10 Typical (S+N)/N and IF Limiting Sensitivity vs. IF Input Signal

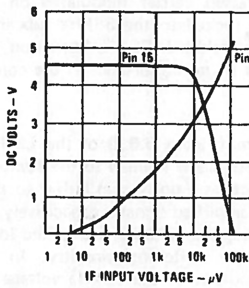


FIGURE 3.4.9 Typical AGC (Pin 15) and Meter Output (Pin 13) vs. IF Input Signal

$$R_S = \frac{V_{MAX(13)}}{I_{FS}} = \frac{5V}{150\mu A} = 33k$$

The maximum current from pin 13 should be limited to approximately 2mA. Short circuit protection has been included on the chip.

The delayed AGC (pin 15) is also a voltage source (Figure 3.4.9). The maximum current should also be limited to approximately 2mA.

Figure 3.4.10 shows the typical limiting sensitivity (measured at pin 1) of the LM3089 when configured per Figure 3.4.3b and using PC layout of Figure 3.4.3a.

3.5 THE LM3189

3.5.1 Introduction

The LM3189 offers all the features of the LM3089 with improvements in performance in some areas, and increased flexibility in others. Since the major functions of the LM3189 are similar to the LM3089, the following sections will detail only the changes that have been made.

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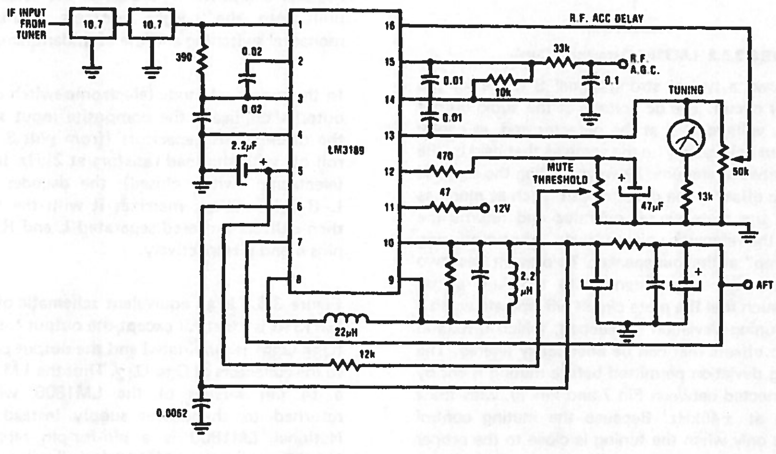

FIGURE 3.5.1 Typical Application of the LM3189

3.5.2 I.F. Amplifier

The input cascode stage has been optimized for low input capacitance and high gain for use with ceramic filters. An improvement in the I.F. amplifier noise performance has been accomplished by reducing the IC bandwidth. If the amplifier bandwidth is significantly higher than that needed to accommodate the operating I.F. frequency, out of band signals can be amplified and multiplied together in the non-linear stages to produce in-band noise components. The IC bandwidth has been decreased to about 15mHz and this will also help make the p.c.b. layout less sensitive. Nevertheless, attention to layout is still important and the I.F. amplifier ground (Pin 4) should be used only for decoupling the I.F. amplifier input.

3.5.3 R.f. a.g.c.

Instead of having a fixed r.f. a.g.c. delay threshold at the 10mV input signal level, the LM3189 allows the designer to select the a.g.c. threshold at any point between 200μV and 200mV, depending on the individual tuner requirements. A control voltage at the previously unused Pin 16 will determine the onset of r.f. a.g.c. action with a threshold level of 1.3V. This control voltage is obtained by a resistive divider connected to the signal strength meter Pin 13 as shown in Figure 3.5.1

3.5.4 Muting

Normally the muting circuit will operate by rectifying the signal that appears across the quad coil. Absence of a signal or noise "holes" in the carrier are peak detected and filtered to give the mute control voltage. The muting circuit of the LM3189 has been modified to include an early mute action when a strong signal with no noise "holes" is mistuned sufficiently. This is done to prevent dc shifts at the audio output from producing audible "thumps" in the loudspeaker.

