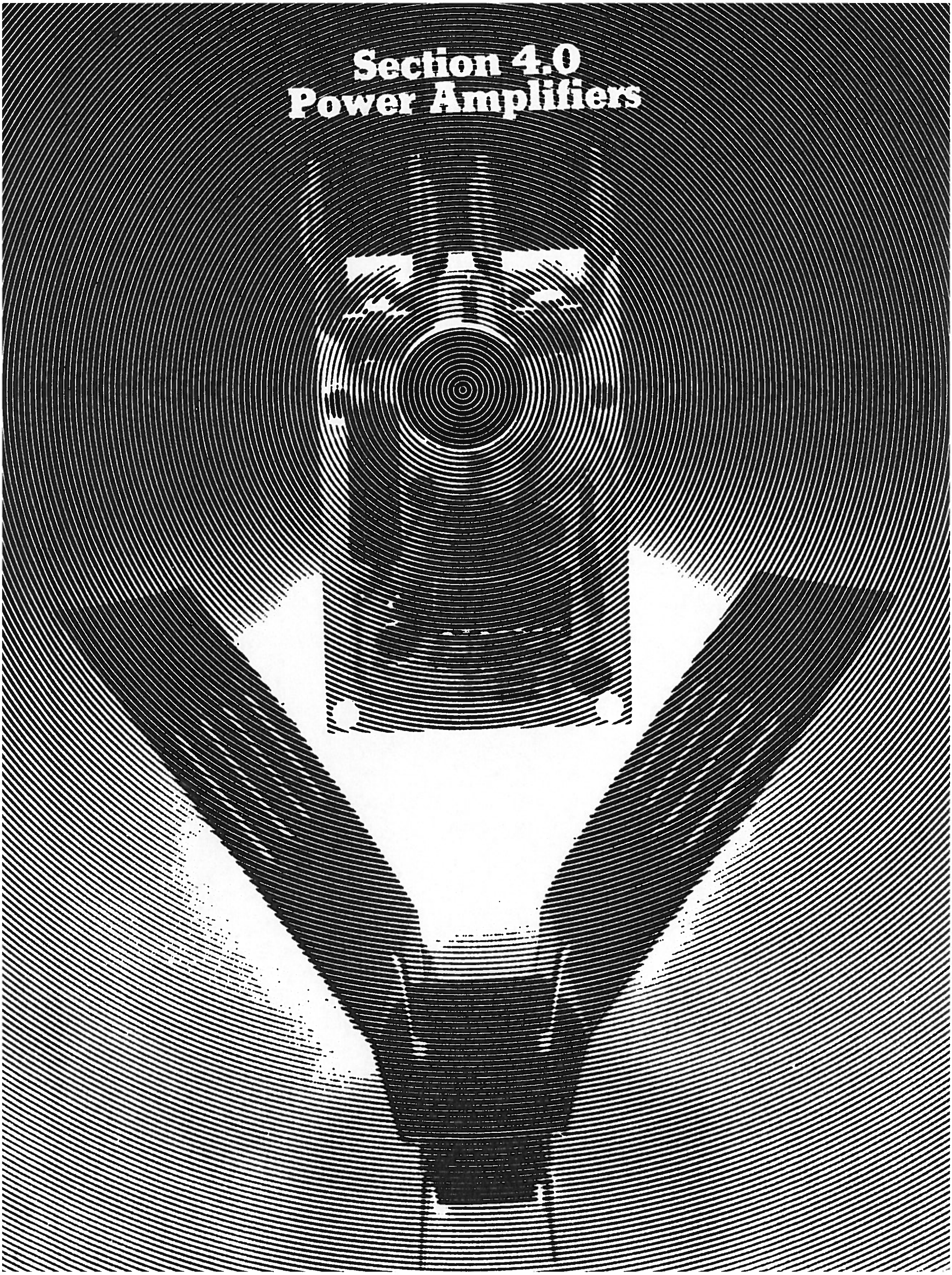
Section 4.0
Power Amplifiers



4.0 Power Amplifiers

4.1 INSIDE POWER INTEGRATED CIRCUITS

Audio power amplifiers manufactured using integrated circuit technology do not differ significantly in circuit design from traditional operational amplifiers. Use of current sources, active loads and balanced differential techniques predominate, allowing creation of high-gain, wide bandwidth, low distortion devices. Major design differences appear only in the class AB high current output stages where unique geometries are required and special layout techniques are employed to guarantee thermal stability across the chip.

The material presented in the following sections serves as a brief introduction to the design techniques used in audio power integrated circuits. Hopefully, a clearer understanding of the internal "workings" will result from reading the discussion, thus making application of the devices easier.

4.1.1 Frequency Response and Distortion

Most audio amplifier designs are similar to Figure 4.1.1. An input transconductance block ($gm = i_O/v_1$) drives a high gain inverting amplifier with capacitive feedback. To this is added an output buffer with high current gain but unity voltage gain. The resulting output signal is defined by:

$$v_O = v_1 \, gm \, X_C \qquad (4.1.1)$$

or, rewriting in terms of gain:

$$A_V = \frac{v_O}{v_1} = gm \, X_C = \frac{gm}{sC} = \frac{gm}{j\omega C} \qquad (4.1.2)$$

Setting Equation (4.1.2) equal to unity allows solution for the amplifier unity gain cross frequency:

$$A_V = 1 = \frac{gm}{j\omega C} = \frac{gm}{j2\pi fC} \qquad (4.1.3)$$

$$f_{UNITY} = \frac{gm}{2\pi C} \qquad (4.1.4)$$

Equation (4.1.2) indicates a single pole response resulting in a 20dB/decade slope of the gain-frequency plot in Figure 4.1.1. There is, of course, a low frequency pole which is determined by the compensation capacitor and the resistance to ground seen at the input of the inverting amplifier. Usually this pole is below 100Hz so it plays only a small role in determining amplifier performance in usual feedback arrangements.

For an amplifier of this type to be stable in unity gain feedback circuits, it is necessary to arrange gm and C so that the unity gain crossover frequency is about 1MHz. This is, in short, due to a few other undesirable phase shifts that are difficult to avoid when using lateral PNP transistors in monolithic realizations of the transconductance as well as the buffer blocks. Figure 4.1.1 shows that if f_{UNITY} is 1MHz then only 34dB of gain is available at 20kHz! Since most audio circuits require more gain, most IC audios are not compensated to unity. Evaluation of an IC audio amplifier will show stability troubles in loops fed back for less than 20dB closed loop gain.

Consider for a moment the problem in audio designs with distortion (THD). The buffer of Figure 4.1.1 is essentially an emitter follower (NPN during positive half cycles and PNP during negative halves due to class B operation). As a result the load presented to the collector of the gain transistor is different depending on which half cycle the output is in. The buffer amplifier itself often contributes in the form of crossover distortion. Suppose for a moment that the amplifier were to be used open loop (i.e., without any AC feedback) and that the result was an output signal distorted 10% at 10kHz. Further assume the open loop gain-frequency is as in Figure 4.1.2 so that the amplifier is running at 60dB of gain. Now add negative feedback around the amplifier to set its gain at 40dB and note that its voltage gain remains flat with frequency throughout the audio band. In this configuration there is 20dB of loop gain (the difference between open loop gain and closed loop gain) which works to correct the distortion in the output waveform by about 20dB, reducing it from the 10% open loop value to 1%. Further study of Figure 4.1.2 shows that there is more loop gain at lower frequencies which should, and does, help the THD at lower frequencies. The reduction in loop gain at high frequencies likewise allows more of the open loop distortion to show.

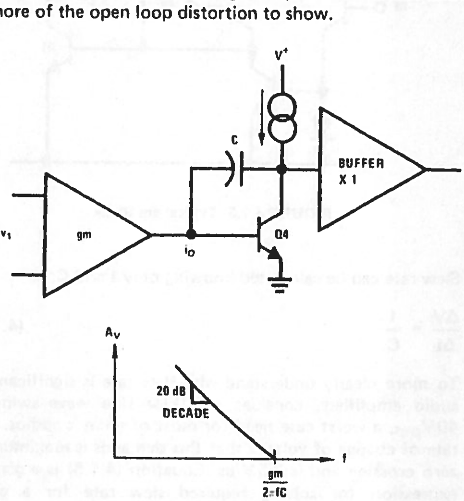


FIGURE 4.1.1 Audio Amp Small Signal Model

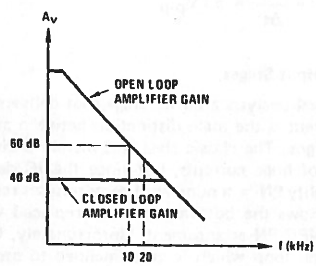


FIGURE 4.1.2 Feedback and "Loop Gain"

4.1.2 Slew Rate

Not only must IC audio amplifiers have more bandwidth than "garden variety" op amps, they must also have higher slew rates. Slew rate is a measure of the ability of an amplifier's large signal characteristics to match its own small signal responses. The transconductance block of Figure 4.1.1 delivers a current out for a given small signal input voltage. Figure 4.1.3 shows an input stage typically used in audio amplifiers. Even for large differential input voltage drives to the PNP bases, the current available can never surpass I. And this constant current (I) charging the compensation capacitor (C) results in a ramp at Q_1's collector. The slope of this ramp is defined as slew rate and usually is expressed in terms of volts per microsecond. Increasing the value of the current source does increase slew rate, but at the expense of increased input bias current and gm. Large gm values demand larger compensation capacitors which are costly in IC designs. The optimum compromise is to use large enough I to achieve adequate slew rate and then add emitter degeneration resistors to the PNPs to lower gm.

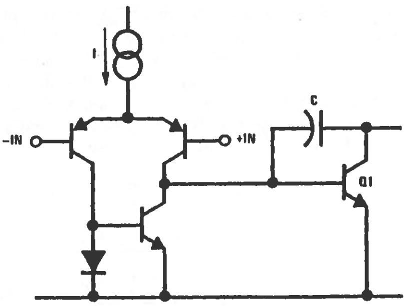


FIGURE 4.1.3 Typical gm Block

Slew rate can be calculated knowing only I and C:

$$\frac{\Delta V}{\Delta t} = \frac{I}{C} \tag{4.1.5}$$

To more clearly understand why slew rate is significant in audio amplifiers, consider a 20kHz sine wave swinging $40V_{p\text{-}p}$, a worst case need for most of today's audios. The rate of change of voltage that this demands is maximum at zero crossing and is 2.5V/µs. Equation (4.1.6) is a general expression for solving required slew rate for a given sinusoid. (See Section 1.2.1.)

$$\text{Slew rate} = \frac{\Delta V}{\Delta t} = \pi f V_{p\text{-}p} \tag{4.1.6}$$

4.1.3 Output Stages

In the final analysis a buffer stage that delivers amperes of load current is the main distinction between audio and op amp designs. The classic class B is merely a PNP and NPN capable of huge currents, but since the IC designer lacks good quality PNPs, a number of compromises results. Figure 4.1.4b shows the bottom side PNP replaced with a composite PNP/NPN arrangement. Unfortunately, Q_2/Q_3 form a feedback loop which is quite inclined to oscillate in the 2-5MHz range. Although the oscillation frequency is well above the audible range, it can be troublesome when placed in proximity to an RF receiver. Among the stabilization techniques that are in use, with varying degrees of success are:

1. Placing an external RC from the output pin to ground to lower the gain of the NPN. This works pretty well and appears on numerous data sheets as an external cure.
2. Utilizing device geometry methods to improve the PNP's frequency response. This has been done successfully in the LM378 and LM379. The only problem with this scheme is that biasing the improved PNP reduces the usable output swing slightly, thereby lowering output power capability.
3. Addition of resistance in series with either the emitter or base of Q_3.
4. Making Q_3 a controlled gain PNP of unity, which has the added advantage of keeping gain more nearly equal for each half cycle.
5. Adding capacitance to ground from Q_3's collector.

These last three work sometimes to some degree at most current levels.

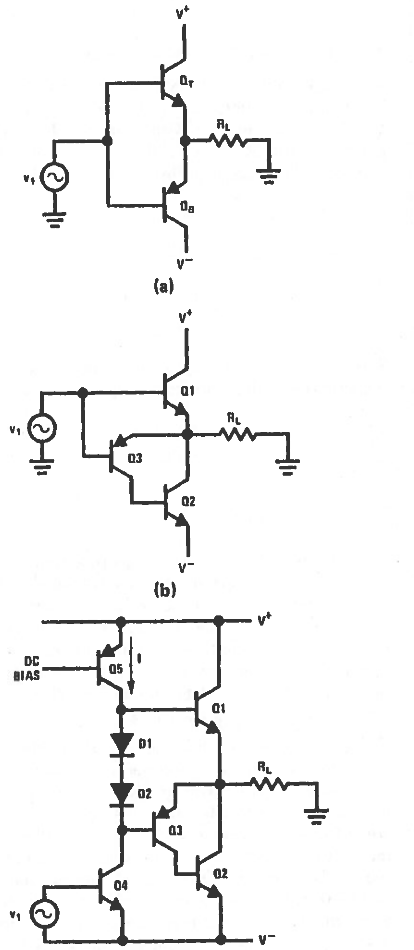


FIGURE 4.1.4 Basic Class B Output Drivers

Figure 4.1.5 illustrates crossover distortion such as would result from the circuit in Figure 4.1.4b. Beginning with Q_1 "on" and the amplifier output coming down from the top half cycle towards zero crossing, it is clear that the emitter of Q_1 can track its base until the emitter reaches zero volts. However, as the base voltage continues below 0.7V, Q_1 must turn off; but Q_2/Q_3 cannot turn on until the input generator gets all the way to –0.7V. Thus, there is a 1.4V of dead zone where the output cannot respond to the input. And since the size of the dead zone is independent of output amplitude, the effect is more pronounced at low levels. Of course feedback works to correct this, but the result is still a somewhat distorted waveform – one which has an unfortunately distasteful sound. Indeed the feedback loop or the composite PNP sometimes rings as it tries to overcome the nonlinearity, generating harmonics that may disturb the receiver in radio applications. The circuit of Figure 4.1.4c adds "AB bias." By running current through D_1 and D_2, the output transistors are turned slightly "on" to allow the amplifier to traverse the zero volts region smoothly. Normally much of the power supply current in audio amplifiers is this AB bias current, running anywhere from 1 to 15mA per amplifier.

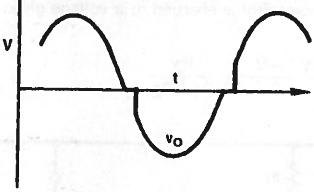


FIGURE 4.1.5 Crossover Distortion

Some amplifiers at high frequencies (say 10kHz) exhibit slightly more crossover distortion when negative going than when positive going through zero. This is explained by the slow composite PNPs' (Q_2/Q_3) delay in turning "on." If the amplifier delivers any appreciable load current in the top half cycle, the emitter current of Q_1 causes its base-emitter voltage to rise and shut "off" Q_3 (since the voltage across D_1 and D_2 is fixed by I). Thus, fast negative going signals demand the composite to go from full "off" to full "on" – and they respond too slowly. As one might imagine, compensating the loop (Q_2 and Q_3) for stability even slows the switching time more. This problem makes very low distortion IC amplifiers (< 0.2%) difficult at the high end of the audio (20kHz).

Another interesting phenomenon occurs when some IC amplifiers oscillate at high frequencies – their power supply current goes up and they die! This usually can be explained by positive going output signals where the fast top NPN transistor (Q_1) turns "on" before the sluggish composite turns "off," resulting in large currents passing straight down through the amplifier (Q_1 and Q_2).

The distortion components discussed so far have all been in terms of circuit nonlinearities and the loop gain covering them up. However, at low frequencies (below 100Hz) thermal problems due to chip layout can cause distortion. In the audio IC, large amounts of power are dissipated in the output driver transistors causing thermal gradients across the die. Since a sensitive input amplifier shares the same piece of silicon, much care must be taken to preserve thermal symmetry to minimize thermal feedback.

Despite the many restrictions on audio IC designs, today's devices do a credible job, many boasting less than 1% THD from 20Hz to 20kHz – not at all a bad feat!

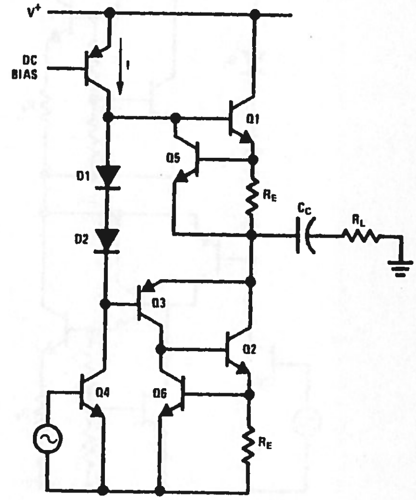


FIGURE 4.1.6 Simple Current Limit

4.1.4 Output Protection Circuitry

By the very nature of audio systems the amplifier often drives a transducer – or speaker – remote from the electronic components. To protect against inadvertent shorting of the speaker some audio ICs are designed to self limit their output current at a safe value. Figure 4.1.6 is a simple approach to current limiting: here Q_5 or Q_6 turns "on" to limit base drive to either of the output transistors (Q_1 or Q_2) when the current through the emitter resistors is sufficient to threshold an emitter base junction. Limiting is sharp on the top side since Q_5 has to sink only the current source (I). However, the current that Q_6 must sink is more nebulous, depending on the alpha holdup of Q_3, resulting in soft or mushy negative side limiting. Other connections can be used to sharpen the limiting action, but they usually result in a marginally stable loop that must be frequency compensated to avoid oscillation during limiting. The major disadvantage to the circuit of Figure 4.1.6 is that as much as 1.4V is lost from loaded output swing due to voltage dropped across the two R_Es.

The improved circuit of Figure 4.1.7 reduces the values of R_E for limiting at the same current but is usable only in Darlington configurations. It suffers from the same negative side softness but only consumes about 0.4V of output swing. There are a few other methods employed, some even consuming less than 0.4V. Indeed it is further possible to add voltage information to the current limit transistor's base and achieve safe operating area protection. Care must be taken in such designs, however, to allow for a leading or lagging current of up to 60° to accommodate the variety of speakers on the market. However, the circuitry shown in Figures 4.1.6 and 4.1.7 is representative of the vast majority of audio ICs in today's marketplace.

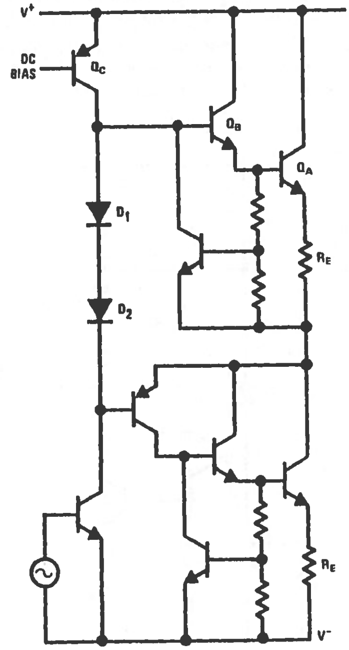


FIGURE 4.1.7 Improved Current Limit

Large amounts of power dissipation on the die cause chip temperatures to rise far above ambient. In audio ICs it is popular to include circuitry to sense chip temperature and shut down the amplifier if it begins to overheat. Figure 4.1.8 is typical of such circuits. The voltage at the emitter of Q_1 rises with temperature due both to the TC of the zener (Z_1) and Q_1's base-emitter voltage. Thus, the voltage at the junction of R_1 and R_2 rises while the voltage required to threshold Q_2's emitter-base junction falls with temperature. In most designs the resistor ratio is set to threshold Q_2 at about 165°C. The collector current of Q_2 is then used to disable the amplifier.

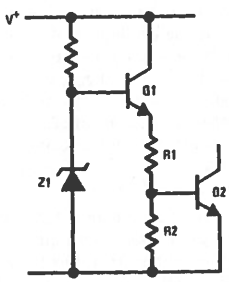


FIGURE 4.1.8 Typical Thermal Shutdown

The addition of thermal shutdowns in audio ICs has done much to improve field reliability. If the heat sinking is inadequate in a discrete design, the devices burn up. In a thermally protected IC the amplifier merely reduces drive to the load to maintain chip temperature at a safe value.

4.1.5 Bootstrapping

A look at the typical Class B output stage of Figure 4.1.4 shows that the output swings positive only until Q_5 saturates (even unloaded). At this point the output voltage swing lost across Q_1 is

$$V^+ - V_{OUTPEAK} = V_{SAT}(Q_5) + V_{BE}(Q_1) \approx 1.1V \quad (4.1.7)$$

Further, the output swings negative until Q_4 saturates when the output voltage swing loss is

$$V^- - V_{OUTPEAK} = V_{SAT}(Q_4) + V_{BE}(Q_3) \approx 0.9V \quad (4.1.8)$$

Despite the fact that there is no load current, the maximum possible output swing is about 2V less than the total supply voltage. While it is possible that with very high load currents the saturation voltages of Q_1 and Q_2 can exceed 1V each, most audio I/C's are limited by Equations (4.1.7) and (4.1.8). For battery operated systems in particular, this loss in output swing can seriously reduce the available output power to the load.

Larger positive swings can be obtained by utilizing "bootstrap" techniques (Figure 4.1.9). In the quiescent state the amplifier output is halfway between the supply voltages so that the capacitor is charged to a voltage given by,

$$V_{CBS} \approx \frac{V^+ - V^-}{2} \times \frac{R_2}{R_1 + R_2} \quad (4.1.9)$$

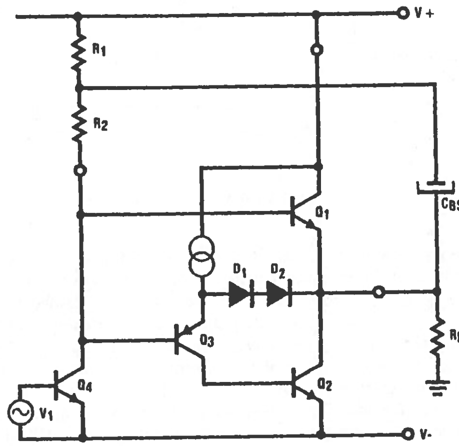


FIGURE 4.1.9 "Bootstrap" Output State

Less the diode drop of Q_1 base-emitter, this is the voltage across R_2. If C_{BS} is very large this voltage will remain constant, even as the output swings positive, so the voltage across R_2 (and consequently the current through R_2) will be maintained. In this way the bootstrap circuitry appears to be a current source to Q_4 and during extreme positive output excursions the base of Q_1 can be pulled above the supply rail, leaving the output swing limited only by Q_1 saturation voltage.

To improve the negative output swing the AB bias network of a current souce and diodes D_1 and D_2 is connected in the emitter circuit of Q_3. Now when Q_4 approaches saturation, the output voltage loss is given by,

$$V^- - V_{OUTPEAK} = V_C(Q_4) + V_{BE}(Q_3) - V_{D1} - V_{D2} \quad (4.1.7)$$

If Q_4 is allowed to saturate, this could be as low as

$0.2V + 0.7V - 1.2V = -0.3V!$

In practice the saturation voltage of Q_2 will define the lowest negative excursion — which will occur before Q_4 saturates.