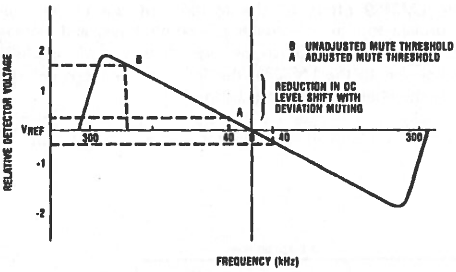

FIGURE 3.5.2 LM3189 Detector S Curve

Figure 3.5.2 shows a typical strong signal S curve for the LM3189 detector circuit. The dc voltage at the audio output will track the dc voltage level at the detector and, at center tuning, the output voltage will be the same as that held by the muting circuit between stations. However, when the signal is mistuned, the dc offset at the detector can reach as much as $\pm 2V_{DC}$ before the mute circuit operates and returns the audio output to the reference voltage level — thus producing an audible "thump" at the loudspeaker. To prevent this, two additional comparators are referenced to the AFC circuit control voltage such that the mute circuit will operate when a predetermined tuning deviation is reached, which results in much smaller dc offsets that can be adequately filtered. The degree of tuning deviation permitted before muting is set by the resistor connected between Pin 7 and Pin 10, with 15kΩ causing muting at ±40kHz. Because the muting control voltage changes only when the tuning is close to the proper point, Pin 12 can be used to indicate "on station" for automatic scanning tuning systems.

3.6 FM STEREO MULTIPLEX

3.6.1 Introduction

The LM1310/1800 is a phase locked loop FM stereo demodulator. In addition to separating left (L) and right (R) signal information from the detected IF output, this IC family features automatic stereo/monaural switching and a 100mA stereo indicator lamp driver. The LM1800 has the additional advantage of 45dB power supply rejection. Particularly attractive is the low external part count and total elimination of coils. A single inexpensive potentiometer performs all tuning. The resulting FM stereo system delivers high fidelity sound while still meeting the cost requirements of inexpensive stereo receivers.

Figures 3.6.1 and 3.6.2 outline the role played by the LM1310/1800 in the FM stereo receiver. The frequency domain plot shows that the composite input waveform contains L+R information in the audio band and L−R information suppressed carrier modulated on 38kHz. A 19kHz pilot tone, locked to the 38kHz subcarrier at the transmitter, is also included. SCA information occupies a higher band but is of no importance in the consumer FM receiver.

The block diagram (Figure 3.6.2) of the LM1800 shows the composite input signal applied to the audio frequency amplifier, which acts as a unity gain buffer to the decoder section. A second amplified signal is capacitively coupled to two phase detectors, one in the phase locked loop and the other in the stereo switching circuitry. In the phase locked loop, the output of the 76kHz voltage controlled oscillator (VCO) is frequency divided twice (to 38, then 19kHz), forming the other input to the loop phase detector. The output of the loop phase detector adjusts the VCO to precisely 76kHz. The 38kHz output of the first frequency divider becomes the regenerated subcarrier which demodulates L−R information in the decoder section. The amplified composite and an "in phase" 19kHz signal, generated in the phase locked loop, drive the "in phase" phase detector. When the loop is locked, the DC output voltage of this phase detector measures pilot amplitude. For pilot signals sufficiently strong to enable good stereo reception the trigger latches, applying regenerated subcarrier to the decoder and powering the stereo indicator lamp. Hysteresis, built into the trigger, protects against erratic stereo/monaural switching and the attendant lamp flicker.

In the monaural mode (electronic switch open) the decoder outputs duplicate the composite input signal except that the de-emphasis capacitors (from pins 3 and 6 to ground) roll off with the load resistors at 2kHz. In the stereo mode (electronic switch closed), the decoder demodulates the L−R information, matrixes it with the L+R information, then delivers buffered separated L and R signals to output pins 4 and 5 respectively.

Figure 3.6.3 is an equivalent schematic of an LM1800. The LM1310 is identical except the output turnaround circuitry (Q_{35}-Q_{38}) is eliminated and the output pins are connected to the collectors of Q_{39}-Q_{42}. Thus the LM1310 is essentially a 14 pin version of the LM1800, with load resistors returned to the power supply instead of ground. The National LM1800 is a pin-for-pin replacement for the UA758, while the LM1310 is a direct replacement for the MC1310.

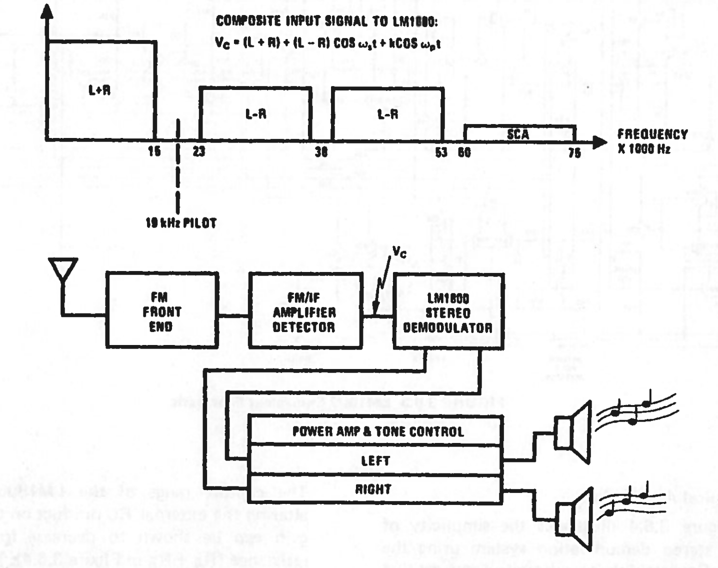

FIGURE 3.6.1 FM Receiver Block Diagram and Frequency Spectrum of LM1800 Input Signal