4.2 DESIGN TIPS ON LAYOUT, GROUND LOOPS AND SUPPLY BYPASSING

Layout, grounding and power supply decoupling of audio power integrated circuits require the same careful attention to details as preamplifier ICs. All of the points discussed in Section 2.2 of this handbook apply directly to the use of power amplifiers and should be consulted before use. The relevant sections are reproduced here for cross-reference and convenience:

- Section 2.2.1 Layout
- Section 2.2.2 Ground Loops
- Section 2.2.3 Supply Bypassing
- Section 2.2.4 Additional Stabilizing Tips

4.3 POWER AMPLIFIER SELECTION

National Semiconductor's line of audio power amplifiers consists of two major families: the "Duals" represented by LM1877/LM2877, LM378, LM378, LM1896/2896, and the "Monos" represented by ten products. Available power output ranges from miniscule 320mW battery operated devices to hefty 9.6W line operated systems. The power driver LM391 is capable of driving output stages, delivering 60W. Although most of the amplifiers are designed for single supply operation, all devices may be operated from split supplies where required. Tables 4.3.1 and 4.3.2 summarize the dual family for ease of selection, while Table 4.3.3 compares the mono devices.

TABLE 4.3.1 Dual Power Amplifier Characteristics

PARAMETER	LM1877/LM2877			LM378/LM379			LM1896/LM2896		
	MIN	TYP	MAX	MIN	TYP	MAX	MIN	TYP	MAX
Supply Voltage	6V	20V	24V	10V	24V	35V	3V	6V	10V/15V
Quiescent Supply Current ($P_{OUT} = 0W$)		25mA	50mA		15mA	65mA		15mA/ 25mA	
Open Loop Gain ($R_s = 0\Omega$, $f = 1kHz$)		70dB		66dB	90dB			100dB	
Input Impedance		4mΩ		3mΩ				100kΩ	
Channel Separation[1] Output Referred ($C_F = 50\mu F$, $f = 1kHz$)	−50dB	−70dB			−36dB[2]			−60dB	
Power Supply Ripple Rejection ($C_F = 50\mu F$, $f = 120Hz$)	−50dB	−68dB			−36dB[2]			−54dB	
Equivalent Input Noise ($R_s = 600\Omega$, BW 20Hz to 20kHz)		3μV			3μV			1.9μV	
THD $f = 1kHz$									
$P_o = 1W$/Channel		0.07%			0.07%			-/0.14%	
$P_o = 50mW$/Channel		0.1%			0.25%			0.09/ 0.27%	

1. $A_V = 34dB$
2. $C_F = 250\mu F$
3. Package Styles:
   - LM1877/LM378/LM1896 — 14 Pin D.I.P.
   - LM2877/LM2896 — 11 Pin S.I.P.
   - LM379 — 14 Pin 'S' Type Power D.I.P.

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TABLE 4.3.2 Dual Amplifier Output Power

V_{CC} (VOLTS)	LOAD (OHMS)	LM1877	TYPICAL OUTPUT POWER (WATTS) AT 10% T.H.D.				LM2896
			LM2877	LM387	LM389	LM1896	
6V	4Ω					1.1W	
	8Ω					600mW	
	8Ω (Bridge)					2.2W	
9V	4Ω						2.5W
	4Ω (Bridge)						7.8W
	8Ω	500mW	500mW			1.3W	
	8Ω (Bridge)						5W
12V	4Ω		1.9W				
	8Ω	1.2W	1.2W	1.6W	1.6W		2.5W
	8Ω (Bridge)						9.0W
14V	8Ω	1.8W	1.8W	1.9W	1.9W		
18V	8Ω		3.6W	3W	3W		
	16Ω			1.8W	1.8W		
20V	8Ω		4.5W		3.8W		
	16Ω			2.4W	2.4W		
24V	8Ω				5.4W		
	16Ω			3.5W	3.6W		
30V	16Ω				5.5W		

1. Specification apply for $T_{TAB} = 25°C$. For operation at higher ambient temperatures, the IC must be derated based on the package/heatsink thermal resistance and a 150°C max junction temperature.

TABLE 4.3.3 Mono Power Amplifier Characteristics

V_{CC} (VOLTS)	DEVICE TYPE	OUTPUT POWER (WATTS) AT 10% THD[2]				GAIN (dB)	OUTPUT PROTECTION
		2Ω	4Ω	8Ω	16Ω		
3V	LM2000/1	480mW	280mW	160mW		ADJUSTABLE	NO
6V	LM383	1.9W	800mW	440mW	240mW	ADJUSTABLE	YES
	LM386/389		340mW	325mW	180mW	26-46dB	NO
	LM388		800mW	600mW	300mW	26-46dB	NO
	LM390		1.0W	650mW	325mW	26-46dB	NO
	LM2000/1	2.0W	1.2W	600mW		ADJUSTABLE	NO
9V	LM383	3.5W	2.1W	1.2W	630mW	ADJUSTABLE	YES
	LM386/389		350mW/-	700mW/520mW	500mW	26-46dB	NO
	LM388		1.8W	1.3W	650mW	26-46dB	NO
	LM390		2.0W	1.4W	700mW	26-46dB	NO
	LM2000	4.8W	2.8W	1.5W		ADJUSTABLE	NO
12V	LM380		2.4W	1.5W	500mW	34dB	NO
	LM383	6.4W	4.0W	2.3W	1.2W	ADJUSTABLE	YES
	LM386/389		350mW/-	820mW/-	1.6W/900mW	26-46dB	NO
	LM388		2.4W	2.2W	1.3W	26-46dB	NO
	LM2000	8.8W	5.0W	2.6W		ADJUSTABLE	NO
14V	LM380		3.3W	2.2W	1.0W	34dB	NO
	LM383	8.9W	5.6W	3.7W	1.7W	ADJUSTABLE	YES
	LM386			830mW	1.3W	26-46dB	NO
	LM388		3.0W	3.0W	1.8W	26-46dB	NO
16V	LM380			3.0W	1.6W	34dB	NO
	LM383	10.5W	7.0W			ADJUSTABLE	YES
	LM386				1.6W	26-46dB	NO
	LM388		3.6W	3.8W	2.3W	26-46dB	NO
18V	LM380			4.0W	2.2W	34dB	NO
	LM383		9.6W	5.5W	2.9W	ADJUSTABLE	YES
	LM384		4.2W	4.0W	2.2W		
22V	LM384		3.5W	5.7W	3.5W	34dB	NO
±22V	LM39IN-60		30W	20W		ADJUSTABLE	YES
±30V	LM39IN-80		60W	40W		ADJUSTABLE	YES

1. Specifications apply for $T_A = 25°C$. For operation at ambient temperatures > 25°C the IC must be derated based on the case style thermal resistance and a maximum 150°C junction temperature
2. P_O increases by 19% at 5% THD and by 30% at 10% THD. Clipping occurs just before 3% THD is reached.

4.4 LM1877, LM1896, LM378, AND LM379 DUAL TWO TO SIX WATT POWER AMPLIFIERS

4.4.1 Introduction

The "Duals" are two channel power amplifiers capable of delivering up to 6 watts into 8 or 16Ω loads. They feature on-chip frequency compensation, output current limiting, thermal shutdown protection, fast turn-on and turn-off without "pops" or pulses of active gain, an output which is self-centering at $V_{CC}/2$, and a 5 to 20MHz gain-bandwidth product. Applications include stereo or multi-channel audio power output for phono, tape or radio use over a supply range of 6 to 35V, as well as servo amplifier, power oscillator and various instrument system circuits. Normal supply is single-ended; however, split supplies may be used without difficulty or degradation in power supply rejection.

4.4.2 Circuit Description of LM378 and LM379

The simplified schematic of Figure 4.4.1 shows the important design features of the amplifier. The differential input stage made up of Q_1-Q_4 uses a double (split) collector PNP Darlington pair having several advantages. The high base-emitter breakdown of the lateral PNP transistor is about 60V, which affords significant input over-voltage protection. The double collector allows operation at high emitter current to achieve good first stage f_t and minimum phase shift while simultaneously operating at low transconductance to allow internal compensation with a physically small capacitor C_1. (Unity gain bandwidth of an amplifier with pole-splitting compensation occurs where the first stage transconductance equals ωC_1.)

Further decrease of transconductance is provided by degeneration caused by resistors at Q_2 and Q_3 emitters, which also allow better large signal slew rate. The second collector provides bias current to the input emitter follower for increased frequency response and slew rate. Full differential input stage gain is provided by the "turnaround" differential to single-ended current source loads Q_5 and Q_6. The input common-mode voltage does not extend below about 0.5V above ground as might otherwise be expected from initial examination of the input circuit. This is because Q_7 is actually preceded by an emitter follower transistor not shown in the simplified circuit.

The second stage Q_7 operates common-emitter with a current source load for high gain. Pole splitting compensation is provided by C_1 to achieve unity gain bandwidth of about 10MHz. Internal compensation is sufficient with closed-loop gain down to about $A_V = 10$.

The output stage is a complementary common-collector class AB composite. The upper, or current sourcing section, is a Darlington emitter follower Q_{12} and Q_{13}. The lower, or current sinking, section is a composite PNP made up of Q_{14}, Q_{15}, and Q_9. Normally, this type of PNP composite has low f_t and excessive delay caused by the lateral PNP transistor Q_9. The usual result is poor unity gain bandwidth and probable oscillation on the negative half of the output waveform. The traditional fix has been to add an external series RC network from output to ground to reduce loop gain of the composite PNP and so prevent the oscillation. In the LM378/LM379 amplifiers, Q_9 is a field-aided lateral PNP to overcome these performance limitations and so reduce external parts count. There is no need for the external RC network, no oscillation is present on the negative half cycle, and bandwidth is better with this output stage. Q_{10} and Q_{11} provide output current limiting at

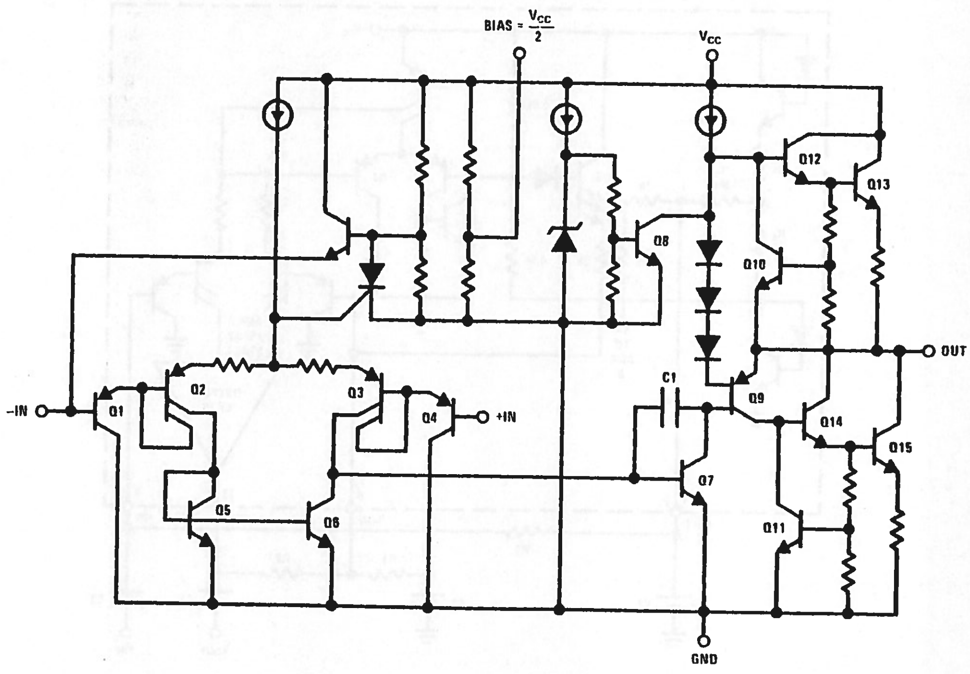

FIGURE 4.4.1 Simplified Schematic Diagram

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about 1.3A, and there is internal thermal limiting protection at 150°C junction temperature. The output may be AC shorted without problem; and, although not guaranteed performance, DC shorts to ground are acceptable. A DC short to supply is destructive due to the thermal protection circuit which pulls the output to ground.

To achieve a stable DC operating point, it is desirable to close the feedback loop with unity DC gain. To achieve this simultaneously with a high AC gain normally requires a fairly large bypass capacitor, C_1, in Figure 4.4.2.

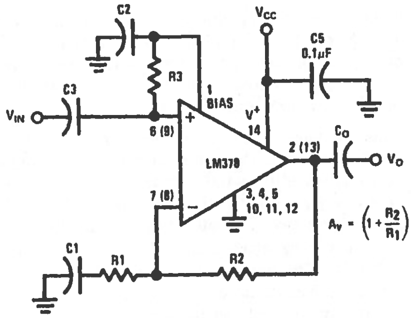

FIGURE 4.4.2 Non-Inverting Amplifier Connection

Establishing the initial charge on this capacitor results in a turn-on delay. An additional capacitor, C_2, is normally required to supply a ripple-free reference to set the DC operating point. To achieve good supply rejection X_{C2} is normally made much smaller than a series resistor from the bias divider circuit (R_S in Figure 4.4.3). Where a supply rejection of 40dB is required with 40dB closed-loop gain, 80dB ripple attenuation is required of R_SC_2. The turn-on time can be calculated as follows:

$$PSRR = \frac{R_S - jX_{C2}}{X_{C2}} \approx \frac{R_S}{X_{C2}} = \omega RC = \omega T$$

$$T = \frac{PSRR}{\omega} = \frac{80\text{dB}}{2\pi\,120\,\text{Hz}} = \frac{10^4}{754} = 13.3\text{sec}$$

$$t_{ON} \approx \frac{T}{3} = 4.5 \text{ seconds to small signal operation}$$

$$t_{ON} \approx 3T = 40 \text{ seconds to full output voltage swing}$$

The 3T delay might normally be considered excessive! The LM378/379 amplifiers incorporate active turn-on circuitry to eliminate the long turn-on time. This circuitry appeared in Figure 4.4.1 as Q_{16} and an accompanying SCR; it is repeated and elaborated in Figure 4.4.3. In operation, the turn-on circuitry charges the external capacitors, bringing output and input levels to $V_{CC}/2$, and then disconnects itself leaving only the $V_{CC}/2$ divider R_B/R_B in the circuit.

The turn-on circuit operation is as follows. When power is applied, approximately $V_{CC}/2$ appears at the base of Q_{16}, rapidly charging C_1 and C_2 via a low emitter-follower output impedance and series resistors of 3k and 1k. This causes the emitters of the differential input pair to rise to $V_{CC}/2$, bringing the differential amp Q_3 and Q_4 into balance. This, in turn, drives Q_3 into conduction. Transistors

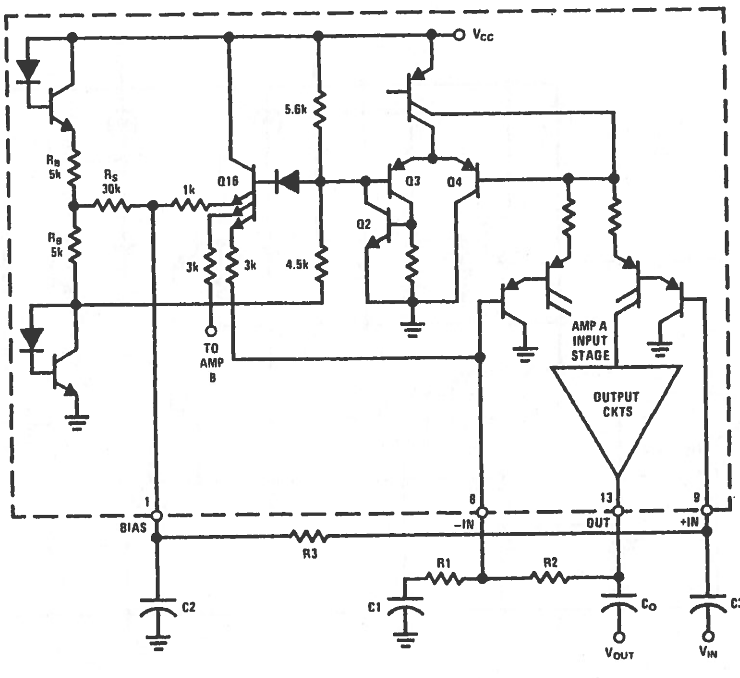

FIGURE 4.4.3 Internal Turn-On Circuitry

Q_2 and Q_3 form an SCR latch which then triggers and clamps the base of Q_{16} to ground, thus disabling the charging circuit. Once the capacitors are charged, the internal voltage divider R_B/R_B maintains the operating point at $V_{CC}/2$. Using $C_2 = 250\mu F$, the $t_{ON} = 3T \approx 0.3s$ and PSRR $\approx$ 75dB at 120Hz due to the 30k resistor R_S. Using $C_2 = 1000\mu F$, PSRR would be 86dB. The internal turn-on circuit prevents the usual "pop" from the speaker at turn-on. The turn-off period is also pop-free, as there is no series of pulses of active gain often seen in other similar amplifiers.

Note that the base of Q_4 is tied to the emitters of only one of the two input circuits. Should only one amplifier be in use, it is important that it be that with input at pins 8 and 9.

4.4.3 External Biasing Connection

The internal biasing is complete for the inverting gain connection of Figure 4.4.4 except for the external C_2 which provides power supply rejection. The bias terminal 1 may be connected directly to C_2 and the non-inverting input terminals 6 and 9. Normal gain-set feedback connections to the inverting inputs plus input and output coupling capacitors complete the circuitry. The output will Q up to $V_{CC}/2$ in a fraction of one second.

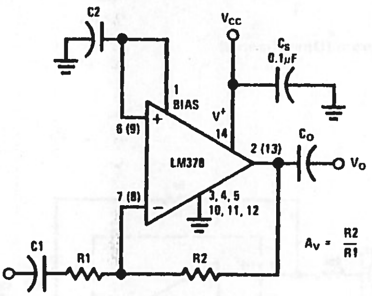

FIGURE 4.4.4 Inverting Amplifier Connection

The non-inverting circuit of Figure 4.4.2 is only slightly more complex, requiring the input return resistor R_3 from input to the bias terminal and additional input capacitor C_3. C_1 must remain in the circuit at the same or larger value than in Figure 4.4.4.

4.4.4 Circuit Description of LM1877 and LM2877

The LM1877 is a dual power amplifier designed to deliver 2W/channel continuously into 8Ω loads. It has an identical pin-out to the older LM377 and is intended as a direct replacement for that device in most applications.

The LM1877 differs internally in several respects from the LM378-LM379 series as shown by Figure 4.4.5 A differential input stage of NPN Darlington pairs is used and is optimized to give low equivalent input noise when the amplifier is driven from low impedance sources. Coupling to the second stage is through the current mirror Q_5. Note that Q_5 will hold the collector of Q_3 at 0.7V above $V_{CC}/2$ (Pin 1 bias level). This will limit the input voltage swing at Q_1 or Q_4 base to +700mV above $V_{CC}/2$. To accomodate input voltage swings that go higher than half supply (comparator or stereo amplifier applications) Pin 1 can be externally connected to Pin 14. The second stage is compensated internally for a unity gain bandwidth of 6mHz, which helps minimize the chance of rf radiation from the LM1877 into adjacent circuits (an AM radio input stage for example).

A large output swing capability is obtained by configuring the output stage and protection circuitry as shown in Section 4.1.4. Therefore an external R-C network from the output to ground is required to suppress oscillations that can occur during negative going signal swings as noted in Section 4.1.3.

Biasing for the amplifier stages is from a ΔV_{BE} reference voltage circuit (Q_{102}-Q_{109}) instead of from a zener, to allow operation with supplies as low as 6 volts. Q_{102}, Q_{103} and Q_{104} form the start-up circuit for the voltage reference by bleeding base current for Q_{105} (and hence Q_{108}) at turn-on. The double collector of Q_{108} will deliver equal currents to Q_{106} and Q_{107} which have a 4:1 ratio in emitter size. For transistors with a current density ratio of R, the difference in base-emitter voltage is given by,

$$\Delta V_{BE} = \frac{kT}{q} \log_e R$$

$$= 36mV \text{ for } R = 4 @ T = 300°K$$

In order for Q_{106} to have the same base voltage as Q_{107} (when it has the same current but one quarter the current density), this 36mV must appear across the 360Ω resistor in Q_{106} emitter. This sets the current level in the devices to $100\mu A$ so that a temperature compensated voltage of $0.7V + (100 \times 10^{-6} \times 5 \times 10^3)V = 1.2V$ appears at the base of Q_{109}. Once the circuit has started up, the current flowing in the 5k resistor in Q_{106} collector circuit wil cause Q_{105} to be shut off.

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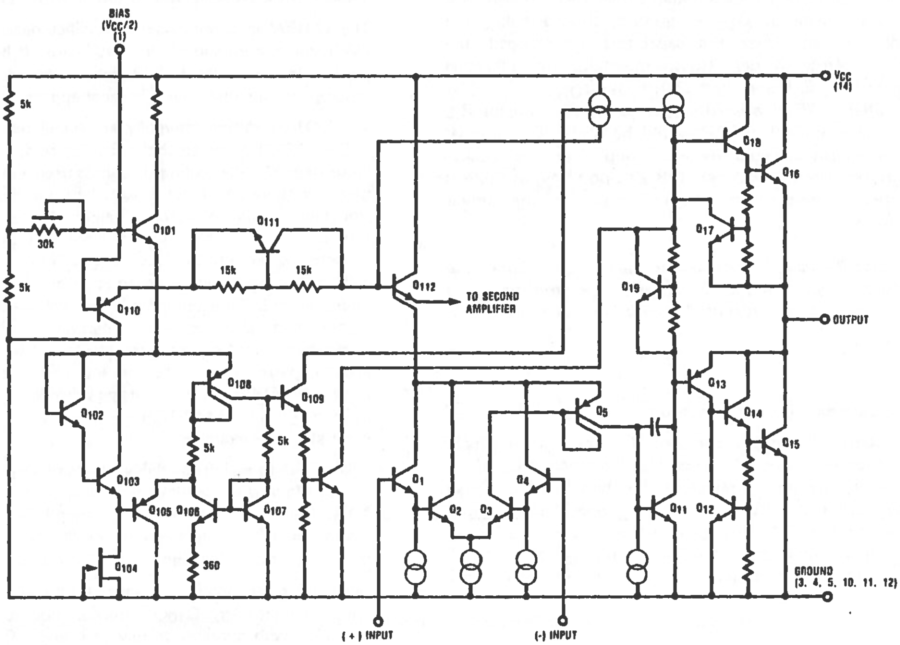


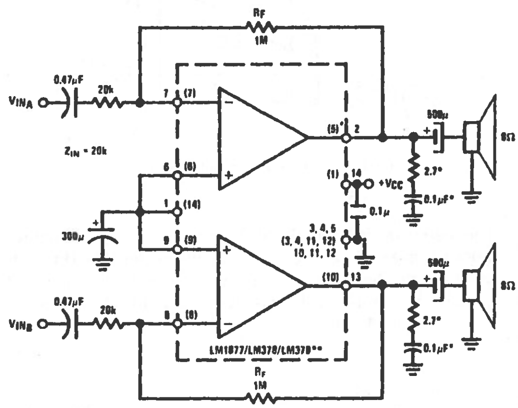
FIGURE 4.4.5 LM1887 Schematic Diagram (One Channel)

4.4.5 The LM1896 and LM2896

The newly introduced LM1896 is a dual power amplifier which has been optimized for maximum power output on low voltage supplies. As shown in Table 4.3.2., with a 6 volt supply, the LM1896 can deliver 1W/Ch into 4Ω or 2W into 8Ω when configured as a bridge amplifier. Good output swing capability is obtained by bootstrapping the output stages (Section 4.1.4) and a unique circuit design ensures low r.f. noise radiation — particularly important for obtaining high sensitivity and good S/N ratios in AM radios. Operation down to 3 volts and a low quiescent current drain of around 12mA make the LM1896 ideally suited for battery operated equipment requiring relatively high audio power output levels.

4.4.6 Stereo Amplifier Applications

The obvious and primary intended application is as an audio frequency power amplifier for stereo music systems. The amplifiers may be operated in either the non-inverting or the inverting modes of Figures 4.4.2 and 4.4.4. The inverting circuit has the lowest parts count so is most economical when driven by relatively low-impedance circuitry. Figure 4.4.6 shows the total parts count for such a stereo amplifier. The feedback resistor value of 1 meg in Figure 4.4.6 is about the largest practical value due to an input bias current max of approximately 1/2μA (100nA typ). This will cause a −0.1 to 0.5V shift in DC output level, thus limiting peak negative signal swing. This output voltage shift can be corrected by the addition of series resistors (equal to the RF in value) in the + input lines. However, when this is done, a potential exists for high frequency instability due to capacitive coupling of the



* (LM1877 ONLY)
** (LM379S PIN NOS. IN PARENTHESES)

		LM1877	LM378	LM379
P_O	=	2W/CH	3W/CH	4W/CH
e_i	=	80mV MAX	98mV MAX	113mV MAX
A_V	=	50	50	50
V_{CC}	=	18V	24V	28V

FIGURE 4.4.6 Inverting Stereo Amplifier

output signal to the + input. Bypass capacitors could be added at + inputs to prevent such instability, but this increases the parts count equal to that of the non-inverting circuit of Figure 4.4.7, which has a superior input impedance. For applications utilizing high impedance tone and volume controls, the non-inverting connection will normally be used.

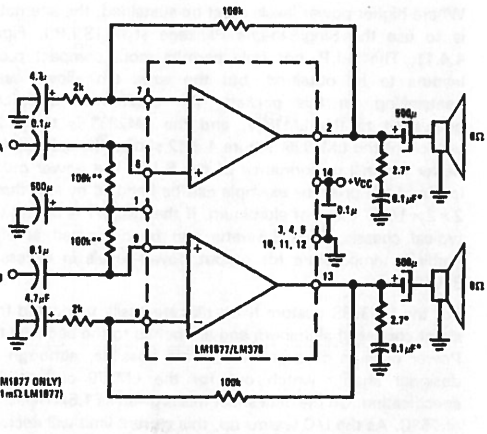


FIGURE 4.4.7 Non-Inverting Stereo Amplifier

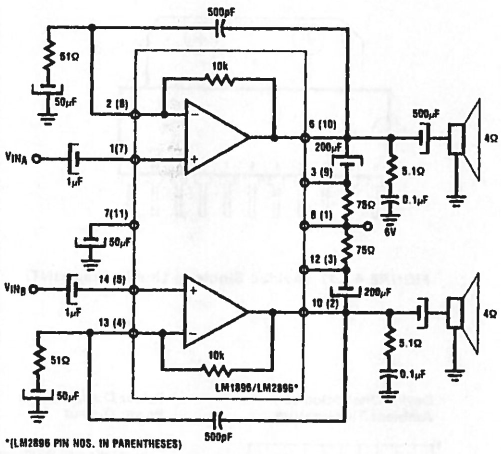


FIGURE 4.4.8 Low Voltage Stereo Amplifier 1.1W/Ch

4.4.7 Power Output per Channel (Both Channels Driven)

Figure 4.4.9 gives the package dissipation for the dual amplifiers with different supply voltages and 8Ω or 16Ω loads. The points at which 3% THD and 10% THD are reached are shown by the straight lines intersecting the curves. At 3% THD the output waveform has noticeable clipping while at 10% THD severe clipping of the output is occurring. It is also worth noting that the maximum amplifier power dissipation will typically be *double* the rated per channel *undistorted* power output into a resistive load. Since many of the smaller audio power amplifiers are rated at 10% THD, knowing that the output power at 10% is 30% larger than the undistorted power output enables a quick calculation to be made of the maximum amplifier dissipation,

$$P_{D(MAX)} \approx \frac{2 \times P_{MAX(RATE)}}{1.3}$$

or 1.5 times the output power at 10% THD.

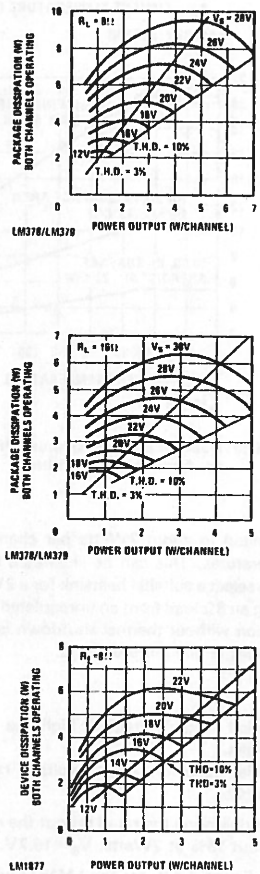


FIGURE 4.4.9 Device Dissipation for 8Ω and 16Ω Loads

Figure 4.4.10 gives the power derating curves for the dual amplifiers. Used in conjunction with Figure 4.4.9, the derating curves will indicate the heatsink requirements for continuous operation at any output power level and ambient temperature. It should be obvious from these curves that in most cases continuous or rms power at the rated output can require substantial heatsinking. Although the LM379 can be effectively heatsinked because of the low thermal resistance of the "S" Package style, with practical heatsinks the LM378 and

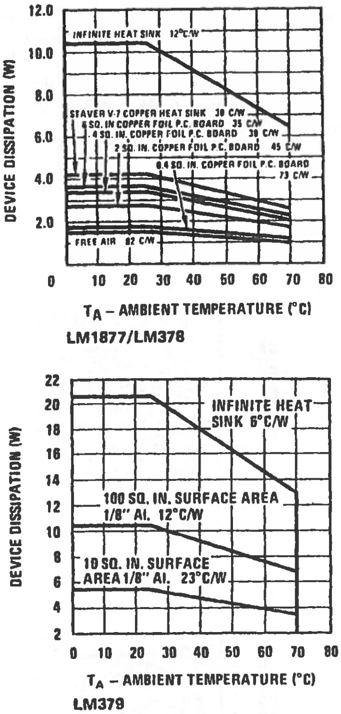


FIGURE 4.4.10 Dual Amplifier Maximum Dissipation vs. Ambient Temperature

LM1877 are limited to about 2Watts per channel output at elevated temperatures. This can be illustrated by the use of these curves to select a suitable heatsink for a 2Watt/Channel amplifier driving an 8Ω load from an unregulated 18Volt power supply. Operation without thermal shutdown is required at a maximum ambient temperature of 55°C.

Solution:

1. Unloaded supply voltage at high line = 18×1.1 = 19.8Volts.
   Amplifier maximum supply voltage rating must be ⩾ 20Volts.
2. When delivering the rated output the supply will sag by about 15% at 2Watts, $V_S = 15.7V$.
3. From Figure 4.4.9, for the LM1877 with this supply voltage, the amplifier can deliver 2Watts before clipping and 2.5Watts at 10% THD.
4. From the same curve the peak device power dissipation is 3.2Watts.
5. Figure 4.4.10 shows that for dissipating 3.2Watts at 55°C, a Staver V7-1 heatsink is needed.

From the above calculations there doesn't seem to be much point in publishing curves for an amplifier driving 8Ω with regulated supplies above 16 volts or unregulated supplies above 18 volts.

Also, a designer appears to be prevented from using higher supply voltages to provide a safety margin from clipping at rated outputs or power outputs in excess of 2.5Watts. This is not always true. Usually for speech or music there is a 30dB ratio between the R.M.S. and peak power levels. It is possible to design the heatsink for power levels 20dB below the rated maximum, anticipating that the heatsink thermal capacity is adequate to carry through peak power levels. In any case, the dual amplifiers have thermal shutdown circuitry to protect the device if sustained peak power levels cause the junction temperature to increase above 150°C.

Where higher power levels must be sustained, the alternative is to use the Single-in-line Package style (S.I.P.), Figure 4.4.11. The S.I.P. not only permits more compact p.c.b. layouts to be obtained, but the large tab allows easier heatsinking. In this package the LM2877 is electrically equivalent to the LM1877, and the LM2896 is the S.I.P. version of the LM1896. Figure 4.4.12 shows the substantially better thermal performance of the S.I.P. The power output levels of the previous example can be handled by less than a $2 \times 2 \times 1/16''$ piece of aluminum. If the LM2877 is bolted to a typical chassis, then 5.5watts can be dissipated at 55°C ambient temperature for output power levels in excess of 3W/Ch!

For the LM379S custom heatsinks are easily fabricated from sheet copper of aluminum and are bolted to the package tab. Power outputs of over 4W/Ch are possible, although the designer should watch out for the LM379 current limit specification. On the data sheet this is given as 1.5A measured at 25°C. As the I/C warms up, this current limit will decrease to between 1A and 1.25A. These peak currents correspond to 0.7 to 0.88 A_{RMS}, which will limit the output power into 8Ω to 4W or 6.2W respectively.

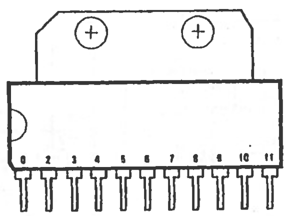


FIGURE 4.4.11 Molded Single-In-Line Package (NT)

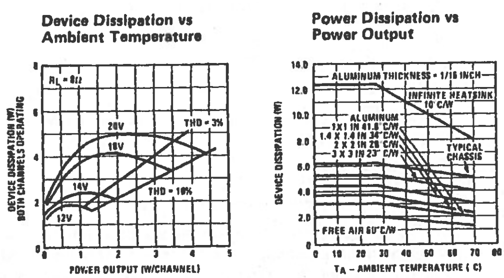


FIGURE 4.4.12 LM2887NT Power Dissipation and Temperature Derating Curves

4.4.8 Stabilization

The LM378/379 series amplifiers are internally stabilized so external compensation capacitors are not required. The high gain × BW provides a bandwidth greater than 50kHz, as seen in Figure 4.4.13. These amplifiers are, however, not intended for closed loop gain below 10. The typical Bode plot of Figure 4.4.14 shows a phase margin of 70° for gain of 5.6 (15dB), which is stable. At unity gain the phase margin is less than 30°, or marginally stable. This margin may vary considerably from device to device due to variation in gain × BW.

4.4.9 Layout

Ground and power connections must be adequate to handle the 1 to 2A peak supply and load currents. Ground loops can be especially troublesome because of these high currents. The load return line should be connected directly to the ground pins of the package on one side and/or the input and feedback ground lines should be connected directly to the ground pins (possibly on the other side of the package). The signal ground should not be connected so as to intercept any output signal voltage drop due to resistance between IC ground and load ground.

4.4.10 Split Supply Operation

The use of split power supplies offers a substantial reduction in parts count for low power stereo systems using dual power amplifiers. Split supply operation requires only redefinition of the ground pins for use with the negative power supply. The only precaution necessary is to observe that when thermal shutdown occurs the output is pulled down to the negative supply, instead of ground. Both supplies require bypassing with 0.1μF ceramic or 0.47μF mylar capacitors to ground.

Single supply operation (Figure 4.4.15) requires 6 resistors and 9 capacitors (excluding power supply parts) and uses the typical power supply shown. The same circuit using split supplies (Figure 4.4.16) requires only 4 resistors and 4 capacitors. This approach allows direct coupling of the amplifier to the speakers since the output DC level is approximately zero volts (offset voltages will be less than 25mV), thereby eliminating the need for large coupling capacitors and their associated degradation of power, distortion and cost. Since the input bias voltages are zero volts, the need for bias resistors and the bias-pin supply bypassing capacitor are also eliminated. Input capacitors are omitted to allow bias currents from the positive inputs to flow directly through the volume pots to ground.

Normally with split supply operation, the current loading of each supply is fairly symmetrical. Nevertheless, care should be taken that at turn-on both supplies increase from zero to full value at the same rate or within a couple volts of each other. If this is not so, referring the non-inverting inputs to ground instead of to pin 1 can cause a latch-up state to occur.

4.4.11 Unity Gain Power Buffers

Occasionally system requirements dictate the need for a unity gain power buffer, i.e., a current amplifier rather than a voltage amplifier. The peak output currents greater than one amp of the LM378/379 family make them a logical choice for this application.

Internal compensation limits stable operation to gains greater than 10 (20dB), thereby requiring additional components if unity gain operation is to be used. Stable unity gain inverting amplifiers (Figure 4.4.17) require only one additional resistor from the negative input to ground, equal in value to one tenth the feedback resistor. A discussion of this technique may be found in Section 2.8.4.

Non-inverting unity gain stability (Figure 4.4.18) can be achieved without additional components by judicious selection of the existing feedback elements. Writing the gain function of Figure 4.4.16 including the frequency dependent term of C_2 yields:

$$A_V = 1 + \frac{R_1}{R_2 + X_{C2}}$$

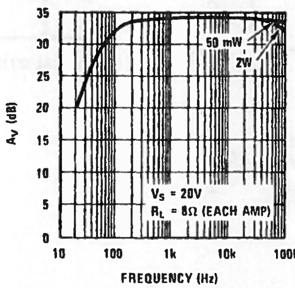


FIGURE 4.4.13 Frequency Response of the Stereo Amp of Figure 4.4.5

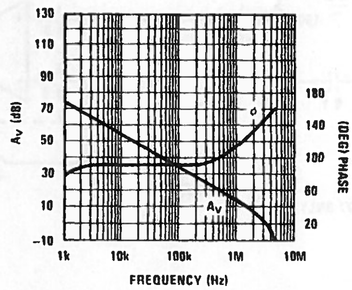


FIGURE 4.4.14 Open Loop Bode Plot (Approximately Worst Case)

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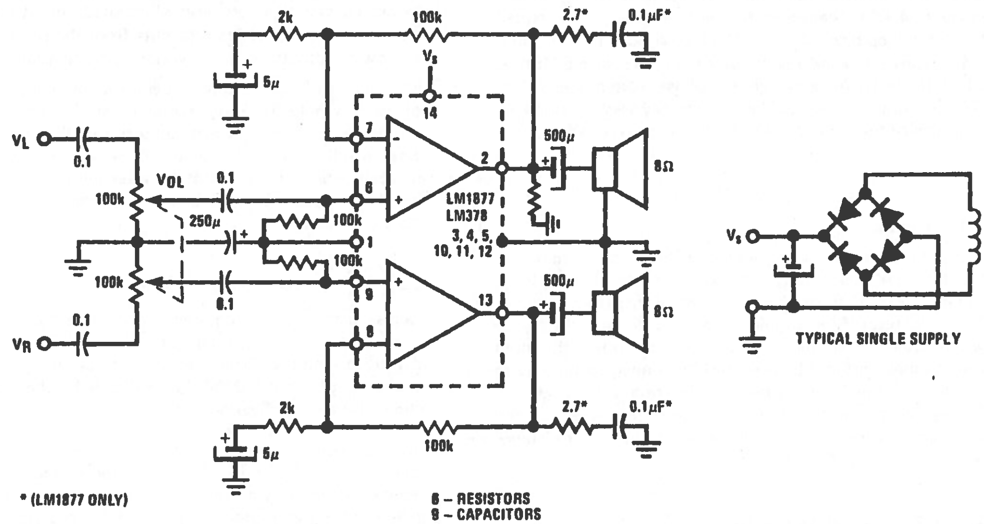


FIGURE 4.4.15 Non-Inverting Amplifier Using Single Supply

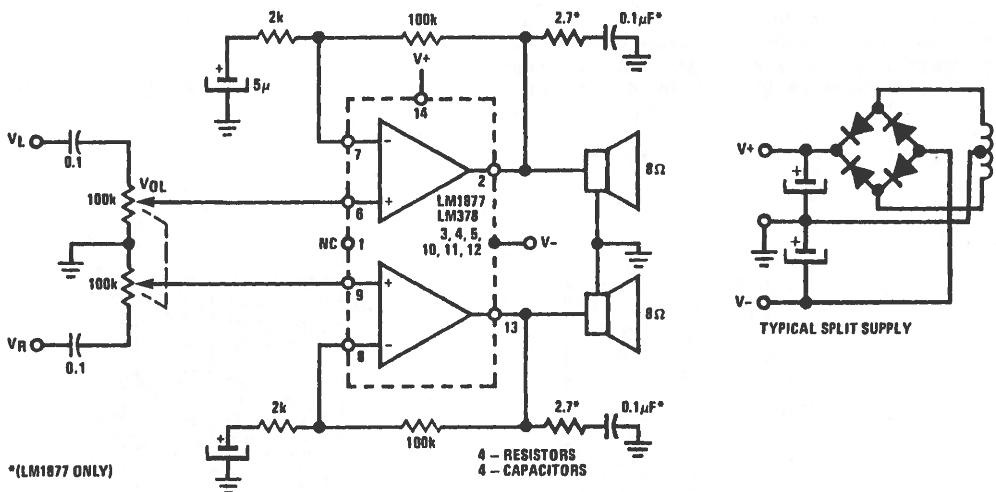


FIGURE 4.4.16 Non-Inverting Amplifier Using Split Supply

Satisfaction of unity gain *circuit* performance over the audio band and gain greater than 10 *amplifier* performance at high frequencies can be accomplished by making the frequency dependent term small (relative to one) over the audio band and allowing it to dominate the gain expression beyond audio. Rewriting the gain term using the Laplace variable S (The variable S is a complex frequency.) results in Equation (4.4.1):

$$A_V = 1 + \frac{R_1}{R_2 + \frac{1}{S C_2}} = \frac{S(R_1 + R_2) C_2 + 1}{S R_2 C_2 + 1}$$

$$\approx \frac{S R_1 C_2 + 1}{S R_2 C_2 + 1} \qquad (4.4.1)$$

$$\text{Zero at } f_z = \frac{1}{2 \pi R_1 C_2} \qquad (4.4.2)$$

$$\text{Pole at } f_p = \frac{1}{2 \pi R_2 C_2} \qquad (4.4.3)$$

Examination of Equation (4.4.1) shows it to have a frequency response zero at f_z (Equation (4.4.2)) and a pole at f_p (Equation (4.4.3)). By selecting f_z to fall at the edge of the audio spectrum (20 kHz as shown) and f_p prior to hitting the open loop response (340 kHz as shown) the frequency response of Figure 4.4.19 is obtained. This response satisfies the unity gain requirements, while allowing the gain to raise beyond audio to insure stable operation.

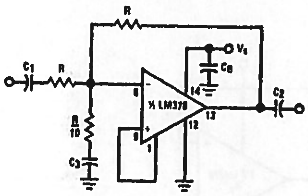


FIGURE 4.4.17 Inverting Unity Gain Amplifier

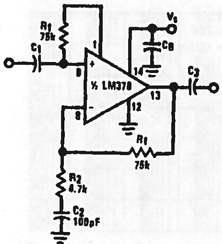


FIGURE 4.4.18 Non-Inverting Unity Gain Amplifier

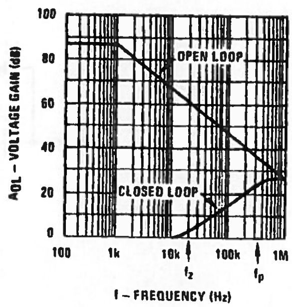


FIGURE 4.4.19 Frequency Response of Non-Inverting Unity Gain Amplifier

4.4.12 Bridge Amplifiers

The dual amplifiers are equally useful in the bridge configuration to drive floating loads, which may be loudspeakers, servo motors or whatever. Double the power output can be obtained in this connection, and output coupling capacitors are not required. Load impedance may be either 8 or 16Ω in the bridge circuit of Figure 4.4.20. Response of this circuit is 20 Hz to 160 kHz as shown in Figure 4.4.21 and distortion is 0.1% midband at 4W, rising to 0.5% at 10 kHz and 50 mW output (Figure 4.4.22). The higher distortion at low power is due to a small amount of crossover notch distortion which becomes more apparent at low powers and high frequencies. The circuit of Figure 4.4.23 is similar except for higher input impedance. In Figure 4.4.23 the signal drive for the inverting amplifier is derived from the feedback voltage of the non-inverting amplifier. Resistors R_1 and R_3 are the input and feedback resistors for A_2, whereas R_1 and R_2 are the feedback network for A_1. So far as A_1 is concerned, R_2 sees a virtual ground at the (−) input to A_2; therefore, the gain of A_1 is $(1 + R_2/R_1)$. So far as A_2 is concerned, its input signal is the voltage appearing at the (−) input to A_1. This equals that at the (+) input to A_1. The driving point impedance at the (−) input to A_1 is very low even though R_2 is 100k. A_1 can be considered a unity gain amplifier with internal $R = R_2 = 100k$ and $R_L = R_1 = 2k$. Then the effective output resistance of the unity gain amplifier is:

$$R_{OUT} = \frac{R_{INTERNAL}}{A_{OL}/A_\beta} = \frac{100k}{600/1} = 167\Omega$$

Layout is critical if output oscillation is to be avoided. Even with careful layout, capacitors C_1 and C_2 may be required to prevent oscillation. With the values shown, the amplifier will drive a 16Ω load to 4W with less than 0.2% distortion midband, rising to 1% at 20 kHz (Figure 4.4.24). Frequency response is 27 Hz to 60 kHz as shown in Figure 4.4.25. The low frequency roll off is due to the double poles $C_3 R_3$ and $C_4 R_1$.