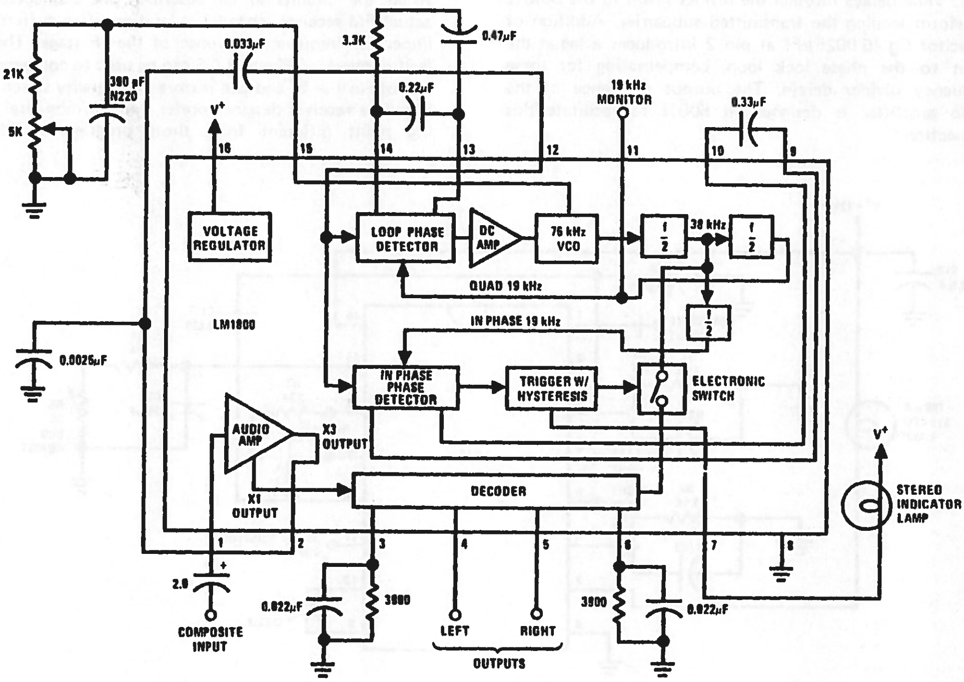

FIGURE 3.6.2 LM1800 Block Diagram

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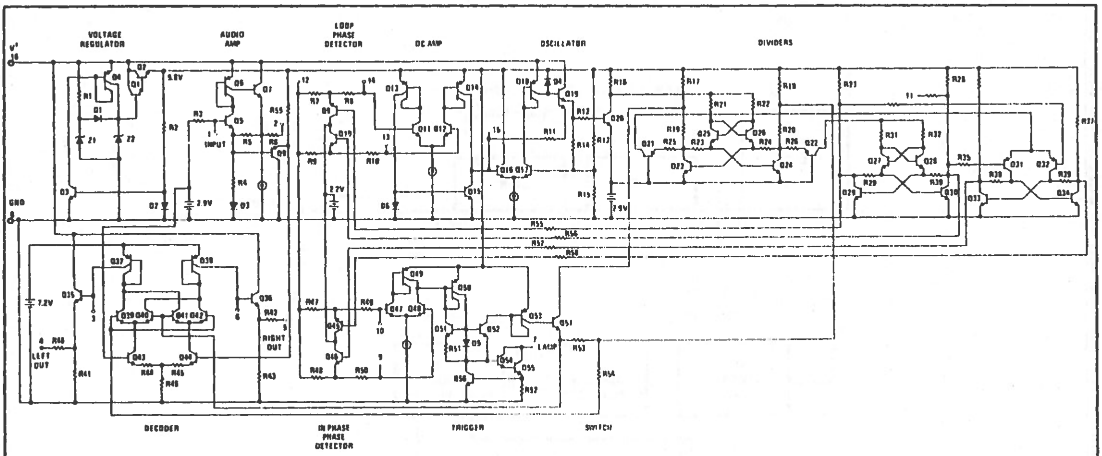

FIGURE 3.6.3 LM1800 Equivalent Schematic

3.6.2 LM1800 Typical Application

The circuit in Figure 3.6.4 illustrates the simplicity of designing an FM stereo demodulation system using the LM1800. R_3 and C_3 establish an adequate loop capture range and a low frequency well damped natural loop resonance. C_8 has the effect of shunting phase jitter, a dominant cause of high frequency channel separation problems. Recall that the 38kHz subcarrier regenerates by phase locking the output of a 19kHz divider to the pilot tone. Time delays through the divider result in the 38kHz waveform leading the transmitted subcarrier. Addition of capacitor C_9 (0.0025μF) at pin 2 introduces a lag at the input to the phase lock loop, compensating for these frequency divider delays. The output resistance of the audio amplifier is designed at 500Ω to facilitate this connection.