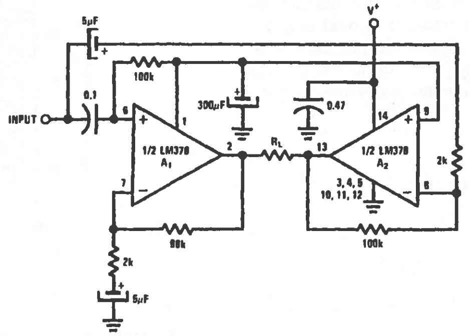


FIGURE 4.4.20 4-Watt Bridge Amplifier

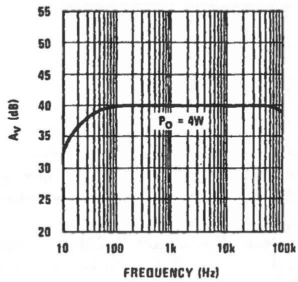


FIGURE 4.4.21 Frequency Response, Bridge Amp of Figure 4.4.20

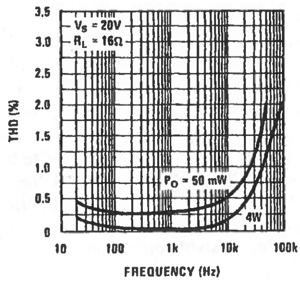


FIGURE 4.4.22 Distoration for Bridge Amp of Figure 4.4.20

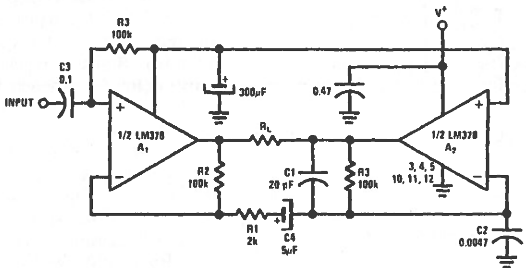


FIGURE 4.4.23 4-Watt Bridge Amplifier with High Input Impedance

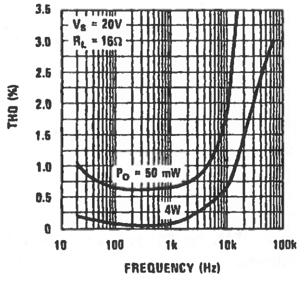


FIGURE 4.4.24 Distoration for Bridge Amp of Figure 4.4.23

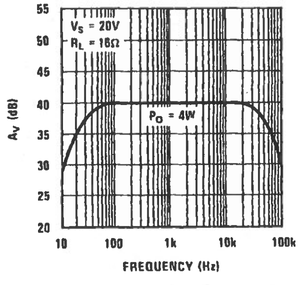


FIGURE 4.4.25 Frequency Response, Bridge Amp of Figure 4.4.23

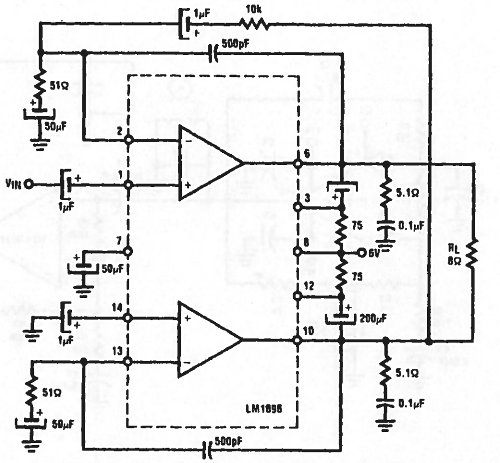


FIGURE 4.4.26 2-Watt Bridge Amplifier with 6V Supply

4.4.13 Power Oscillator

One half of an LM378 may be connected as an oscillator to deliver up to 2W to a load. Figure 4.4.27 shows a Wien bridge type of oscillator with FET amplitude stabilization in the negative feedback path. The circuit employs internal biasing and operates from a single supply. C_3 and C_6 allow unity gain DC feedback and isolate the bias from ground. Total harmonic distortion is under 1% to 10kHz, and could possibly be improved with careful adjustment of R_5. The FET acts as the variable element in the feedback attenuator R_4 to R_6. Minimum negative feedback gain is set by the resistors R_4 to R_6, while the FET shunts R_6 to increase gain in the absence of adequate output signal. The peak detector D_2 and C_8 senses output level to apply control bias to the FET. Zener diode D_1 sets the output level although adjustment could be made if R_9 were a potentiometer with R_8 connected to the slider. Maximum output level with the values shown is $5.3V_{RMS}$ at 60 Hz. C_7 and the attenuator R_7 and R_8 couple 1/2 the signal of the FET drain to the gate for improved FET linearity and low distortion. The amplitude control loop could be replaced by an incandescent lamp in non-critical circuits (Figure 4.4.25), although DC offset will suffer by a factor of about 3 (DC gain of the oscillator). R_{10} matches R_3 for improved DC stability, and the network R_{11}, C_9 increases high frequency gain for improved stability. Without this RC, oscillation may occur on the negative half cycle of output waveform. A low inductance capacitor, C_5, located directly at the supply leads on the package is important to maintain stability and prevent high frequency oscillation on negative half cycle of the output waveform. C_5 may be 0.1µF ceramic, or 0.47µF mylar. Layout is important; especially take care to avoid ground loops as discussed in the section on amplifiers. If high frequency instability still occurs, add the R_{12}, C_{10} network to the output.

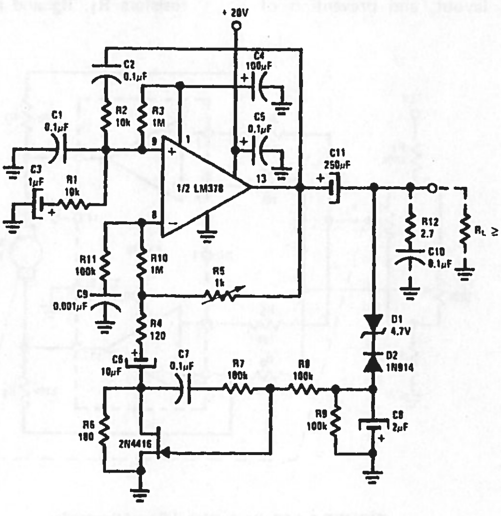


FIGURE 4.4.27 Wien Bridge Power Oscillator

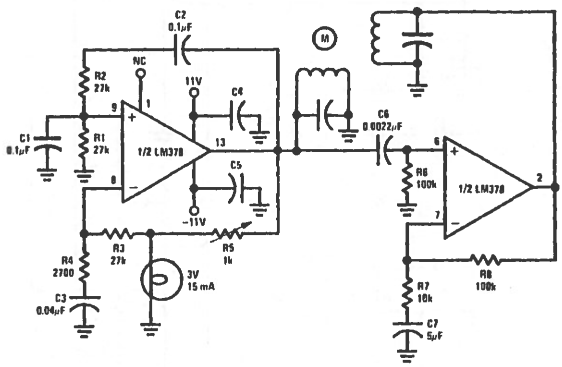


FIGURE 4.4.28 Two-Phase Motor Drive

4.4.14 Two-Phase Motor Drive

Figure 4.4.28 shows the use of the LM378 to drive a small 60Hz two phase servo motor up to 3W per phase. Applications such as a constant (or selectable) speed phonograph turntable drive are adequately met by this circuit. A split supply is used to simplify the circuit, reduce parts count, and eliminate several large bypass capacitors. An incandescent lamp is used in a simple amplitude stabilization loop. Input DC is minimized by balancing DC resistance at (+) and (−) amplifier inputs ($R_1 = R_3$ and $R_6 = R_8$). High frequency stability is assured by increasing closed-loop gain from approximately 3 at 60Hz to about 30 above 40kHz with the network consisting of R_3, R_4 and C_3. The inter-stage coupling $C_6 R_6$ network shifts phase by 85° at 60Hz to provide the necessary two phase motor drive signal. The gain of the phase shift network is purposely low so that the buffer amplifier will operate at a gain of 10 for adequate high frequency stability. As in other circuits, the importance of supply bypassing, careful layout, and prevention of output ground loops is to be stressed. The motor windings are tuned to 60Hz with shunt capacitors. This circuit will drive 8Ω loads to 3W each.

4.4.15 Proportional Speed Controller

A low cost proportional speed controller may be simply designed using a LM378 amplifier. For use with 12-24V_{DC} motors at continuous currents up to several hundred milliamps, this circuit allows remote adjustment of angular displacements in a drive shaft. Typical applications include rooftop rotary antennas and motor-controlled valves.

Proportional control (Figure 4.4.29) results from an error signal developed across the Wheatstone bridge comprised of resistors R_1, R_2 and potentiometers P_1,P_2. Control P_1 is

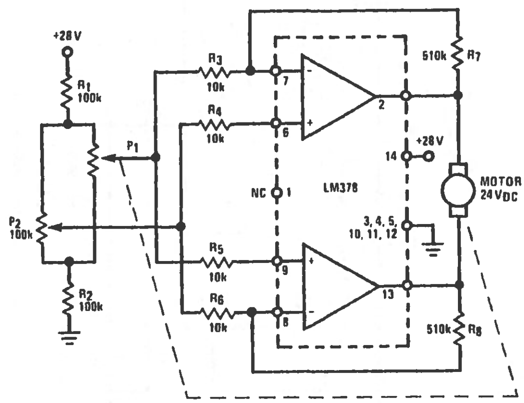


FIGURE 4.4.29 Proportional Speed Controller

mechanically coupled to the motor shaft as depicted by the dotted line and acts as a continuously variable feedback sensor. Setting position control P_2 creates an error voltage between the two inputs which is amplified by the LM378 (wired as a difference bridge amplifier); the magnitude and polarity of the output signal of the LM378 determines the speed and direction of the motor. As the motor turns, potentiometer P_1 tracks the movement, and the error signal, i.e., difference in positions between P_1 and P_2, becomes smaller and smaller until ultimately the system stops when the error voltage reaches zero volts.

Actual gain requirements of the system are determined by the motor selected and the required range. Figure 4.4.26 demonstrates the principle involved in proportional speed control and is not intended to specify final resistor values.

4.4.16 Complete Systems

The dual power amplifiers are useful in table or console radios, phonographs, tape players, intercoms, or any low to medium power music systems.

Figures 4.4.30 through 4.4.32 describe the complete electronic section of a 2-channel sound system with inputs for AM radio, stereo FM radio, phono, and tape playback. Figure 4.4.30 combines the power amplifier pair with loudness, balance, and tone controls. The tone controls allow boost or cut of bass and/or treble. Transistors Q_1 and Q_2 act as input line amplifiers with the triple function of (1) presenting a high impedance to the inputs, especially ceramic phono; (2) providing an amplified output signal to a tape recorder; and (3) providing gain to make up for the loss in the tone controls. Feedback tone controls of the Baxandall type employing transistor gain could be used; but then, with the same transistor count, the first two listed functions of Q_1Q_2 would be lost. It is believed that this circuit represents the lowest parts count for the complete system. Figure 4.4.31 is the additional circuitry for input switching and tape playback amplifiers. The LM382 with capacitors as shown provides for NAB tape playback compensation. For further information on the LM382 or the similar LM381 and LM387, refer to Section 2.0

Figure 4.4.32 shows the relationship between signal source impedance and gain or input impedance for the amplifier stage Q_1Q_2. Stage gain may be set at a desired value by choice of either the source impedance or insertion of resistors in series with the inputs (as R_1 to R_4 in Figure 4.4.31). Gain is variable from −15 to +24 dB by choice of series R from 0 to 10 meg. Gain required for $e_{IN} = 100$ to 200 mV (approximate value of recovered audio from FM stereo or AM radio) is about 18 to 21 dB overall for 2W into an 8Ω speaker at 1 Hz or 21 to 24 dB for 4W.

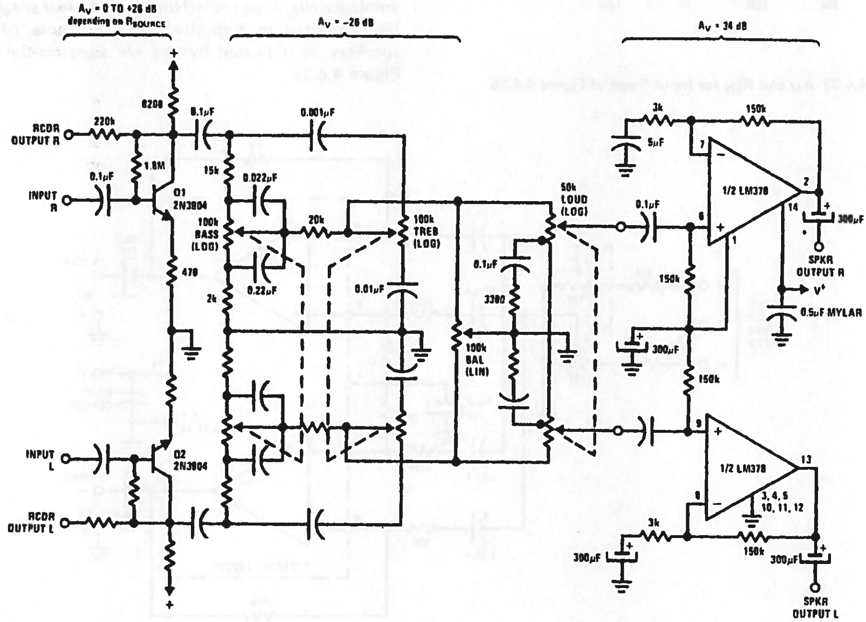


FIGURE 4.4.30 Two-Channel Power Amplifier and Control Circuits

4

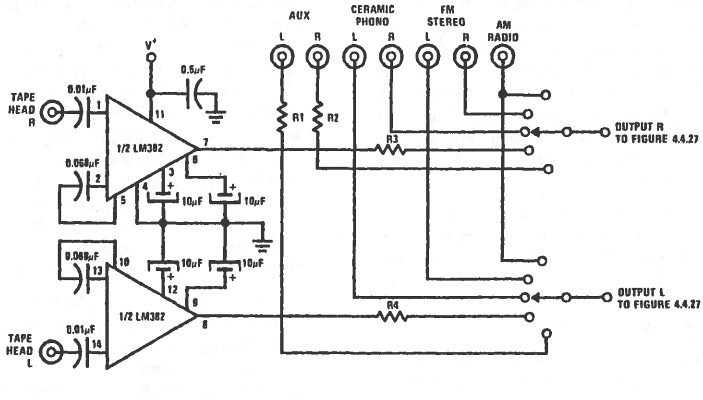


FIGURE 4.4.31 Two-Channel Tape-Playback Amplifier and Signal Switching

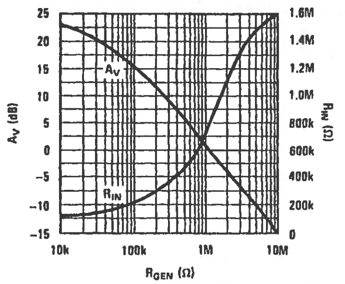


FIGURE 4.4.32 A_V and R_{IN} for Input Stage of Figure 4.4.26

4.4.17 Rear Channel Ambience Amplifier

The rear channel "ambience" circuit of Figure 4.4.33 can be added to an existing stereo system to extract a difference signal (R − L or L − R) which, when combined with some direct signal (R or L), adds some fullness, or "concert hall realism" to reproduction of recorded music. Very little power is required at the rear channels, hence an LM1877 will suffice for most "ambience" applications. The inputs are merely connected to the existing speaker output terminals of a stereo set, and two more speakers are connected to the ambience circuit outputs. Note that the rear speakers should be connected in opposite phase to those of the front speakers, as indicated by the +/− signs on the diagram of Figure 4.4.33.

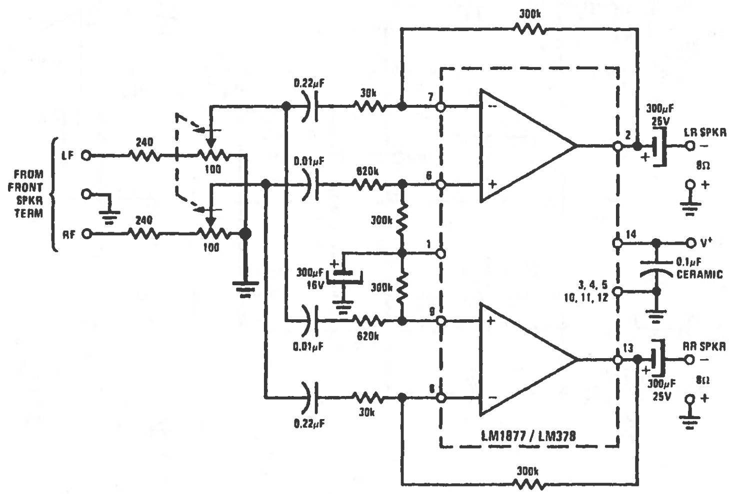


FIGURE 4.4.33 Rear Speaker Ambience (4-Channel) Amplifier

4.4.18 Ceramic Cartridge Stereo Phonograph

Ceramic cartridges, with a high output level of several hundred millivolts, can be used with the LM1877 as the only active gain element to provide a complete and inexpensive 2W/Ch stereo phonograph system. A suitable circuit is shown in Figure 4.4.34 where the cartridge is loaded directly with the 500kΩ gain control potentiometers. The LM1877 is configured in the non-inverting mode to minimize loading on the cartridge at maximum volume settings and a simple bass tone control circuit is added in the feedback network (see Section 2.14.7). Response of the tone control circuit is shown in Figure 4.4.35. At midband and higher frequencies the capacitors can be considered as short circuits, which gives a midband gain

$$A_V = \left(\frac{10k + 1k}{1k}\right) \times \left(\frac{510k + 51k}{51k}\right) = 121 \text{ or } 42\text{dB}.$$

For a typical ceramic cartridge output of 200mV to 300mV, this gain is more than adequate to ensure clipping at the speaker output with moderate gain control settings. The amplifier is capable of delivering 2 watts continuously in both channels at the 10% distortion level into 8Ω loads on a 14V supply. With a 16V supply, 2.5 watts continuous is available (See Figure 4.4.9).

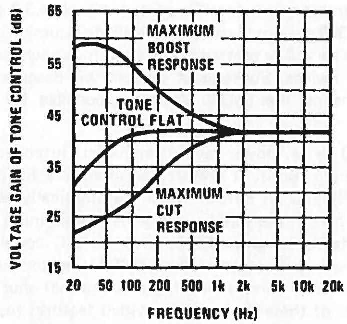


FIGURE 4.4.35 Frequency Response of Bass Tone Control

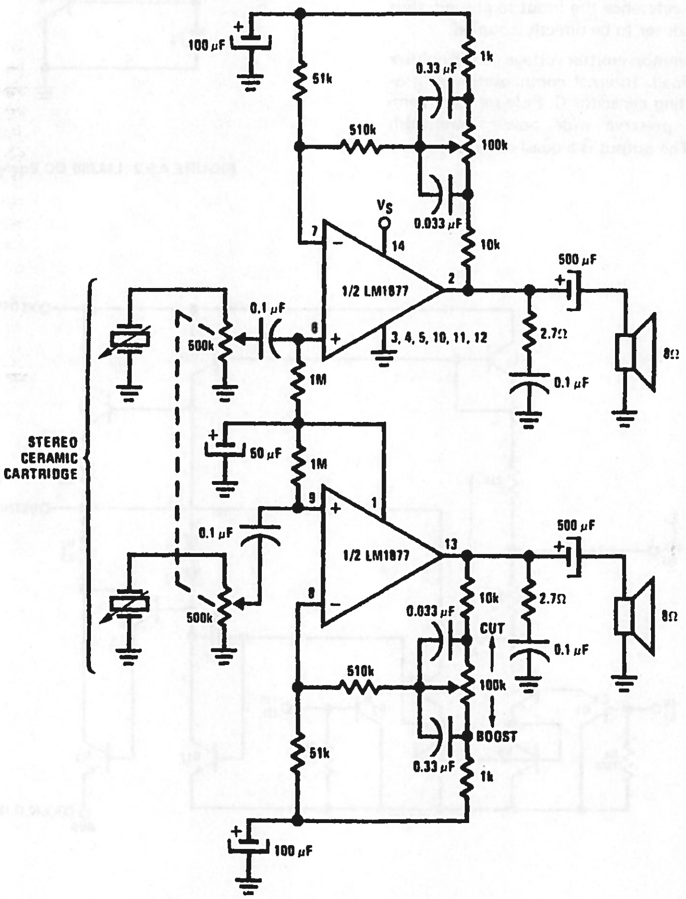


FIGURE 4.4.34 Stereo Phongraph Amplifier with Bass Tone Control

4.5 LM380 AUDIO POWER AMPLIFIER

4.5.1 Introduction

Most of the mono power amplifiers listed in Table 4.3.3 derive from the LM380 design; therefore, a detailed discussion of the internal circuitry will be presented as a basis for understanding each of the devices. Subsequent sections will describe only the variations on the LM380 design responsible for each unique part.

The LM380 is a power audio amplifier intended for consumer applications. It features an internally fixed gain of 50 (34dB) and an output which automatically centers itself at one half of the supply voltage. A unique input stage allows inputs to be ground referenced or AC coupled as required. The output stage of the LM380 is protected with both short circuit current limiting and thermal shutdown circuitry. All of these internally provided features result in a minimum external parts count integrated circuit for audio applications.

4.5.2 Circuit Description

Figure 4.5.1 shows a simplified circuit schematic of the LM380. The input stage is a PNP emitter-follower driving a PNP differential pair with a slave current-source load. The PNP input is chosen to reference the input to ground, thus enabling the input transducer to be directly coupled.

The second stage is a common emitter voltage gain amplifier with a current-source load. Internal compensation is provided by the pole-splitting capacitor C. Pole-splitting compensation is used to preserve wide power bandwidth (100kHz at 2W, 8Ω). The output is a quasi-complementary pair emitter-follower.

The output is biased to half the supply voltage by resistor ratio R_2/R_1. Simplifying Figure 4.5.1 still further to show the DC biasing of the output stage results in Figure 4.5.2, where resistors R_1 and R_2 are labeled R. Since the transistor operates with effectively zero volts base to collector, the circuit acts as a DC amplifier with a gain of one half (i.e., $A_V = R/[R + R]$) and an input of V^+; therefore, the output equals $V^+/2$.

The amplifier AC gain is internally fixed to 34dB (or 50V/V). Figure 4.5.3 shows this to be accomplished by the internal feedback network R_2-R_3. The gain is twice that of the ratio R_2/R_3 due to the slave current-source (Q_5, Q_6) which provides the full differential gain of the input stage.

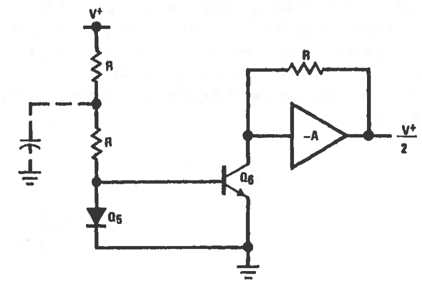


FIGURE 4.5.2 LM380 DC Equivalent Circuit

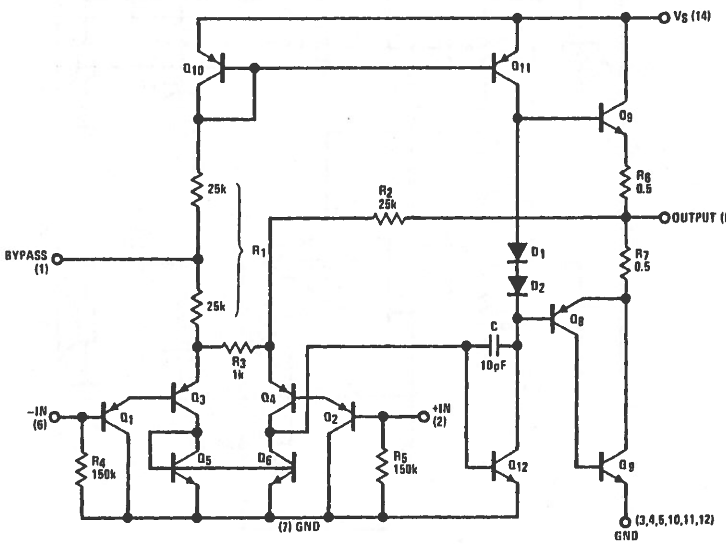


FIGURE 4.5.1 LM380 Simplified Schematic

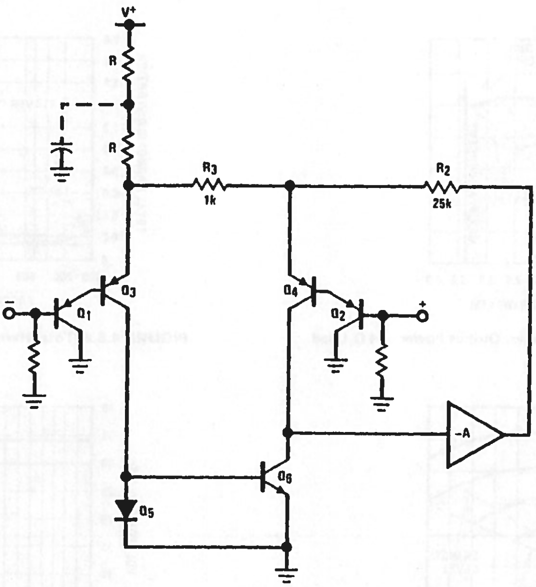


FIGURE 4.5.3 LM380 AC Equivalent Circuit

A gain difference of one exists between the negative and positive inputs, analogous to inverting and non-inverting amplifiers. For example, an inverting amplifier with input resistor equal to 1k and a 50k feedback resistor has a gain of 50V/V, while a non-inverting amplifier constructed from the same resistors has a gain of 51V/V. Driving the inverting terminal of the LM380, therefore, results in a gain of 50, while driving the non-inverting will give a gain of 51.

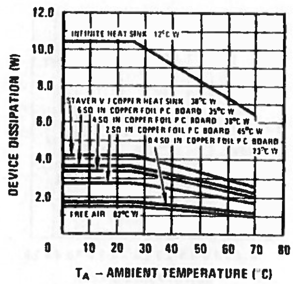


FIGURE 4.5.4a Device Dissipation vs. Maximum Ambient Temperature

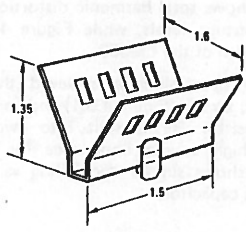
4.5.3 General Operating Characteristics

The output current of the LM380 is rated at 1.3A peak. The 14 pin dual-in-line package is rated at 35°C/W when soldered into a printed circuit board with 6 square inches of 2 ounce copper foil (Figure 4.5.4). Since the device junction temperature is limited to 150°C via the thermal shutdown circuitry, the package will support 2.9W dissipation at 50°C ambient or 3.6W at 25°C ambient.

Figure 4.5.4a shows the maximum package dissipation vs. ambient temperature for various amounts of heat sinking. (Dimensions of the Staver V7 heat sink appear as Figure 4.5.4b.)

Figures 4.5.5a, -b, and -c show device dissipation versus output power for various supply voltages and loads.

The maximum device dissipation is obtained from Figure 4.5.4 for the heat sink and ambient temperature conditions under which the device will be operating. With this maximum allowed dissipation, Figures 4.5.5a, -b, and -c show the maximum power supply allowed (to stay within dissipation limits) and the output power delivered into 4, 8 or 16Ω loads. The three percent total harmonic distortion line is approximately the onset of clipping.



*—Staver Co.
Bayshore, N.Y.

FIGURE 4.5.4b Staver* "V7" Heat Sink

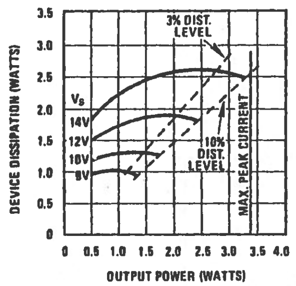


FIGURE 4.5.5a Device Dissipation vs. Output Power — 4 Ω Load

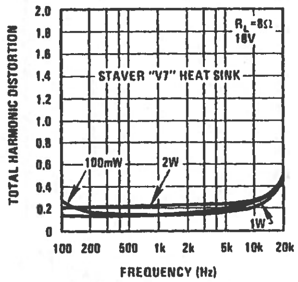


FIGURE 4.5.6 Total Harmonic Distortion vs. Frequency

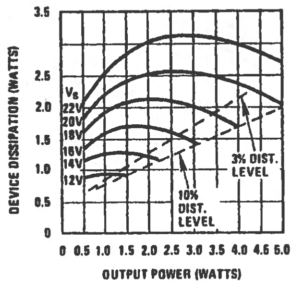


FIGURE 4.5.5b Device Dissipation vs. Output Power — 8 Ω Load

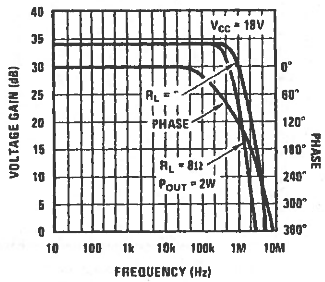


FIGURE 4.5.7 Output Voltage Gain vs. Frequency

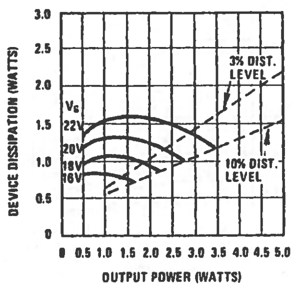


FIGURE 4.5.5c Device Dissipation vs. Output Power — 16 Ω Load

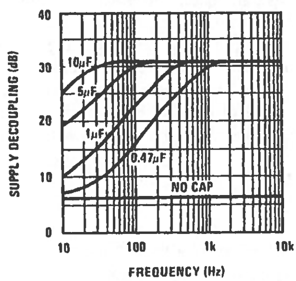


FIGURE 4.5.8 Supply Decoupling vs. Frequency

Figure 4.5.6 shows total harmonic distortion vs. frequency for various output levels, while Figure 4.5.7 shows the power bandwidth of the LM380.

Power supply decoupling is achieved through the AC divider formed by R_1 (Figure 4.5.1) and an external bypass capacitor. Resistor R_1 is split into two 25kΩ halves providing a high source impedance for the integrator. Figure 4.5.8 shows supply decoupling vs. frequency for various bypass capacitors.

4.5.4 Biasing

The simplified schematic of Figure 4.5.1 shows that the LM380 is internally biased with the 150kΩ resistance to ground. This enables input transducers which are referenced to ground to be direct-coupled to either the inverting or non-inverting inputs of the amplifier. The unused input may be either: (1) left floating, (2) returned to ground through a resistor or capacitor, or (3) shorted to ground. In most applications where the non-inverting input is used, the inverting input is left floating. When the inverting input is used and the non-inverting input is left floating, the amplifier may be found to be sensitive to board layout since stray coupling to the floating input is positive feedback. This can be avoided by employing one of three alternatives: (1) AC grounding the unused input with a small capacitor. This is preferred when using high source impedance transducers. (2) Returning the unused input to ground through a resistor. This is preferred when using moderate to low DC source impedance transducers and when output offset from half supply voltage is critical. The resistor is made equal to the resistance of the input transducer, thus maintaining balance in the input differential amplifier and minimizing output offset. (3) Shorting the unused input to ground. This is used with low DC source impedance transducers or when output offset voltage is non-critical.

4.5.5 Oscillation

The normal power supply decoupling precautions should be taken when installing the LM380. If V_S is more than 2" to 3" from the power supply filter capacitor it should be decoupled with a $0.1\mu F$ disc ceramic capacitor at the V_S terminal of the IC.

The R_C and C_C components in Figure 4.5.9 and throughout this section suppress a 5 to 10MHz small amplitude oscillation which can occur during the negative swing into a load which draws high current. The oscillation is of course at too high a frequency to pass through a speaker, but it should be guarded against when operating in an RF sensitive environment.

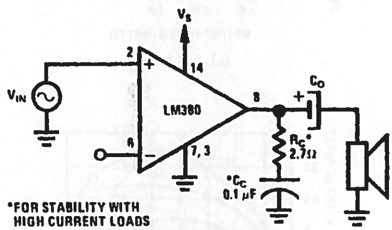


FIGURE 4.5.9 Oscillation Suppression Components

4.5.6 RF Precautions — See Section 2.3.10

4.5.7 Inverting Amplifier Application

With the internal biasing and compensation of the LM380, the simplest and most basic circuit configuration requires only an output coupling capacitor as seen in Figure 4.5.10.

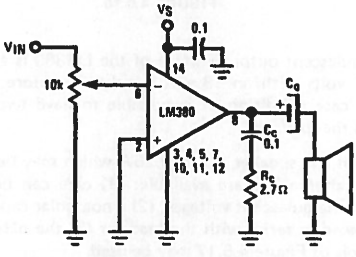


FIGURE 4.5.10 Minimum Component Configuration

4.5.8 Ceramic Phono Amplifier

An application of this basic configuration is the phonograph amplifier where the addition of volume and tone controls is required. Figure 4.5.11 shows the LM380 with a voltage divider volume control and high frequency roll-off tone control.

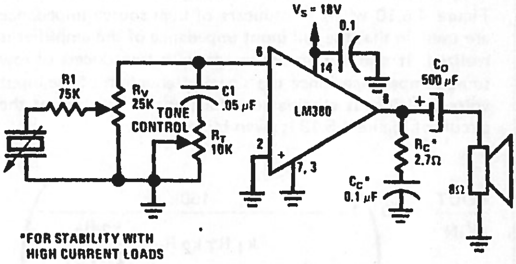


FIGURE 4.5.11 Ceramic Phono Amp

4.5.9 Common Mode Volume and Tone Controls

When maximum input impedance is required or the signal attenuation of the voltage divider volume control is undesirable, a "common mode" volume control may be used as seen in Figure 4.5.12.

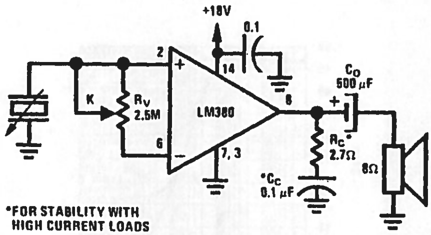
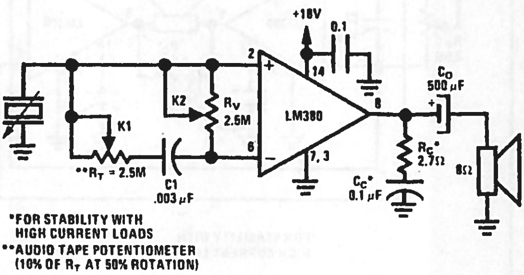


FIGURE 4.5.12 "Common Mode" Volume Control

With this volume control the source loading impedance is only the input impedance of the amplifier when in the full-volume position. This reduces to one half the amplifier input impedance at the zero volume position. Equation (4.5.1) describes the output voltage as a function of the potentiometer setting.

$$V_{OUT} = 50\,V_{IN}\left(1 - \frac{150 \times 10^3}{k_1 R_V + 150 \times 10^3}\right) \qquad 0 \leqslant k_1 \leqslant 1 \tag{4.5.1}$$

This "common mode" volume control can be combined with a "common mode" tone control as seen in figure 4.5.13.



*FOR STABILITY WITH
HIGH CURRENT LOADS
**AUDIO TAPE POTENTIOMETER
(10% OF R_T AT 50% ROTATION)

FIGURE 4.5.13 "Common Mode" Volume and Tone Control

This circuit has a distinct advantage over the circuit of Figure 4.5.10 when transducers of high source impedance are used, in that the full input impedance of the amplifier is realized. It also has an advantage with transducers of low source impedance, since the signal attenuation of the input voltage divider is eliminated. The transfer function of the circuit of Figure 4.5.13 is given by:

$$\frac{V_{OUT}}{V_{IN}} = 50\left(1 - \frac{150k}{150k + \dfrac{k_1 R_T k_2 R_V + \dfrac{k_2 R_V}{j2\pi f C_1}}{k_1 R_T + k_2 R_V + \dfrac{1}{j2\pi f C_1}}}\right)$$

$$0 \leqslant K_1 \leqslant 1$$
$$0 \leqslant K_2 \leqslant 1$$

(4.5.2)

Figure 4.5.14 shows the response of the circuit of Figure 4.5.13.

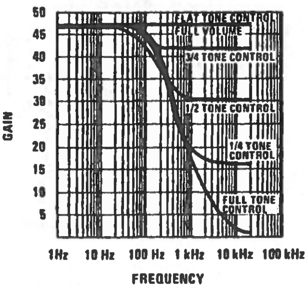
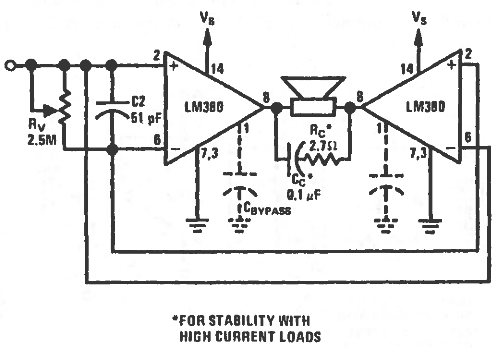


FIGURE 4.5.14 Tone Control Response

4.5.10 Bridge Amplifier

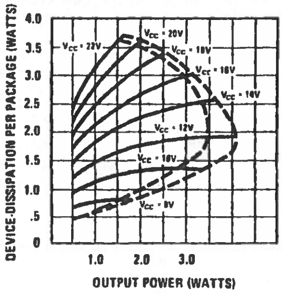
Where more power is desired than can be provided with one amplifier, two amps may be used in the bridge configuration shown in Figure 4.5.15.



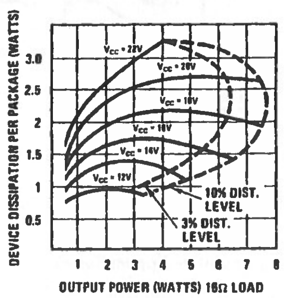
*FOR STABILITY WITH HIGH CURRENT LOADS

FIGURE 4.5.15 Bridge Configuration

This provides twice the voltage swing across the load for a given supply, thereby increasing the power capability by a factor of four over the single amplifier. However, in most cases the package dissipation will be the first parameter limiting power delivered to the load. When this is the case, the power capability of the bridge will be only twice that of the single amplifier. Figures 4.5.16a and -b show output power vs. device package dissipation for both 8 and 16Ω loads in the bridge configuration. The 3% and 10% harmonic distortion contours double back due to the thermal limiting of the LM380. Different amounts of heat sinking will change the point at which the distortion contours bend.



(a) 8Ω Load



(b) 16Ω Load

FIGURE 4.5.16

The quiescent output voltage of the LM380 is specified at 9 ± 1 volts with an 18 volt supply. Therefore, under the worst case condition, it is possible to have two volts DC across the load.

With an 8Ω speaker this is 0.25A which may be excessive. Three alternatives are available: (1) care can be taken to match the quiescent voltages, (2) a non-polar capacitor may be placed in series with the load, or (3) the offset balance controls of Figure 4.5.17 may be used.

The circuits of Figures 4.5.15 and 4.5.17 employ the "common mode" volume control as shown before. However, any of the various input connection schemes discussed previously may be used. Figure 4.5.18 shows the bridge configuration with the voltage divider input. As discussed in the "Biasing" section the undriven input may be AC or DC grounded. If V_S is an appreciable distance from the power supply (> 3") filter capacitor it should be decoupled with a 1 μF tantalum capacitor.

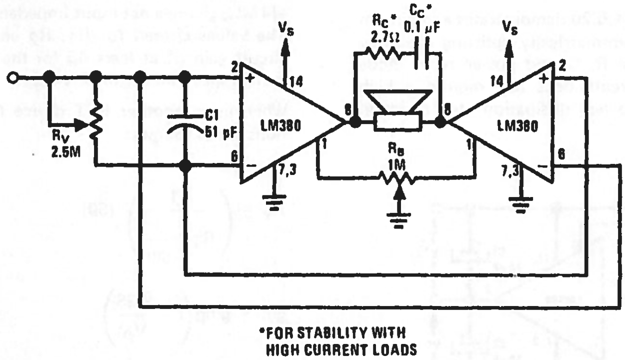


FIGURE 4.5.17 Quiescent Balance Control

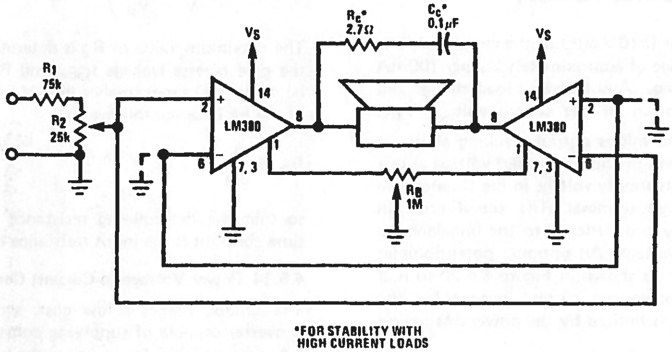


FIGURE 4.5.18 Voltage Divider Input

4.5.11 Intercom

The circuit of Figure 4.5.19 provides a minimum component intercom. With switch S_1 in the talk position, the speaker of the master station acts as the microphone with the aid of step-up transformer T_1.

A turns ratio of 25 and a device gain of 50 allows a maximum loop gain of 1250. R_V provides a "common mode" volume control. Switching S_1 to the listen position reverses the role of the master and remote speakers.

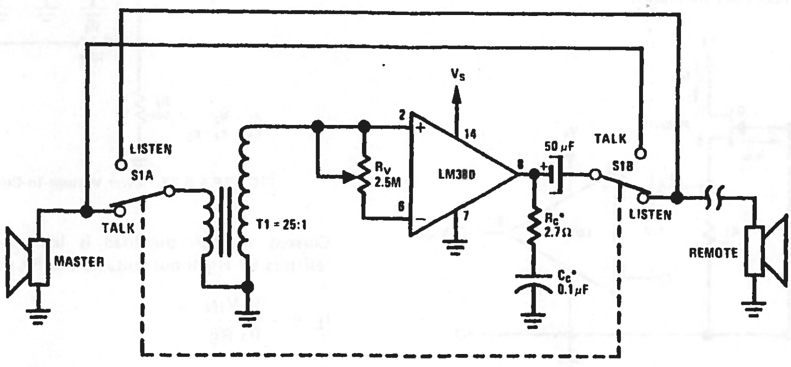


FIGURE 4.5.19 Intercom

4.5.12 Low Cost Dual Supply

The circuit shown in Figure 4.5.20 demonstrates a minimum parts count method of symmetrically splitting a supply voltage. Unlike the normal R, C, and power zener diode technique the LM380 circuit does not require a high standby current and power dissipation to maintain regulation.

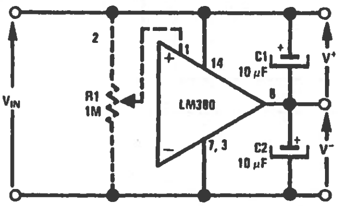


FIGURE 4.5.20 Dual Supply

With a 20V input voltage (±10V output) the circuit exhibits a change in output voltage of approximately 2% per 100mA of unbalanced load change. Any balanced load change will reflect only the regulation of the source voltage V_{IN}.

The theoretical plus and minus output tracking ability is 100% since the device will provide an output voltage at one half of the instantaneous supply voltage in the absence of a capacitor on the bypass terminal. The actual error in tracking will be directly proportional to the imbalance in the quiescent output voltage. An optional potentiometer may be placed at pin 1 as shown in Figure 4.5.20 to null output offset. The unbalanced current output for the circuit of Figure 4.5.20 is limited by the power dissipation of the package.

In the case of sustained unbalanced excess loads, the device will go into thermal limiting as the temperature sensing circuit begins to function. For instantaneous high current loads or short circuits the device limits the output current to approximately 1.3A until thermal shutdown takes over or until the fault is removed.

4.5.13 High Input Impedance Circuit

The junction FET isolation circuit shown in Figure 4.5.21 raises the input impedance to 22MΩ for low frequency input signals. The gate to drain capacitance (2pF maximum for the PN4221 shown) of the FET limits the input impedance as frequency increases.

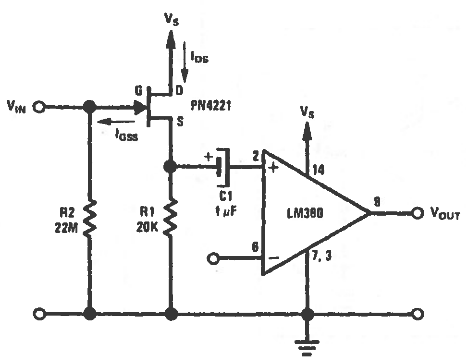


FIGURE 4.5.21 High Input Impedance

At 20kHz the reactance of this capacitor is approximately −j4MΩ, giving a net input impedance magnitude of 3.9MΩ. The values chosen for R_1, R_2 and C_1 provide an overall circuit gain of at least 45 for the complete range of parameters specified for the PN4221.

When using another FET device the relevant design equations are as follows:

$$A_V = \left(\frac{R_1}{R_1 + \frac{1}{gm}}\right)(50) \tag{4.5.3}$$

$$gm = gm_0\left(1 - \frac{V_{GS}}{V_P}\right) \tag{4.5.4}$$

$$V_{GS} = I_{DS} R_1 \tag{4.5.5}$$

$$I_{DS} = I_{DSS}\left(1 - \frac{V_{GS}}{V_P}\right)^2 \tag{4.5.6}$$

The maximum value of R_2 is determined by the product of the gate reverse leakage I_{GSS} and R_2. This voltage should be 10 to 100 times smaller than V_P. The output impedance of the FET source follower is:

$$R_o = \frac{1}{gm} \tag{4.5.7}$$

so that the determining resistance for the interstage RC time constant is the input resistance of the LM380.

4.5.14 Power Voltage-to-Current Converter

The LM380 makes a low cost, simple voltage-to-current converter capable of supplying constant AC currents up to 1A over variable loads using the circuit shown in Figure 4.5.22.

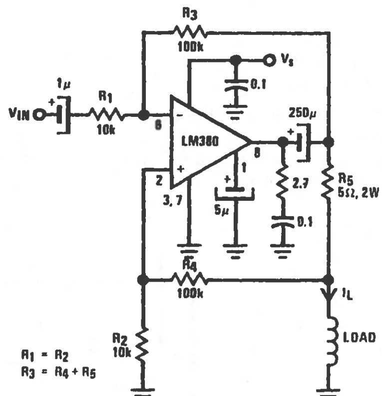


FIGURE 4.5.22 Power Voltage-to-Current Converter

Current through the load is fixed by the gain setting resistors R_1-R_3, input voltage, and R_5 per Equation (4.5.8).

$$I_L = -\frac{R_3 V_{IN}}{R_1 R_5} \tag{4.5.8}$$

For AC signals the minus sign of Equation (4.5.8) merely shows phase inversion. As shown, Figure 4.5.22 will deliver

$1/2 A_{RMS}$ to the load from an input signal of $250mV_{RMS}$, with THD less than 0.5%. Maximum current variation is typically 0.5% with a load change from 1-5Ω.

Flowmeters, or other similar uses of electromagnets, exemplify application of Figure 4.5.22. Interchangeable electromagnets often have different impedances but require the same constant AC current for proper magnetization. The low distortion, high current capabilities of the LM380 make such applications quite easy.

4.5.15 Muting

Muting, or operating in a squelched mode may be done with the LM380 by pulling the bypass pin high during the mute, or squelch period. Any inexpensive, general purpose PNP transistor can be used to do this function as diagrammed in Figure 4.5.23.

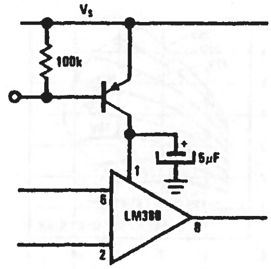


FIGURE 4.5.23 Muting the LM380

During the mute cycle, the output stage will be switched off and will remain off until the PNP transistor is turned off again. Muting attach and release action is smooth and fast.

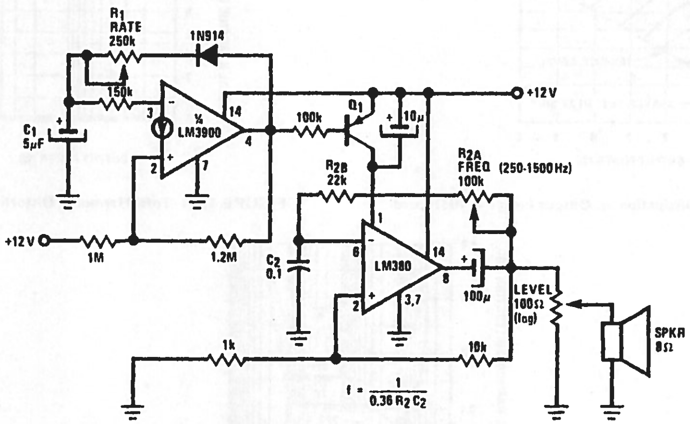


FIGURE 4.5.24 Siren with Programmable Frequency and Rate Adjustment

4.5.16 Siren

Use of the muting technique described in section 4.5.15 allows the LM380 to be configured into a siren circuit with programmable frequency and rate adjustment (Figure 4.5.24. The LM380 operates as an astable oscillator with frequency determined by R_2-C_2. Adding Q_1 and driving its base with the output of an LM3900 wired as a second astable oscillator acts to gate the output of the LM380 on and off at a rate fixed by R_1-C_1. The design equations for the LM3900 astable are given in detail in application note AN-72, page 20, and should be consulted for accurate variation of components. For experimenting purposes (i.e., playing around), changing just about any component will alter the siren effect.