The capture range of the LM1800 can be changed by altering the external RC product on the VCO pin. The loop gain can be shown to decrease for a decrease in VCO resistance ($R_4 + R_5$ in Figure 3.6.4). Maintaining a constant RC product, while increasing the capacity from 390pF to 510pF, narrows the capture range by about 25%. Although the resulting system has slightly improved channel separation, it is more sensitive to VCO tuning.

When the circuits so far described are connected in an actual FM receiver, channel separation often suffers due to imperfect frequency response of the IF stage. The input lead network of Figure 3.6.5 can be used to compensate for roll off in the IF and will restore high quality stereo sound. Should a receiver designer prefer a stereo/monaural switching point different from those programmed into the

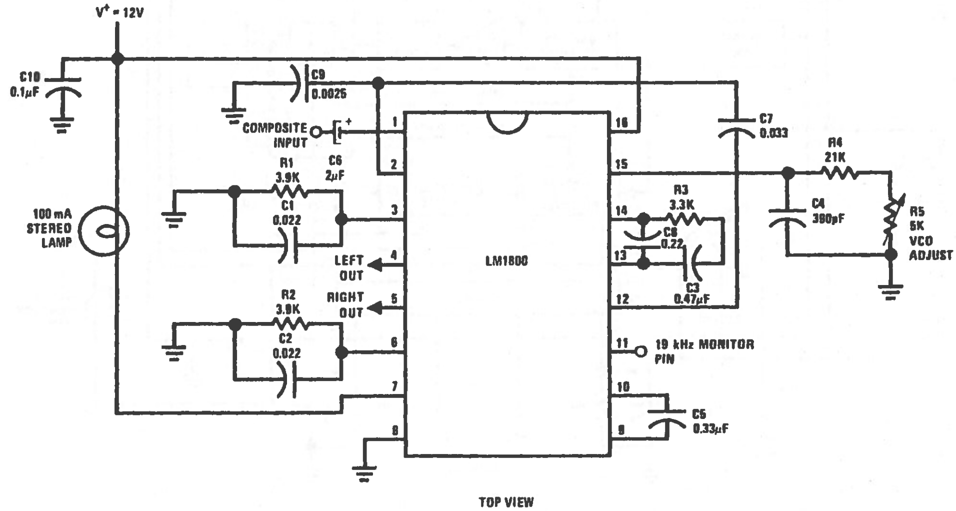

FIGURE 3.6.4 LM1800 Typical Application

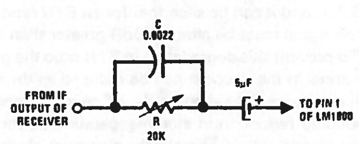

FIGURE 3.6.5 Compensation for Receiver IF Rolloff

LM1800 (pilot: 15mV$_{RMS}$ on, 6.0mV$_{RMS}$ off typical), the circuit of Figure 3.6.6 provides the desired flexibility. The user who wants slightly increased voltage gain through the demodulator can increase the size of the load resistors (R_1 and R_2 of Figure 3.6.4), being sure to correspondingly change the de-emphasis capacitors (C_1 and C_2). Loads as high as 5600Ω may be used (gain of 1.4). Performance of the LM1800 is virtually independent of the supply voltage used (from 10 to 16V) due to the on-chip regulator.

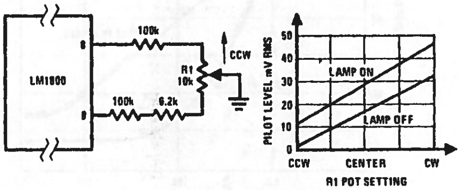

FIGURE 3.6.6 Stereo/Monaural Switch Point Adjustment

Although the circuit diagrams show a 100mA indicator lamp, the user may desire an LED. This presents no problem for the LM1800 so long as a resistor is connected in series to limit current to a safe value for the LED. The lamp or LED can be powered from any source (up to 18V), and need not necessarily be driven from the same supply as the LM1800.

3.6.3 LM1310 Typical Application

Figure 3.6.7 shows a typical stereo demodulator design using the LM1310. Capture range, lamp sensitivity adjustment and input lead compensation are all accomplished in the same manner as for the LM1800.

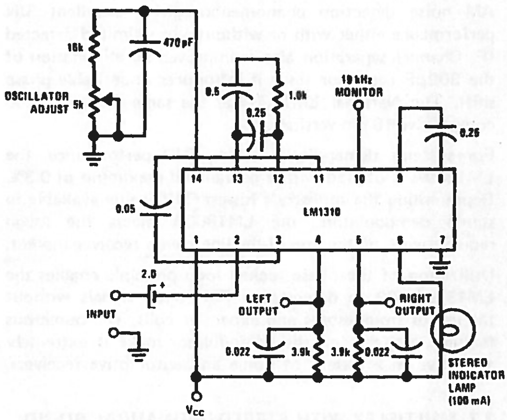

FIGURE 3.6.7 LM1310 Typical Application

3.6.4 Special Considerations of National's LM1310/1800

A number of FM stereo systems use the industry standard IF (LM3089) with an industry standard demodulator (LM1310/1800) as in Figure 3.8.8.

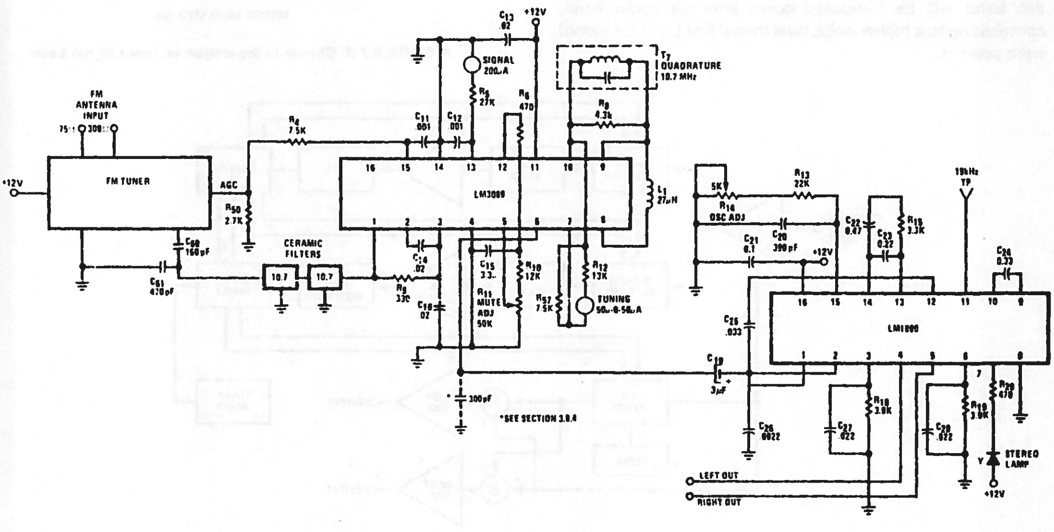

FIGURE 3.6.8 LM3089/LM1800 Application

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The optional 300pF capacitor on pin 6 of the LM3089 is often used to limit the bandwidth presented to the demodulator's input terminals. As the IF input level decreases and the limiting stages begin to come out of limiting, the detector noise bandwidth increases. Most competitive versions of the LM1310 would inadvertently AM detect this noise in their input "audio amplifier," resulting in decreased system signal-to-noise. They therefore require the 300pF capacitor, which serves to eliminate this noise from the demodulator's input by decreasing bandwidth, and thus the system maintains adequate S/N.

The National LM1310 has been designed to eliminate the AM noise detection phenomenon, giving excellent S/N performance either with or without a bandlimited detected IF. Channel separation also is improved by elimination of the 300pF capacitor since it introduces undesirable phase shift. The National LM1800 has the same feature, as do competitive 16 pin versions.

For systems demanding superior THD performance, the LM1800A is offered with a guaranteed maximum of 0.3%. Representing the industry's lowest THD value available in stereo demodulators, the LM1800A meets the tough requirements of the top-of-the-line stereo receiver market.

Utilization of the phase locked loop principle enables the LM1310/1800 to demodulate FM stereo signals without the use of troublesome and expensive coils. The numerous features available on the demodulator make it extremely attractive in a variety of home and automotive receivers.

3.7 MULTIPLEX WITH STEREO/MONAURAL BLEND

3.7.1 Introduction — Why Blend?

The signal to noise ratio of a strong, or local, stereo FM transmission is usually more than adequate. However, as many listeners to automotive radios will know, when the signal becomes weak, the S/N ratio in stereo is noticeably inferior to the S/N ratio of an equivalent strength monaural signal. Reference back to Figure 3.6.1 will show why this is the case. For a stereo broadcast a much wider frequency spectrum is used, in order to include the L-R channel information from 23kHz to 53kHz. When decoded, noise in this band will be translated down into the audio band, contributing to a higher noise level than if just L + R (or mono) were present.