4.6 LM384 AUDIO POWER AMPLIFIER

4.6.1 Introduction

Higher allowed operating voltage, thus higher output power, distinguishes the LM384 from the LM380 audio amplifier. Typical power levels of 7.5W (10% THD) into 8Ω are possible when operating from a supply voltage of 26V. All other parameters remain as discussed for the LM380. The electrical schematic is identical to Figure 4.5.1.

4.6.2 General Operating Characteristics

Package power dissipation considerations regarding heat-sinking are the same as the LM380 (Figure 4.5.4). Device dissipation versus output power curves for 4, 8 and 16Ω loads appear as Figures 4.6.1-4.6.3.

Figure 4.6.4 shows total harmonic distortion vs. output power, while total harmonic distortion vs. frequency for various output levels appears as Figure 4.6.5.

A typical 5W amplifier (V_s = 22V, R_L = 8Ω, THD = 10%) is shown by Figure 4.6.6. Note the extreme simplicity of the circuit. For applications where output ripple and small, high-frequency oscillations are not a problem, all capacitors except the 500μF output capacitor may be eliminated – along with the 2.7Ω resistor. This creates a complete amplifier with only *one* external capacitor and *no* resistors.

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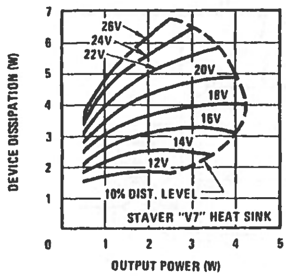


FIGURE 4.6.1 Device Dissipation vs. Output Power — 4 Ω Load

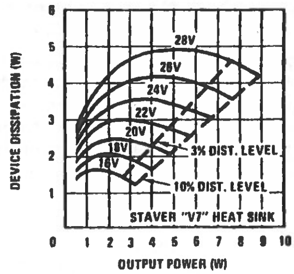


FIGURE 4.6.2 Device Dissipation vs. Output Power — 8 Ω Load

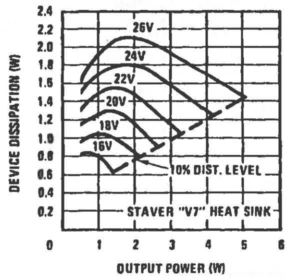


FIGURE 4.6.3 Device Dissipation vs. Output Power — 16 Ω Load

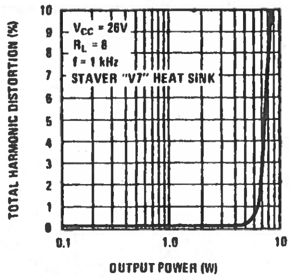


FIGURE 4.6.4 Total Harmonic Distortion vs. Output Power

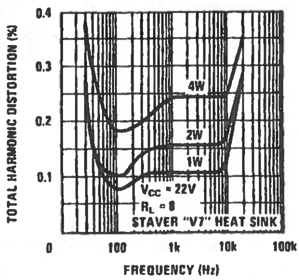


FIGURE 4.6.5 Total Harmonic Distortion vs. Frequency

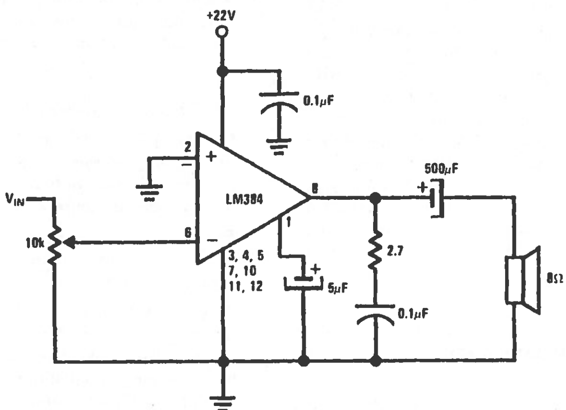


FIGURE 4.6.6 Typical 5W Amplifier

4.7 LM386 LOW VOLTAGE AUDIO POWER AMPLIFIER

4.7.1 Introduction

The LM386 is a power amplifier designed for use in low voltage consumer applications. The gain is internally set to 20 to keep external part count low, but the addition of an external resistor and capacitor between pins 1 and 8 will increase the gain to any value up to 200.

The inputs are ground referenced while the output is automatically biased to one half the supply voltage. The quiescent power drain is only 24mW when operating from a 6V supply, making the LM386 ideal for battery operation.

Comparison of the LM386 schematic (Figure 4.7.1) with that of the LM380 (Figure 4.5.1) shows them to be essentially the same. The major difference is that the LM386 has two gain control pins (1 and 8), allowing the internally set gain of 20V/V (26dB) to be externally adjusted to any value up to 200V/V (46dB). Another important difference lies in the LM386 being optimized for low current drain, battery operation.

4.7.2 General Operating Characteristics

Device dissipation vs. output power curves for 4, 8 and 16Ω loads appear as Figures 4.7.2-4.7.4. Expected power output as a function of typical supply voltages may be noted from these curves. Observe the "Maximum Continuous Dissipation" limit denoted on the 4 and 8Ω curves as a dashed line. The LM386 comes packaged in the 8-pin mini-DIP leadframe having a thermal resistance of 187°C/W, junction to ambient. There exists a maximum allowed junction temperature of 150°C, and assuming ambient temperature equal to 25°C, then the maximum dissipation permitted is 660mW (P_{DMAX} = [150°C − 25°C]/[187°C/W]). Operation at increased ambient temperatures means derating the device at a rate of 187°C/W. Note from Figure 4.7.3 that operation from a 12V supply limits continuous output power to a maximum of 250mW for allowed limits of package dissipation. It is therefore important that the power supply voltage be picked to optimize power output vs. device dissipation.

Figure 4.7.5 gives a plot of voltage gain vs. frequency, showing the wideband performance characteristic of the LM386. Both gain extremes are shown to indicate the narrowing effect of the higher gain setting.

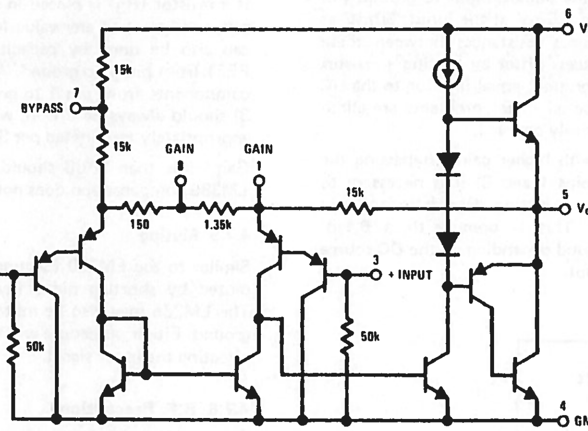


FIGURE 4.7.1 LM386 Simplified Schematic

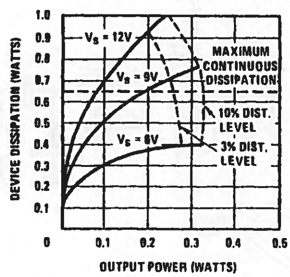


FIGURE 4.7.2 Device Dissipation vs. Output Power — 4Ω Load

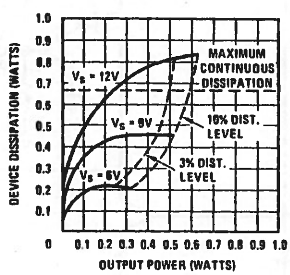


FIGURE 4.7.3 Device Dissipation vs. Output Power — 8Ω Load

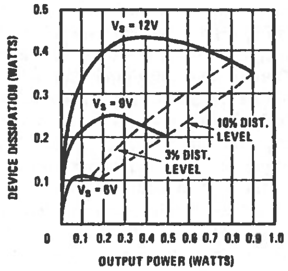


FIGURE 4.7.4 Device Dissipation vs. Output Power — 16Ω Load

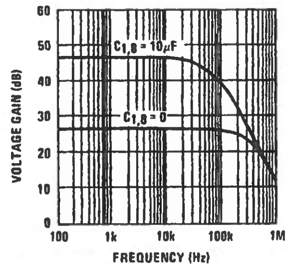


FIGURE 4.7.5 Voltage Gain vs. Frequency

4.7.3 Input Biasing

The schematic (Figure 4.7.1) shows that both inputs are biased to ground with a 50kΩ resistor. The base current of the input transistors is about 250nA, so the inputs are at about 12.5mV when left open. If the DC source resistance driving the LM386 is higher than 250kΩ it will contribute very little additional offset (about 2.5mV at the input, 50mV at the output). If the DC source resistance is less than 10kΩ, then shorting the unused input to ground will keep the offset low (about 2.5mV at the input, 50mV at the output). For DC source resistances between these values we can eliminate excess offset by putting a resistor from the unused input to ground, equal in value to the DC source resistance. Of course all offset problems are eliminated if the input is capacitively coupled.

When using the LM386 with higher gains (bypassing the 1.35kΩ resistor between pins 1 and 8) it is necessary to bypass the unused input, preventing degradation of gain and possible instabilities. This is done with a 0.1μF capacitor or a short to ground depending on the DC source resistance on the driven input.

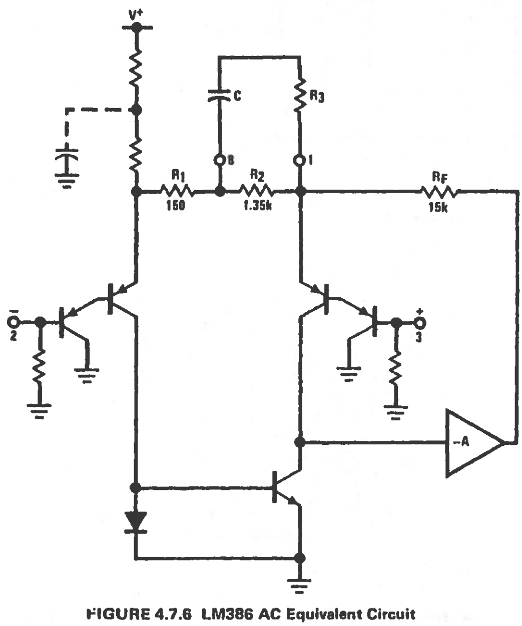


FIGURE 4.7.6 LM386 AC Equivalent Circuit

4.7.4 Gain Control

Figure 4.7.6 shows an AC equivalent circuit of the LM386, highlighting the gain control feature. To make the LM386 a more versatile amplifier, two pins (1 and 8) are provided for gain control. With pins 1 and 8 open the 1.35kΩ resistor sets the gain at 20 (26dB). If a capacitor is put from pin 1 to 8, bypassing the 1.35kΩ resistor, the gain will go up to 200 (46dB).

If a resistor (R_3) is placed in series with the capacitor, the gain can be set to any value from 20 to 200. Gain control can also be done by capacitively coupling a resistor (or FET) from pin 1 to ground. When adding gain control with components from pin 1 to ground, the *positive* input (pin 3) should always be driven, with the negative input (pin 2) appropriately terminated per Section 4.7.3.

Gains less than 20dB should not be attempted since the LM386 compensation does not extend below 9V/V (19dB).

4.7.5 Muting

Similar to the LM380 (Section 4.5.15), the LM386 may be muted by shorting pin 7 (bypass) to the supply voltage. The LM386 may also be muted by shorting pin 1 (gain) to ground. Either procedure will turn the amplifier off without affecting the input signal.

4.7.6 R.F. Precautions

In AM radio applications in particular, r.f. interference caused by radiated wideband noise voltage at the speaker terminals needs to be considered. The pole splitting compensation used in monolithic audio power amplifiers to preserve a wide power bandwidth capability means that there will be plenty of excess

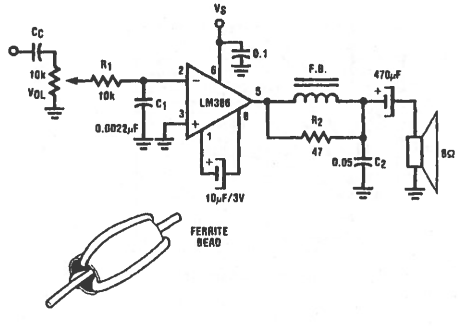


FIGURE 4.7.7 AM Radio Application

gain at frequencies well beyond the audio bandwidth. Noise voltages at these frequencies are amplified and delivered to the load where they can be radiated back to the AM radio ferrite antenna.

Any p.c. board should be layed out to locate the power amplifier as far as possible from the antenna circuit. Extremely tight twisting of the speaker and power supply leads is a must if optimum sensitivity for the radio is to be obtained.

If r.f. radiation still causes a reduction in sensitivity the circuit can be modified as shown in Figure 4.7.7. A typical radio application will use fairly high gain (200V/V) so the device gain is increased by connecting a 10μF capacitor between Pins 1 and 8. To band limit the input signal to 5-10kHz, a two pole filter configuration is used. The first pole is determined by the radio detector circuit and a second pole is added by the R_1C_1 network at the input to the LM386. Any r.f. noise is substantially reduced by placing a ferrite bead (F.B) at the output. A Ferroxcube K5-001-001/3B with 3 turns taken through the bead is suitable for this application. The R_2C_2 network is necessary to stabilize the output stage (Section 4.5.5) but R_2 will also load the ferrite bead, reducing the level of r.f. attenuation. In this instance, a 47Ω resistor is optimum — a smaller value will simply degrade AM sensitivity and a larger value will not ensure stability for all parts. If other ferrite beads are used, a new value for R_2 that will guarantee stability and minimize degradation of AM sensitivity can be found by a few trials.

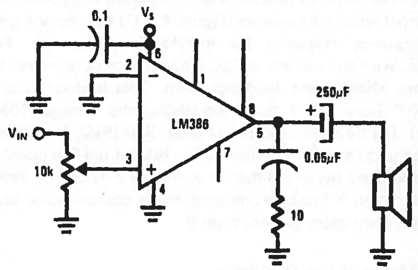


FIGURE 4.7.8 Amplifier with Gain = 20V/V (26dB) Minimum Parts

4.7.7 Typical Applications

Three possible variations of the LM386 as a standard audio power amplifier appear as Figures 4.7.8-4.7.10. Possible gains of 20, 50 and 20V/V are shown as examples of various gain control methods. The addition of the 0.05μF capacitor and 10Ω resistor is for suppression of the "bottom side fuzzies" (i.e., bottom side oscillation occurring during the negative swing into a load drawing high current — see Section 4.5.5).

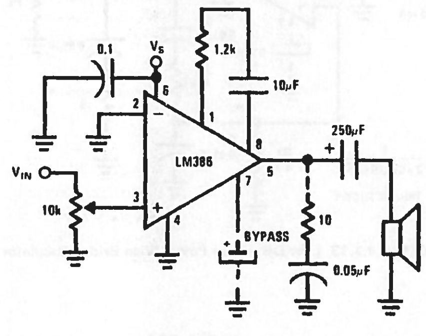


FIGURE 4.7.9 Amplifier with Gain = 50V/V (34dB)

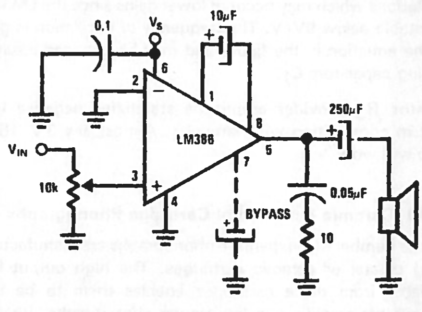
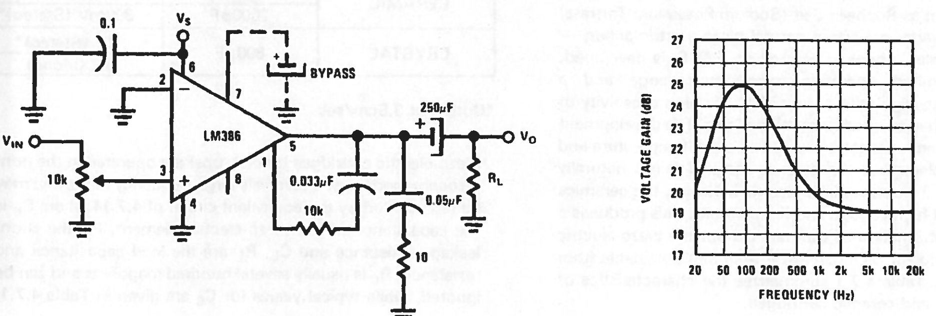


FIGURE 4.7.10 Amplifier with Gain = 200V/V (46dB)



(a) Amplifier with Bass Boost

(b) Frequency Response with Bass Boost

FIGURE 4.7.11 LM386 with Bass Boost

Additional external components can be placed in parallel with the internal feedback resistors (Figure 4.7.11) to tailor the gain and frequency response for individual applications. For example, we can compensate poor speaker bass response by frequency shaping the feedback path. This is done with a series RC from pin 1 to 5 (paralleling the internal 15kΩ resistor). For 6dB effective bass boost: R ≈ 15kΩ, the lowest value for good stable operation is R = 10kΩ if pin 8 is open. If pins 1 and 8 are bypassed then R as low as 2kΩ can be used. This restriction is because the amplifier is compensated only for closed-loop gains greater than 9.

4.7.8 Square Wave Oscillator

A square wave oscillator capable of driving an 8Ω speaker with 0.5W from a 9V supply appears as Figure 4.7.12. Altering either R_1 or C_1 will change the frequency of oscillation per the equation given in the figure. A reference voltage determined by the ratio of R_3 to R_2 is applied to the positive input from the LM386 output. Capacitor C_1 alternately charges and discharges about this reference value, causing the output to switch states. A triangle output may be taken from pin 2 if desired. Since DC offset voltages are not relevant to the circuit operation, the gain is increased to 200V/V by a short circuit betwen the pins 1 and 8, thus saving one capacitor.

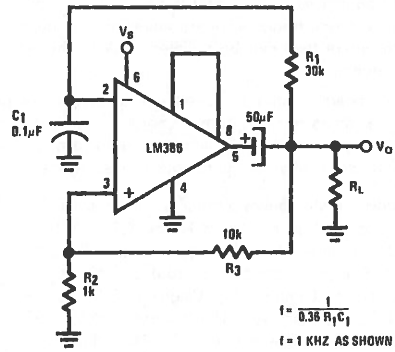


FIGURE 4.7.12 Square Wave Oscillator

4.7.9 Power Wien Bridge Oscillator

The LM386 makes a low cost, low distortion audio frequency oscillator when wired into a Wien brige configuration (Figure 4.7.13). Capacitor C_2 raises the "open loop" gain to 200V/V. Closed-loop gain is fixed at approximately ten by the ratio of R_1 to R_2. A gain of ten is necessary to guard against spurious oscillations which may occur at lower gains since the LM386 is not stable below 9V/V. The frequency of oscillation is given by the equation in the figure and may be changed easily by altering capacitors C_1.

Resistor R_3 provides amplitude stabilizing negative feedback in conjunction with lamp L_1. Almost any 3V, 15mA lamp will work.

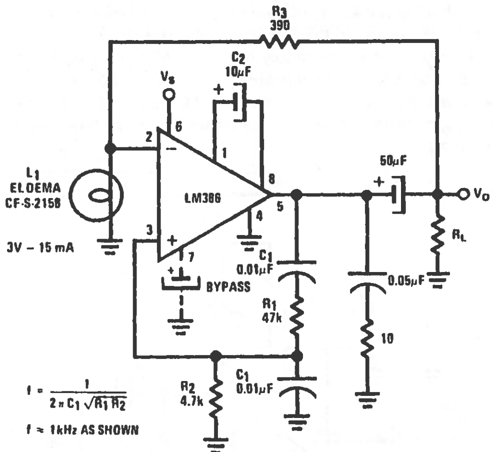


FIGURE 4.7.13 Low Distortion Power Wien Bridge Oscillator

4.7.10 Ceramic and Crystal Cartridge Phonographs

A large number of inexpensive phonographs are manufactured using crystal or ceramic cartridges. The high output level available from these cartridges enables them to be used without pre-amplifiers in low power phonographs. Because the power amplifier is the only active gain element in such systems, the amplifier design should take into account the unique characteristics of piezo-electric cartridges.

Crystal cartridges are typically made from a single crystal material known as Rochelle Salt (Sodium Potassium Tartrate) which, like quartz, exhibits a natural piezo-electric action — when the crystal is bent or twisted an E.M.F. is developed. Despite a limited operating temperature range and a susceptibility to high relative humidity, the high sensitivity of Rochelle Salt has ensured its continued use. The development of modern ceramic titanates has solved many temperature and humidity problems but the ceramic material is not naturally piezo-electric. To obtain piezo-electric behavior, the ceramics are "poled" at high voltage and temperature. This produces a permanent deformation of the material but the piezo-electric action after "poling" is much lower than that obtainable from Rochelle Salt. Table 4.7.1 summarizes the characteristics of typical crystal and ceramic cartridges.

TABLE 4.7.1

CARTRIDGE TYPE	CAPACITANCE	OUTPUT AT 5cm/sec (f = 1kHz)
CERAMIC	800pF	500mV
	2000pF	300mV (Stereo)*
CRYSTAL	800pF	2V (Stereo)*
		3V (Mono)

*Output at 3.5cm/sec

Piezo-electric cartridges (or pick-ups) are operated in the non-resonant mode over a relatively large frequency range and may be represented by the equivalent circuit of 4.7.14 where C_C is the capacitance of the piezo-electric element, R_C the shunt leakage resistance and C_L, R_L are the load capacitance and resistance. R_C is usually several hundred megohms and can be ignored, while typical values for C_C are given in Table 4.7.1.

The E.M.F. generated by any piezo-electric cartridge depends on the amplitude of the movement of the stylus. If discs were recorded with a constant amplitude characteristic, above the cut-off frequency determined by the cartridge capacitance and the load resistance, the response would be essentially flat with frequency, Figure 4.7.15. Note that any load capacitance reduces the output at *all* frequencies above cut-off and that the cut-off frequency moves lower since

$$f_C = \frac{1}{2\pi C_T R_L}$$

where C_T is the paralleled capacitance of the cartridge and the load capacitance.

Since discs are not cut with a constant amplitude versus frequency characteristic (See Section 2.11), when an ideal piezo cartridge plays back a R.I.A.A. recorded disc, there will be a 12.5dB drop in response between 500Hz and 2.1kHz. Before an amplifier response is designed to accomodate this, the designer should realize that crystal cartridges have mechanical compensation to provide relatively flat response through this region, so that a flat amplifier response is all that is required. Ceramic cartridges however, may or may not have mechanical compensation and the decision to compensate electronically will probably depend on the cost objectives (See Section 4.8.7).

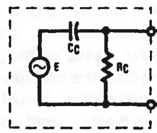


FIGURE 4.7.14 Cartridge Equivalent Circuit

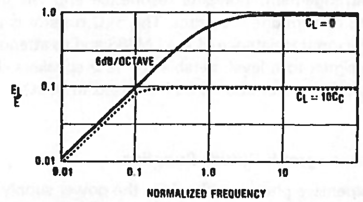


FIGURE 4.7.15 Cartridge Frequency Response

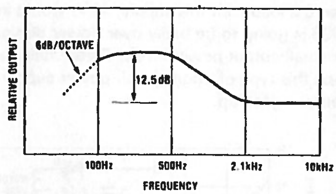


FIGURE 4.7.16 Cartridge Response to RIAA Recorded Disc

4.7.11 LM386 Crystal Cartridge Amplifiers

Where a crystal cartridge is used, the most economical design with the LM386 is shown in Figure 4.7.17. The input stage configuration is the result of a trade-off between cartridge load R_L (which together with the cartridge capacitance will set the low −3dB frequency) and the need to mask variations of input impedance presented by the LM386.