Typical quieting curves for an FM stereo radio are shown in Figure 3.7.1, and it can be seen that for an S/N ratio of 50dB, the stereo signal must be almost 20dB greater than the mono signal. To prevent this degradation in S/N ratio the gain of the (L-R) channel in the decoder can be reduced as the r.f. signal strength decreases. Simultaneously, of course, there will be a corresponding reduction in stereo separation as the decoder gradually blends into a completely monaural signal output. This smooth loss of separation is much less noticeable than an abrupt switching into mono at a predetermined signal level. If an acceptable S/N ratio is 50dB then the quieting curve to be followed is given by the dashed line in Figure 3.7.1. The required decrease in L-R gain is given by Figure 3.7.2 which also shows the change in stereo separation with signal level.

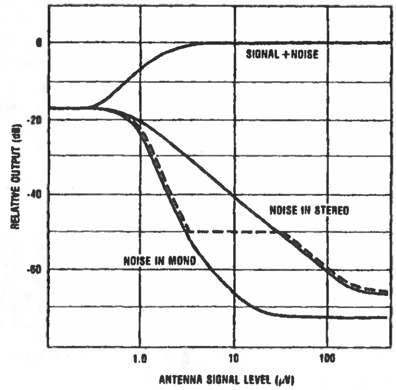

FIGURE 3.7.1 FM Radio S + N and N vs. Input Signal Level

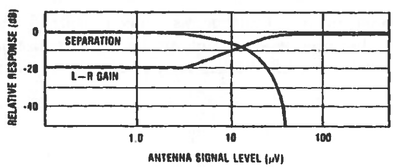

FIGURE 3.7.2 Change in Separation vs. Input Signal Level

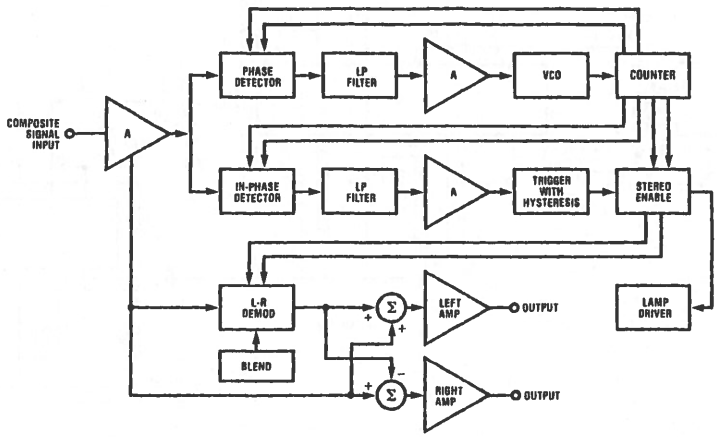

FIGURE 3.7.3 LM4500A Block Diagram

3.7.2 The LM4500A

The LM4500A is an improved stereo decoder with a new demodulation technique which minimizes subcarrier harmonics, and has a built-in blend circuit to optimize S/N ratios under weak FM signal conditions.

The block diagram of Figure 3.7.3 illustrates that the LM4500A has the same circuit functions as an LM1800, but with the addition of the blend circuit which operates on the L-R demodulator section. In this demodulation section both in-phase and antiphase components of the L-R signal are available and these can be gradually combined to finally produce complete cancellation of the L-R signal. The control voltage, which must be proportional to the r.f. signal strength, is obtained from the signal strength meter drive output of the FM IF. Usually a potentiometer adjustment will be needed to compensate for different Tuner/IF combinations. The change in separation with this control voltage is given by the curve of Figure 3.7.4

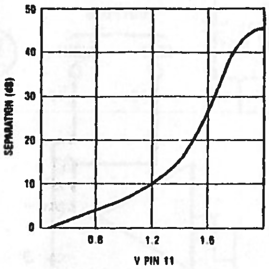

FIGURE 3.7.4 LM4500A Stereo Separation vs. Pin 11 Control Voltage

Not shown in the block diagram of Figure 3.7.3 are the different decoder switching waveforms used by the LM4500A. Conventional decoders, such as the LM1800, use square waves in-phase and anti-phase, which have a precise duty cycle, zero mean level and no even harmonics. While this provides excellent performance for the standard U.S. stereo broadcasts, problems with third harmonic radiation interference can occur in Europe where closer station spacing and the A.R.I. (Automotive Radio Information) signals are utilized. The third harmonic of the subcarrier is 114kHz, and an adjacent transmitter sideband can mix with this to produce audible components. Similarly, the third harmonic of the pilot carrier, at 57kHz, can mix with the A.R.I. system signal causing phase modulation of the V.C.O. and this results in intermodulation distortion.

The LM4500A avoids these problems by generating switching waveforms composed from square waves phase shifted such that their third harmonics are in antiphase and cancel out, Figure 3.7.5.

A complete schematic of the external components required for an LM4500A is shown in Figure 3.7.6 and this circuit exhibits at least 40dB stereo separation (optimized by P_2) and an 83dB S/N ratio. The subcarrier harmonics are typically better than 70dB down and the stereo T.H.D. is 0.07% with a 1.5V(p-p) composite signal level.

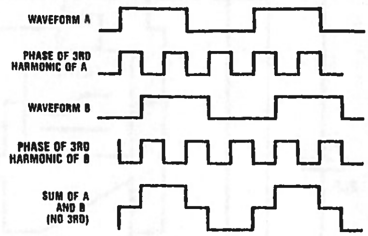

FIGURE 3.7.5 LM4500A Switching Waveform Generation

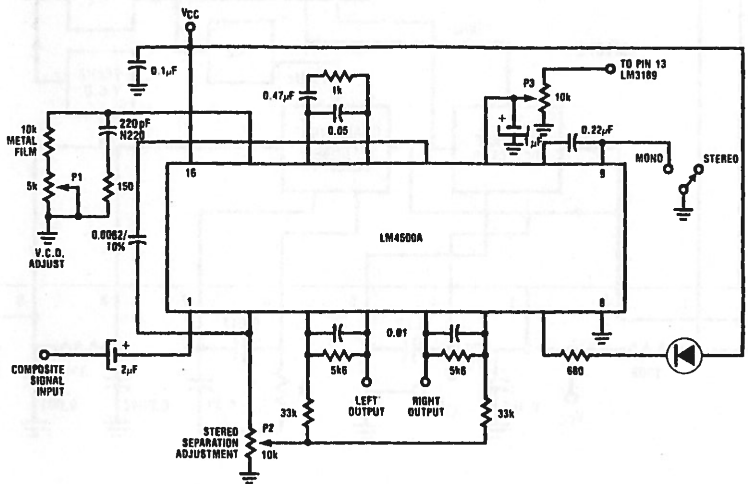

FIGURE 3.7.6 Typical Application of the LM4500A

3.7.3 The LM 1870

The LM1870 is another new stereo decoder IC from National that incorporates the variable blend feature. Instead of adding in-phase and antiphase components of the demodulated L-R signal, the LM1870 achieves stereo to mono blend before the demodulator, Figure 3.7.7. The composite input signal follows two paths, one of which has a flat, wideband frequency resonse. The other has a 2 pole low pass filter response and the output from both paths are summed in a multiplier circuit which is controlled by the r.f. signal strength. This control voltage is derived from the signal strength meter drive of the LM3189. Figure 3.7.8 shows the net result. As the r.f. signal level decreases, the h.f. portion of the composite signal containing the L-R information is decreased. At the same time the upper frequency response of the L + R signal is modified to further reduce the audible noise. Typical L+R and L-R response curves are shown in Figures 3.7.9 and 3.7.10.