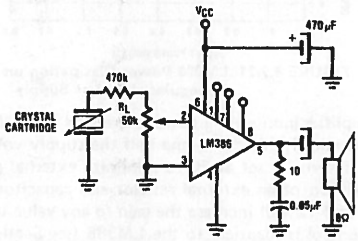


FIGURE 4.7.17 Low Cost Phono Amplifier

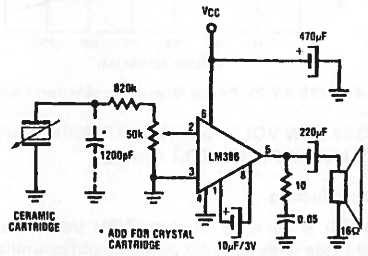


FIGURE 4.7.18 Ceramic Cartridge Amplifier

The resistor R_L is large enough to define the cartridge load for all settings of the volume control, but a signal attenuator is also formed by this resistor and the input resistance of the LM386 (50k) in parallel with the volume potentiometer. With a large valued potentiometer, the amount of signal attenuation will depend of the input resistance of the LM386 which can change by −30% to +100% from device to device. A 50k volume control will mask this variation to less than 4dB for worst case device input resistance change. A smaller volume control will give even less possible variation in output level but the signal become correspondingly more attenuated. Decreasing R_L to restore more signal input to the LM386 will cause further degradation in the cartridge bass response.

4.7.12 Ceramic Cartridge Amplifiers

While the circuit of 4.7.17 can provide a reasonable compromise of output power and bass frequency response with crystal cartridges, the lower output level of ceramic cartridges will require some changes.

In the circuit of 4.7.18 the gain of the LM386 has been raised to 200V/V by connecting a capacitor between pin 1 and pin 8. This will also allow R_L to be increased to 820kΩ, which for a 2000pF capacitance cartridge will give a bass cut-off frequency of under 100Hz. This circuit can be used to accomodate the higher output crystal cartridges without overload simply by adding a 1200pF capacitor across the cartridge terminals. This reduces the crystal cartridge output

$$\text{by } \frac{800}{800+1200} = 0.4 \text{ or } 8\text{dB}$$

and extends the bass response down to 100Hz (compared to the usual bass cut-off of 200Hz). However, for either ceramic or crystal cartridge, extended bass response should be approached with caution, since problems can result from low frequency mechanical feedback between the speaker and the tone arm in complete phonograph units.

This is no problem for stereo units with separated speakers, but for more compact monaural phonographs the circuit of Figure 4.7.18 may cause a low frequency resonance at higher volume settings. It is possible to reconfigure the cartridge loading to prevent this (Figure 4.7.19), by connecting a large valued potentiomenter across the cartridge. For ceramic cartridges 500kΩ is suitable and for crystal cartridges 1mΩ is recommended. At low volume setting the cartridge response is dictated by the size of the potentiometer. At higher volume settings where mechanical feedback could occur, the potentiometer becomes shunted by the series resistance (R) and the input resistance of the LM386. Proper choice of R (dependent on the particular phonograph tone arm and speaker arrangement) prevents resonance and will give the impression of a loudness control. The 5kΩ resistor is used to swamp the input resistance of the LM386 and to attenuate the cartridge signal to a level suitable for 16Ω speakers. For 8Ω speakers, this resistor should be increased to 10kΩ.

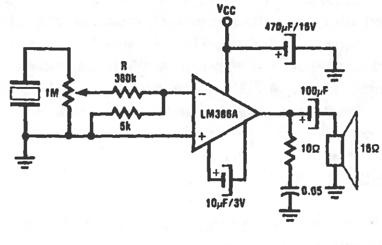


FIGURE 4.7.19 Circuit to Reduce Tone Arm/ Speaker Resonance

4.7.13 Phonograph Power Supplies

Most inexpensive phonographs drive the power supply for the electronics from an overwinding on the phonograph motor, and have a no-load voltage from around 12V to 16V. Inspection of the power dissipation curves for an LM386 driving an 8Ω load with this supply, 12V, would indicate that the LM386 is going to be badly over power dissipation limits, even for small output power levels. Fortunately this is not the case since this type of phonograph power supply sags as the power output goes up.

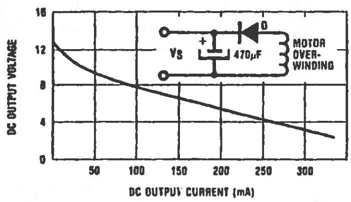


FIGURE 4.7.20 Power Supply Regulation Curve

A typical plot of supply voltage versus output current for a half wave rectified, capacitive input filter power supply is given in Figure 4.7.20. The equivalent internal resistance of the supply (contributed mainly by the winding) is approximately 26Ω. Using this supply regulation curve to plot the intenal power dissipation of the LM386 as the load current increases (Figure 4.7.21) shows that at no time does the power dissipation exceed 600mW. Nevertheless, it is important to check that the peak supply voltage under no-load conditions does not exceed the maximum supply voltage rating for the device.

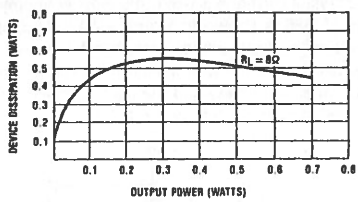


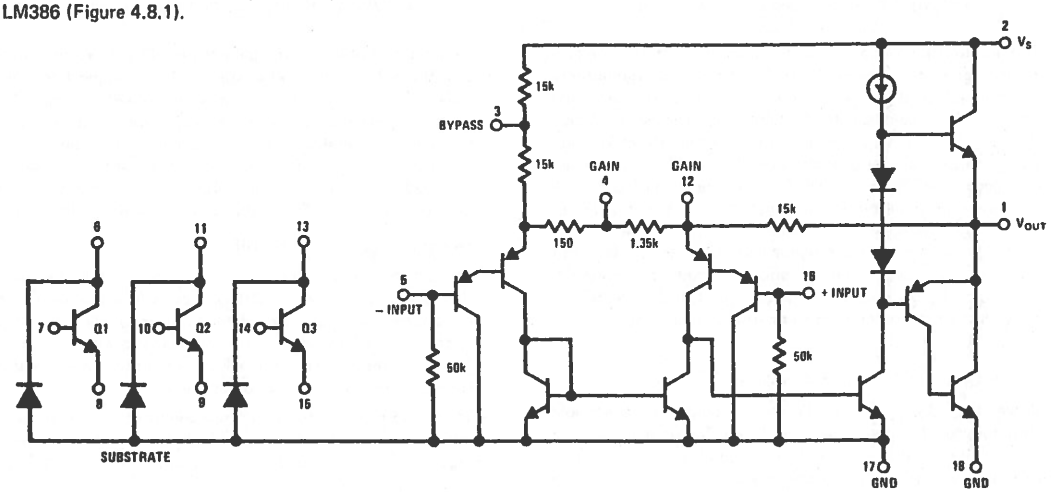
FIGURE 4.7.21 LM386 Power Dissipation on Unregulated Power Supply

4.8 LM389 LOW VOLTAGE AUDIO POWER AMPLIFIER WITH NPN TRANSISTOR ARRAY

4.8.1 Introduction

The LM389 is an array of three NPN transistors on the same substrate with an audio power amplifier similar to the LM386 (Figure 4.8.1).

The amplifier inputs are ground referenced while the output is automatically biased to one half the supply voltage. The gain is internally set at 20 to minimize external parts, but the addition of an external resistor and capacitor between pins 4 and 12 will increase the gain to any value up to 200. Gain control is identical to the LM386 (see Section 4.7.4).



The three transistors have high gain and excellent matching characteristics. They are well suited to a wide variety of applications in DC through VHF systems.

4.8.2 Supplies and Grounds

The LM389 has excellent supply rejection and does not require a well regulated supply. However, to eliminate possible high frequency stability problems, the supply should be decoupled to ground with a 0.1 μF capacitor. The high current ground of the output transistor, pin 18, is brought out separately from small signal ground, pin 17. If the two ground leads are returned separately to supply, the parasitic resistance in the power ground lead will not cause stability problems. The parasitic resistance in the signal ground can cause stability problems and it should be minimized. Care should also be taken to insure that the power dissipation does not exceed the maximum dissipation (825mW) of the package for a given temperature.

4.8.3 Muting

Muting is accomplished in the same manner as for the LM386 (Section 4.7.5), with the exception of applying to different pin numbers.

4.8.4 Transistors

The three transistors on the LM389 are general purpose devices that can be used the same as other small signal transistors. As long as the currents and voltages are kept within the absolute maximum limitations, and the collectors are never at a negative potential with respect to pin 17, there is no limit on the way they can be used.

For example, the emitter-base breakdown voltage of 7.1V can be used as a zener diode at currents from 1 μA to 5mA. These transistors make good LED driver devices; V_{SAT} is only 150mV when sinking 10mA.

In the linear region, these transistors have been used in AM and FM radios, tape recorders, phonographs, and many other applications. Using the characteristic curves on noise voltage and noise current, the level of the collector current can be set to optimize noise performance for a given source impedance (Figures 4.8.2-4.8.4).

4.8.5 Typical Applications

The possible applications of three NPN transistors and a 0.5W power amplifier seem limited only by the designer's imagination. Many existing designs consist of three transistors plus a small discrete power amplifier; redesign with the LM389 is an attractive alternative – typical of these are battery powered AM radios. The LM389 makes a cost-saving single IC AM radio possible as shown in Figure 4.8.5.

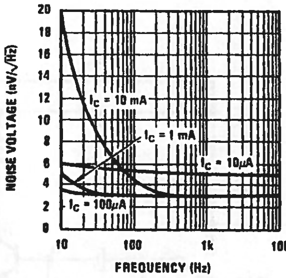


FIGURE 4.8.2 Noise Voltage vs. Frequency

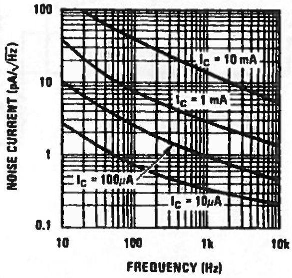


FIGURE 4.8.3 Noise Current vs. Frequency

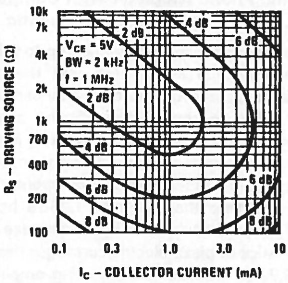
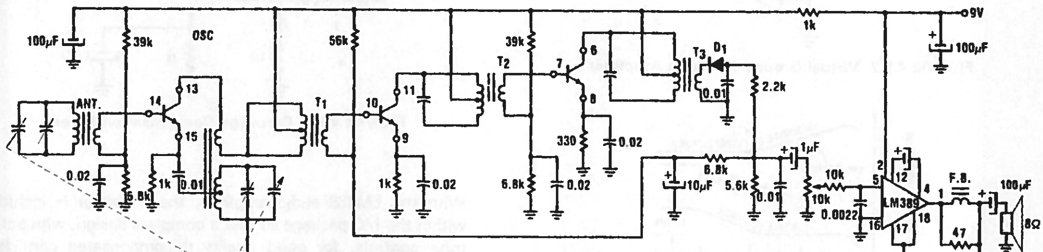


FIGURE 4.8.4 Contours of Constant Noise Figure

4.8.6 Tape Recorder

A complete record/playback cassette tape machine amplifier appears as Figure 4.8.6. Two of the transistors act as signal amplifiers, with the third used for automatic level control during the "record" mode. The complete circuit consists of only the LM389 plus one diode and the passive components.



OSC: TOKO YMO-S2A188R
T1:
T2:
T3: TOKO RHC-1A5006DX
ANT: L = 650μH, Qu = 250

TUNING CAPACITOR: TOKO 2A-20HQZ
FERRITE BEAD: PHILIPS 2809-34401 (3 TURNS THROUGH BEAD)

FIGURE 4.8.5 AM Radio

4

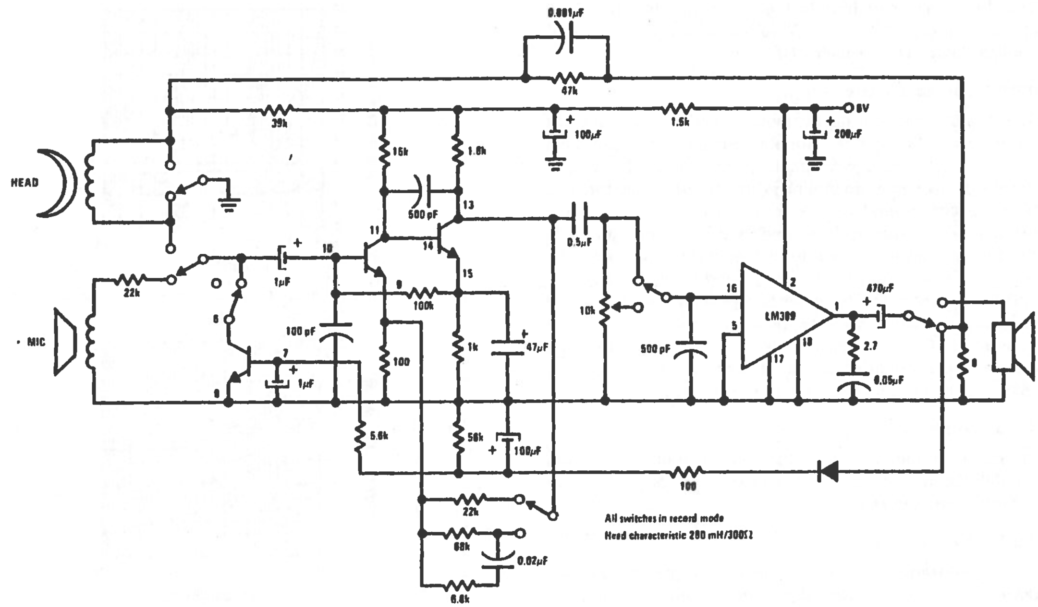


FIGURE 4.8.6 Tape Recorder

4.8.7 Ceramic Phono Amplifier with Compensation for R.I.A.A. Recording Characteristic

All the phonograph amplifiers described up to this point, have been designed on the assumption that the cartridge has mechanical compensation (true for crystal cartridges) or that the 12.5dB fall in response when playing a disc with the R.I.A.A. recording characteristic indicated by Figure 4.7.16, is acceptable. The existence of uncompensated ceramic cartridge implies a need for elecronic compensation — that is an amplifier response that will give 12.5dB boost between 500Hz and 2.1kHz. To achieve this, we can take advantage of the characteristics of piezo-electric cartridges described earlier in Section 4.7.10. Consider the inverting amplifier circuit of Figure 4.8.7. If $R_1C_1 = R_2C_2$ then the frequency response would be flat. Further, if C_1 is the cartridge capacitance, it should be possible to select R_1R_2 and C_2 to compensate for the low frequency roll-off of the cartridge and give a rising reponse between 500Hz and 2.1kHz.

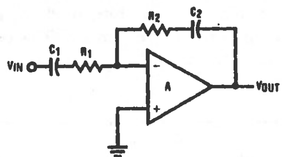


FIGURE 4.8.7 Virtual Ground Inverting Amplifier

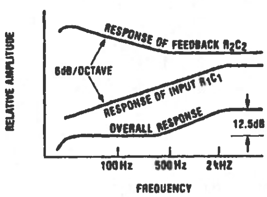


FIGURE 4.8.8 Response of Inverting Amplifier

For a cartridge capacitance of 2000pF, R_1 is selected for a break frequency of 2.1kHz. R_2 and C_2 are chosen to give a break frequency of 500Hz. The amplifier response with each of these networks and the combined response is shown in Figure 4.8.8.

It would be difficult to implement this type of equalization with LM386 amplifiers because of the variation of input resistance and the need for a volume control. Instead a single transistor cartridge-compensation stage can be built to precede the power amplifier, Figure 4.8.9. For the 2000pF cartridge, R_1 is 39kΩ. R_2 is chosen to give slightly more than unity gain so that the output at medium to high frequencies is the same as the cartridge rating (measured at 1kHz where the response is −6dB for an uncompensated cartridge on a R.I.A.A. recording). With $R_2 = 62k\Omega$, a 0.005μF capacitor gives the 500Hz break frequency.

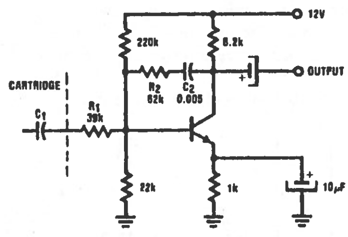


FIGURE 4.8.9 Cartridge-Compensation Stage

With the LM389 audio amplifier, the transistor is included within the I/C package so that a complete design, with active tone controls, for good quality uncompensated cartridges appears as shown in Figure 4.8.10. See Section 2.14.7 for the design of the the tone control circuit.

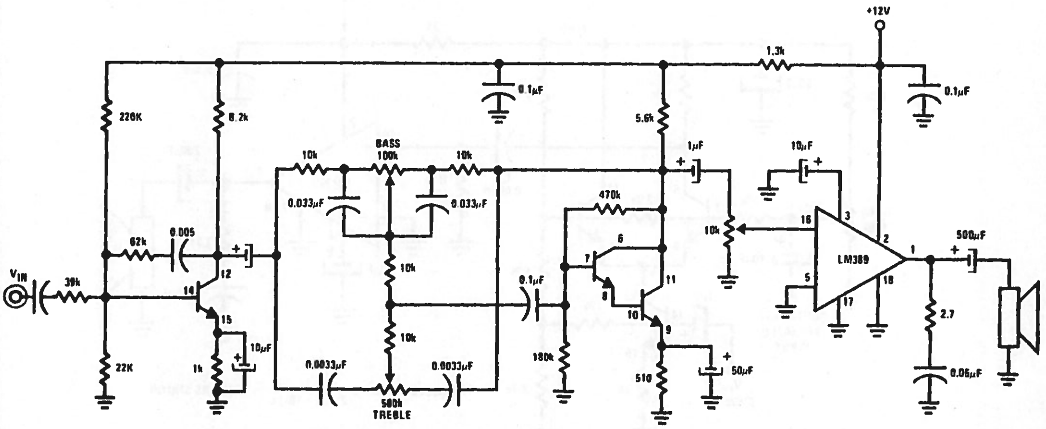


FIGURE 4.8.10 Ceramic Phono Amplifier with Tone Controls

4.8.8 Siren

The siren circuit of Figure 4.8.11 uses one of the LM389 transistors to gate the power amplifier on and off by applying one of the muting techniques discussed in Section 4.8.3. The other transistors form a cross-coupled multivibrator circuit that controls the rate of the square wave oscillator. The power amplifier is used as the square wave oscillator with individual frequency adjust provided by potentiometer R_{2B}.

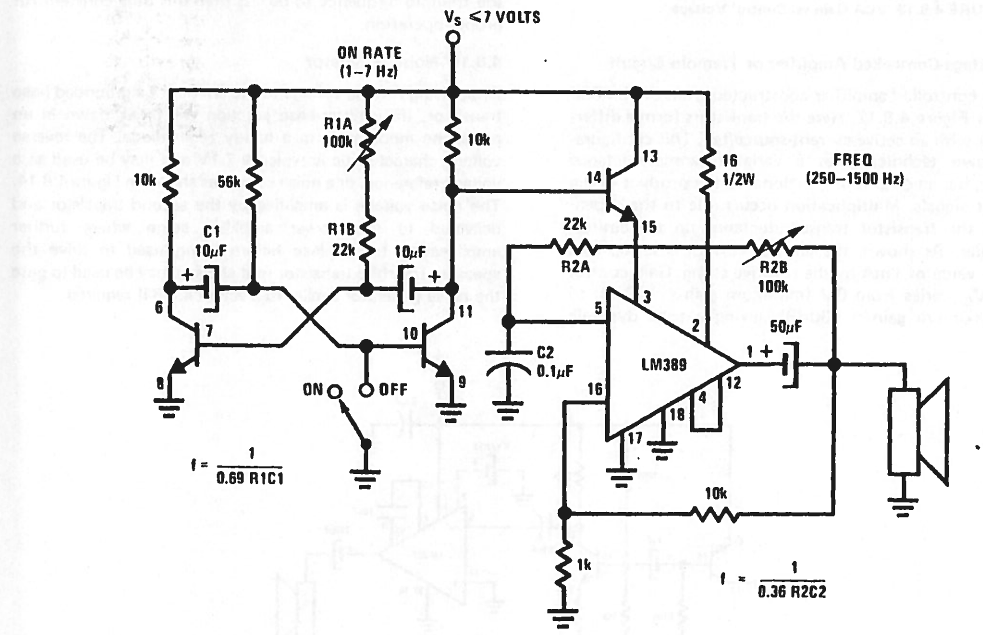


FIGURE 4.8.11 Siren

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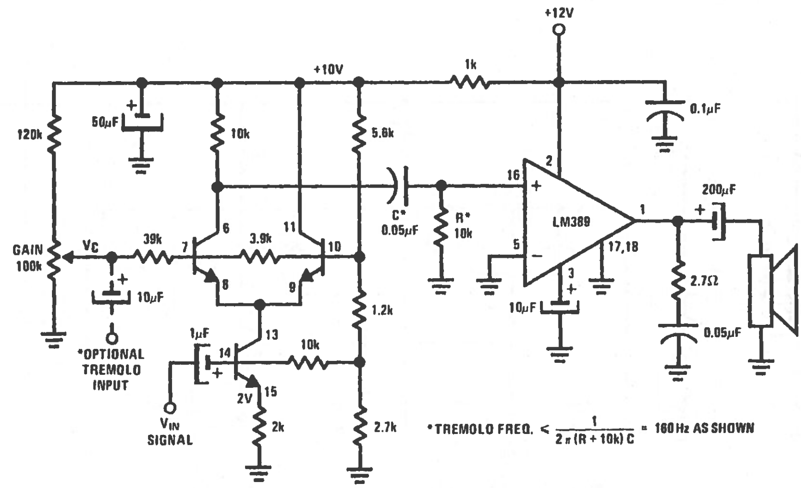


FIGURE 4.8.12 Voltage-Controlled Amplifier or Tremolo Circuit

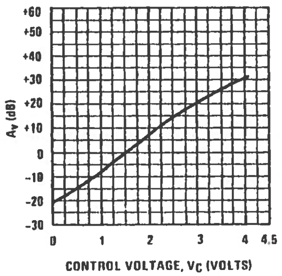


FIGURE 4.8.13 VCA Gain vs. Control Voltage

4.8.9 Voltage-Controlled Amplifier or Tremolo Circuit

A voltage-controlled amplifier constructed from the LM389 appears as Figure 4.8.12. Here the transistors form a differential pair with an active current-source tail. This configuration, known technically as a variable-transconductance multiplier, has an output proportional to the product of the two input signals. Multiplication occurs due to the dependence of the transistor transconductance on the emitter current bias. As shown, the emitter current is set up to a quiescent value of 1 mA by the resistive string. Gain control voltage, V_C, varies from 0V (minimum gain = −20dB) to 4.5V (maximum gain = +30dB), giving a total dynamic range of 50dB (Figure 4.8.13). V_{IN} signal levels should be restricted to less than 100mV for good distortion performance. The output of the differential gain stage is capacitively fed to the power amplifier via the R-C network shown, where it is used to drive the speaker.

Tremolo (amplitude modulation of an audio frequency by a sub-audio oscillator — normally 5-15Hz) applications require feeding the low frequency oscillator signal into the optional input shown. The gain control pot may be set for optimum "depth." Note that the interstage R-C network forms a high pass filter (160Hz as shown), thus requiring the tremolo frequency to be less than this time constant for proper operation.