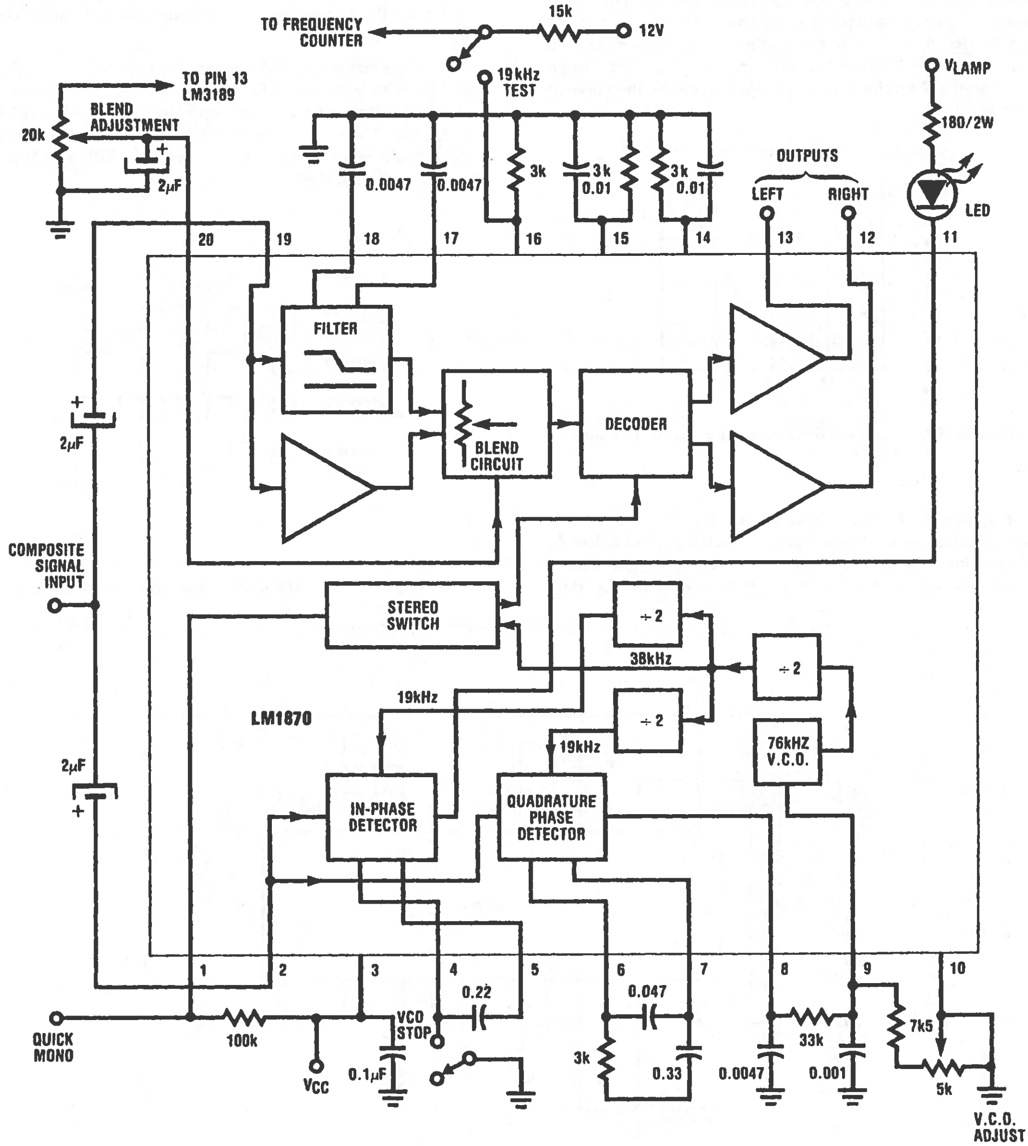

FIGURE 3.7.7 LM1870 Typical Application

The stereo performance of the LM1870 is very constant for small changes (<2%) in the free running frequency of the V.C.O. Low temperature coefficient components should be used for the oscillator capacitor and tuning resistors. Tuning the V.C.O. is done by adjusting the 5kΩ pot to obtain 19kHz ±20Hz with no signal input at Pin 2. 19kHz is available at Pin 16 if a resistor is connected from Pin 16 to the supply voltage. In normal operation, Pin 16 is connected via a resistor to ground which programs the blend characteristic, Figure 3.7.9.

Although the LM1870 outputs are low impedance and capable of sinking or sourcing 1mA, if the supply pin (Pin 3) is open or grounded, then both outputs are at a high impedance. This facilitates switching in AM-FM radios since the outputs do not have to be disconnected when the radio is in the AM mode.

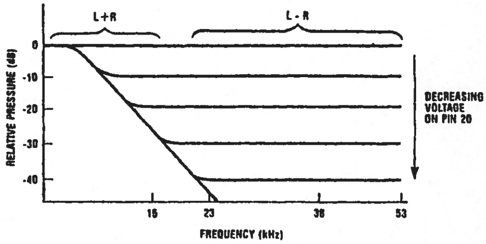

FIGURE 3.7.8 Response vs. Frequency of LM1870 Blend Circuit

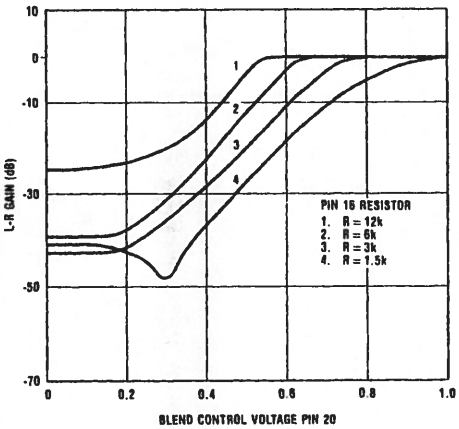

FIGURE 3.7.9 L-R Gain vs. Blend Control Voltage

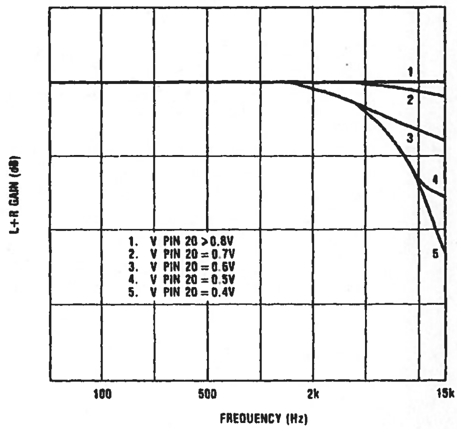

FIGURE 3.7.10 L + R Frequency Response

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