4.8.10 Noise Generator

By applying reverse voltage to the emitter of a grounded base transistor, the emitter-base junction will break down in an avalanche mode to form a handy zener diode. The reverse voltage characteristic is typically 7.1V and may be used as a voltage reference, or a noise source as shown in Figure 4.8.14. The noise voltage is amplified by the second transistor and delivered to the power amplifier stage where further amplification takes place before being used to drive the speaker. The third transistor (not shown) may be used to gate the noise generator similar to Section 4.8.8 if required.

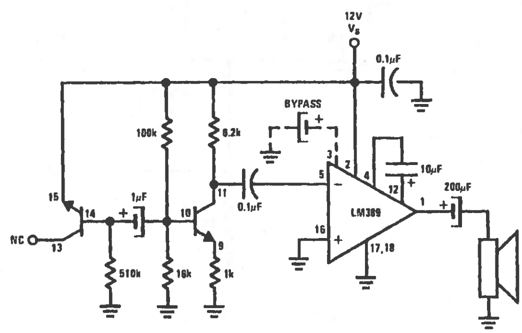


FIGURE 4.8.14 Noise Generator Using Zener Diode

4.8.11 Logic Controlled Mute

Various logic functions are possible with the three NPN transistors, making logic control of the mute function possible. Figures 4.8.15 - 4.8.17 show standard AND, OR and Exclusive-OR circuits for controlling the muting transistor. Using the optional mute scheme of shorting pin 12 to ground gives NAND, NOR and Exclusive NOR

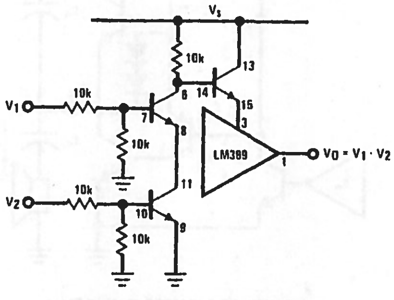


FIGURE 4.8.15 AND Muting

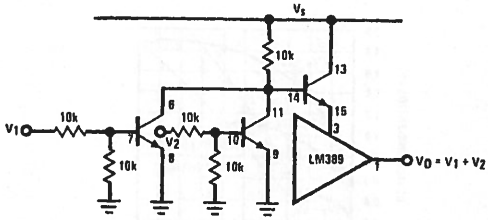


FIGURE 4.8.16 OR Muting

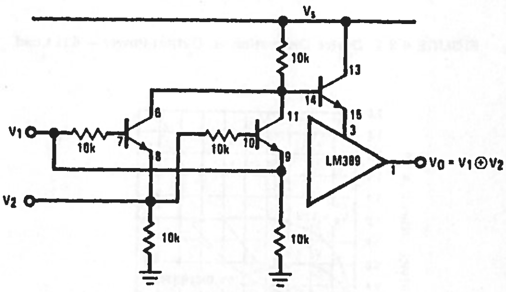


FIGURE 4.8.17 Exclusive-OR Muting

4.9 LM388 BOOTSTRAPPED AUDIO POWER AMPLIFIER

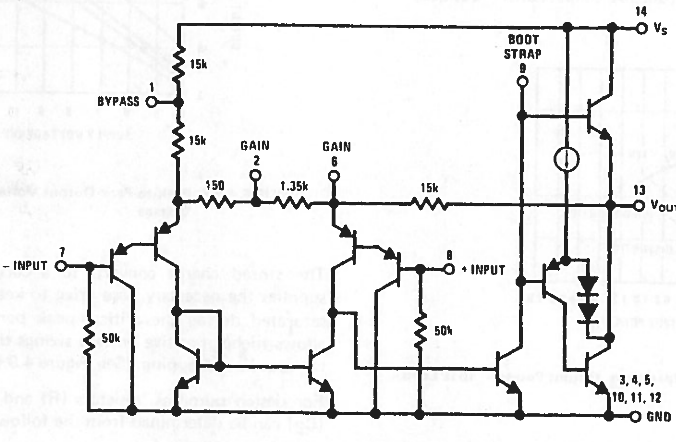


FIGURE 4.9.1 LM388 Simplified Schematic

4.9.1 Introduction

The LM388 audio power amplifier, designed for low voltage, medium power consumer applications, extends the LM386 design concept one step further by incorporating a bootstrapped output stage (Figure 4.9.1). Bootstrapping allows power levels in excess of 1W to be obtained from battery powered products (Figures 4.9.2-4.9.4). Packaging the LM388 into National's 14-pin copper lead-frame (same as LM380) extends maximum package dissipation to values where heatsinking is eliminated for most designs.

4.9.2 General Operating Characteristics

The gain, internally set to 20V/V, is externally controlled in the same manner as the LM386. Consult Section 4.7.4 for details. Input biasing follows LM386 procedures outlined in Section 4.7.3; likewise, muting is the same as Section 4.7.5.

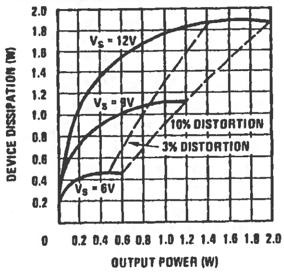


FIGURE 4.9.2 Device Dissipation vs. Output Power — 4Ω Load

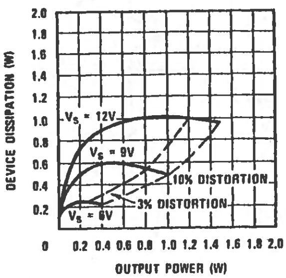


FIGURE 4.9.3 Device Dissipation vs. Output Power — 8Ω Load

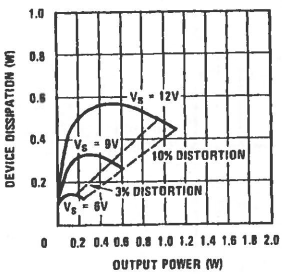


FIGURE 4.9.4 Device Dissipation vs. Output Power — 16Ω Load

4.9.3 Bootstrapping (See also section 4.1.5.)

The base of the top side output transistor is brought out to pin 9 for bootstrapping. The term "bootstrapping" (derived from the expression, ". . . pull oneself up by one's bootstraps") aptly describes the effect. Figure 4.9.5 shows the output stage with the external parts necessary for standard bootstrapping operation. Capacitor C_B charges to approximately $V_S/4$ during the quiescent state of the amplifier and then acts to pull the base of the top transistor up ("by the bootstraps") as the output stage goes through its positive swing — actually raising pin 9 to a *higher* potential than the supply at the top of the swing. This occurs since the voltage on a capacitor cannot change instantaneously, but must decay at a rate fixed by the resistive discharge path.

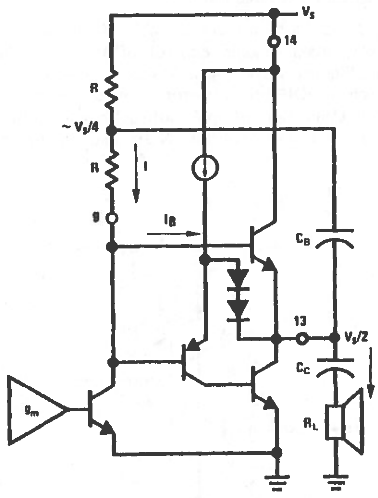


FIGURE 4.9.5 LM388 Output Stage

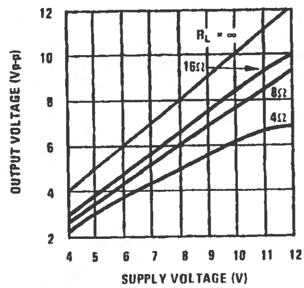


FIGURE 4.9.6 Peak-to-Peak Output Voltage Swing vs. Supply Voltage

The stored charge converts to a current with time and supplies the necessary base drive to keep the top transistor saturated during the critical peak period. The net effect allows higher positive voltage swings than can be achieved without bootstrapping. (See Figure 4.9.6.)

For design purposes, resistors (R) and bootstrap capacitor (C_B) can be determined from the following:

$$I_B = \frac{I_L}{\beta} = \frac{V_S/2 - V_{BE}}{2R} \approx \frac{V_S}{4R}$$

$$\therefore\ I_L = \frac{\beta V_S}{4R}$$

$$\text{also, } I_{L(max)} = \frac{V_S/2}{R_L}$$

$$\text{so, } \frac{\beta V_S}{4R} = \frac{V_S}{2R_L}$$

$$\text{or, } R = \frac{\beta R_L}{2} \qquad (4.9.1)$$

To preserve low frequency performance the pole due to C_B and R/2 (parallel result of R-R) is set equal to the pole due to C_C and R_L:

$$\frac{R}{2} C_B = R_L C_C \tag{4.9.2}$$

Substituting Equation (4.9.1) into (4.9.2) yields:

$$C_B = \frac{4\,C_C}{\beta} \tag{4.9.3}$$

Letting $\beta = 100$ (nominal) gives:

$$R = 50\,R_L \tag{4.9.4}$$

$$C_B = \frac{C_C}{25} \tag{4.9.5}$$

For reduced component count the load can replace the upper resistor, R (Figure 4.9.7). The value of bootstrap resistors R+R must remain the same, so the lower R is increased to 2R (assuming speaker resistance to be negligible). Output capacitor (C_C) now serves the dual function of bootstrapping and coupling. It is sized about 5% larger since it now supplies base drive to the upper transistor.

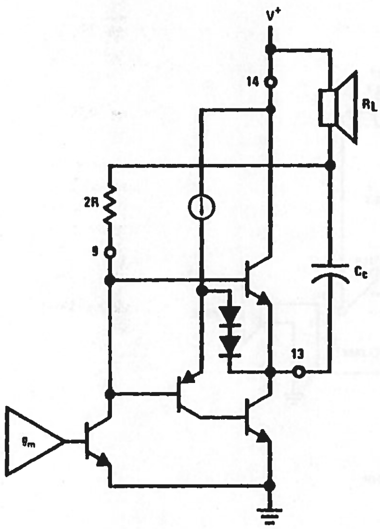


FIGURE 4.9.7 Bootstrapping with Load to Supply

Examples of both bootstrapping methods appear as Figures 4.9.8 and 4.9.9. Note that the resistor values are slightly larger than Equation (4.9.4) would dictate. This recognizes that $I_{L(max)}$ is, in fact, always less than $(V_S/2)/R_L$ due to saturation and V_{BE} losses.

A third bootstrapping method appears as Figure 4.9.10, where the upper resistor is replaced by a diode (with a subsequent increase in the resistance value of the lower resistor). Addition of the diode allows capacitor C_B to be decreased by about a factor of four, since no stored charge is allowed to discharge back into the supply line.

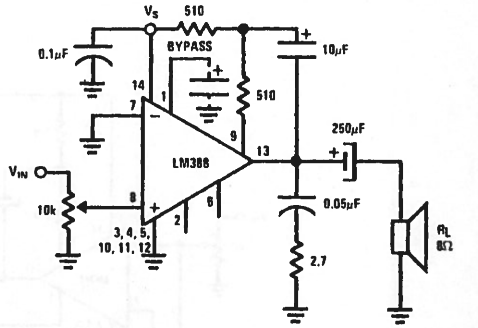


FIGURE 4.9.8 Load Returned to Ground (Amplifier with Gain = 20)

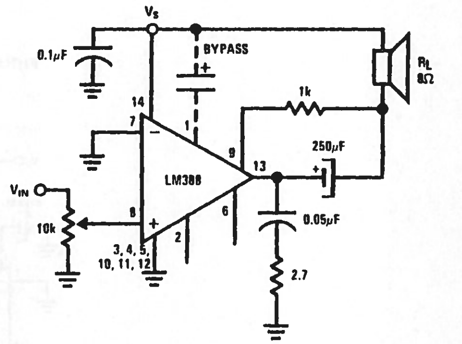


FIGURE 4.9.9 Load Returned to V_S (Amplifier with Gain = 20)

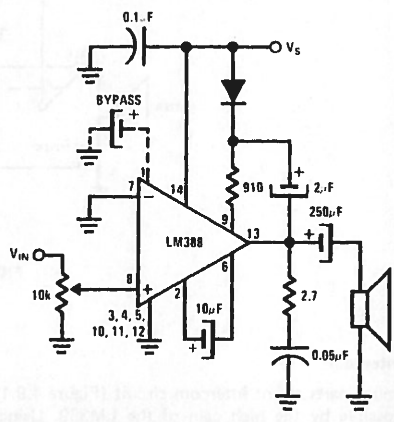


FIGURE 4.9.10 Amplifier with Gain = 200 and Minimum C_B

4.9.4 Bridge Amplifier

For low voltage applications requiring high power outputs, the bridge connected circuit of Figure 4.9.11 can be used. Output power levels of 1.0W into 4Ω from 6V and 3.5W into 8Ω from 12V are typical. Coupling capacitors are not necessary since the output DC levels will be within a few tenths of a volt of each other. Where critical matching is required the 500k potentiometer is added and adjusted for zero DC current flow through the load.

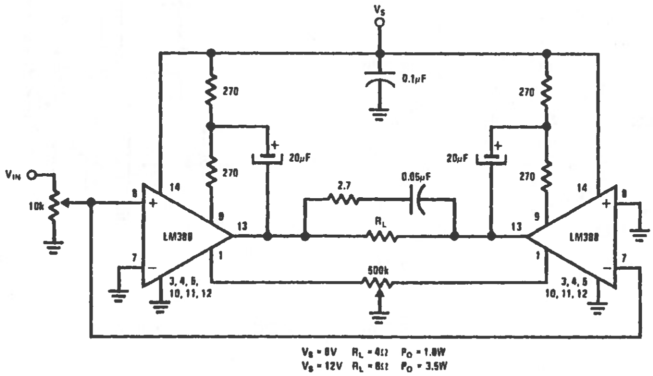


FIGURE 4.9.11 Bridge Amp

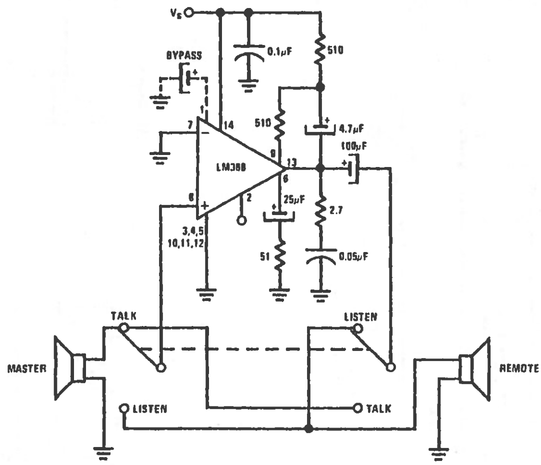


FIGURE 4.9.12 Intercom

4.9.5 Intercom

A minimum parts count intercom circuit (Figure 4.9.12) is made possible by the high gain of the LM388. Using the gain control pin to set the AC gain to approximately 300V/V ($A_V \approx 15k/51\,\Omega$) allows elimination of the step-up transformer normally used in intercom designs (e.g., Figure 4.5.22). The 2.7Ω-0.05μF R-C network suppresses spurious oscillations as described for the LM380 (Section 4.5.5).

4.9.6 FM Scanners and Two Way Walkie Talkies

Designed for the high volume consumer market, the LM388 ideally suits applications in FM scanners and two way walkie talkie radios. Requirements for this market generally fall into three areas:

1. Low cost FM scanners; $V_S = 6V$, $P_O = 0.25W$
2. Consumer walkie talkie (including CB); $V_S = 12V$, $P_O = 0.5W$
3. High quality hand-held portables; $V_S = 7.5V$, $P_O = 0.5W$

Since all equipment is battery operated, current consumption is important; also, the amplifier must be squelchable, i.e., turned off with a control signal. The LM388 meets both of these requirements. When squelched, the LM388 draws only 0.8mA from a 7.5V power supply.

A typical high quality hand held portable application with noise squelch appears as Figure 4.9.13. Diodes D_1 and D_2 rectify noise from the limiter or the discriminator of the receiver, producing a DC current to turn on Q_1, which clamps the LM388 in an off condition.

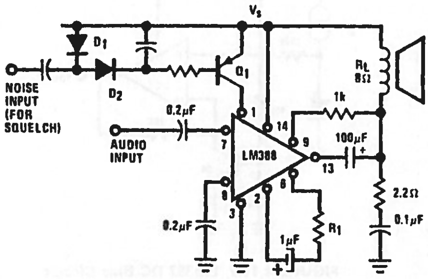


FIGURE 4.9.13 LM388 Squelch Circuit for FM Scanners and Walkie Talkies

As shown, the following performance is obtained:

- Voltage gain equals 20 to 200 (selectable with R_1).
- Noise (output squelched) equals 20μV.
- $P_O = 0.53W$ ($V_S = 7.5V$, $R_L = 8\Omega$, THD = 5%)
- $P_O = 0.19W$ ($V_S = 4.5V$, $R_L = 8\Omega$, THD = 5%)
- Current consumption ($V_S = 7.5V$):
  squelched – 0.8mA
  $P_O = 0.5W$ – 110mA

4.10 LM390 1 WATT BATTERY OPERATED AUDIO POWER AMPLIFIER

Battery operated consumer products often employ 4Ω speaker loads for increased power output. The LM390 meets the stringent output voltage swings and higher currents demanded by low impedance loads. Bootstrapping of the upper output stage (Figure 4.10.1) maximizes positive swing, while a unique biasing scheme (Figure 4.10.2) used on the lower half allows negative swings down to within one saturation drop above ground. Special processing techniques are employed to reduce saturation voltages to a minimum. The result is a monolithic solution to the difficulties of obtaining higher power levels from low voltage supplies. The LM390 delivers 1W into 4Ω (6V) at a lower cost than any competing approach, discrete or IC Figure 4.10.3).

In all other respects (including pin-out) the LM390 is identical to the LM388 (Section 4.9). Gain control, input biasing, muting, and bootstrapping are all as explained previously for the LM386 and LM388.

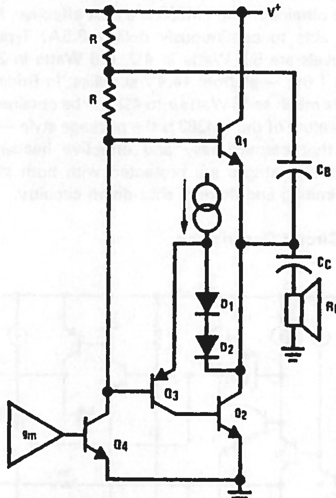


FIGURE 4.10.2 LM390 Output Stage

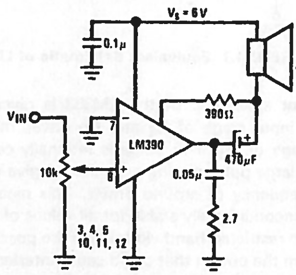


FIGURE 4.10.3 1 Watt Power Amplifier for 6 Volt Systems

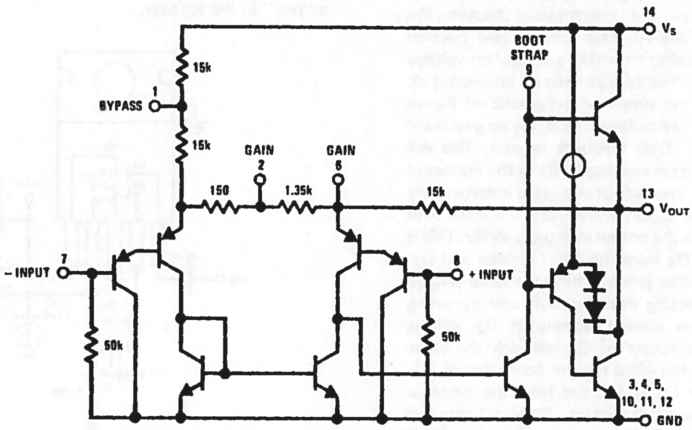


FIGURE 4.10.1 LM390 Simplified Schematic

4

4.11 LM383 8 WATT AUDIO POWER AMPLIFIER

4.11.1 Introduction

The limited supply voltage available in automotive applications requires amplifiers with an extremely high current output capability to drive low impedance loads if high power outputs are to be obtained. The LM383 is a cost effective, high power amplifer able to continuously deliver 3.5A. Typical output power levels are 5.5 Watts in 4Ω, 8.6 Watts in 2Ω and 9.3 Watts in 1.6Ω — all from 14.4V supplies. In Bridge amplifier circuits as much as 16 Watts into 4Ω can be obtained! Another unique feature of the LM383 is the package style — a five lead TO-220 that permits easy and effective heatsinking. The LM383 output stages are protected with both short circuit current limiting and thermal shut-down circuitry.

4.11.2 Circuit Description

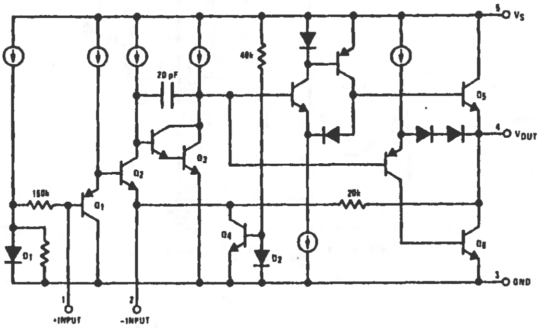


FIGURE 4.11.1 Equivalent Schematic of LM383

An equivalent schematic of the LM383 is given in Figure 4.11.1. The input stage of Q_1 and Q_2 drives the transconductance stage of Q_3. This stage is internally compensated with a fairly large pole splitting capacitor to give a unity gain crossover frequency of around 3mHz. This means that the amplifier is unconditionally stable for all values of closed loop gain, and the restricted bandwidth limits the possibility of r.f. radiation from the output that could cause interference in AM radio applications (see Section 2.3.10). The bandwidth for a closed loop gain of 40dB is still 30kHz (Figure 4.11.9), with careful design of the output stages keeping the *open loop* THD at 1%. The available pin-outs prevent boot-strapping the upper output stage, but the AB bias scheme (see Section 4.1.5) allows a negative swing to within a saturation voltage of ground (Figure 4.11.8). The LM383 uses an interesting dc bias scheme, shown in the simplified schematic of Figure 4.11.2 which has two main advantages. First, the dc gain is set internally to unity by the 20kΩ feedback resistor. This will minimize input offset voltages causing shifts in the quiescent output voltage. Secondly, the output voltage is automatically established at one half the supply voltage and will track with supply voltage to maximize the output swing capability. This is accomplished by biasing Q_4 from the 40kΩ resistor and D_2. Since D_2 and Q_4 base-emitter junction have the same voltage across them, then (neglecting base currents and assuming matching geometries) the current flowing in D_2 will be "mirrored" in Q_4. The collector of Q_4 will sink the same current as that flowing in the 40kΩ resistor connected to V^+. The collector current for Q_4 is sourced from the amplifier output stage through the 20kΩ resistor. Since V^+ appears across the 40kΩ resistor, $V^+/2$ is forced across the 20kΩ resistor and the output will track at one half supply voltage.

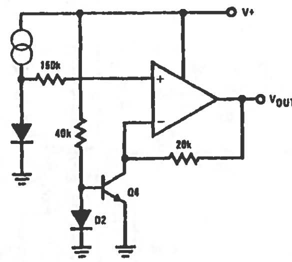


FIGURE 4.11.2 LM383 DC Bias Circuit

4.11.3 General Operating Characteristics

The closed loop gain of the LM383 is set by external components, Figure 4.11.3 showing a typical non-inverting amplifier circuit with A_V set by the ratio of R_1 and R_2. In practical terms the input dynamic range ($\pm 0.5V_{MAX}$) will determine the lowest useable gain for a given output power and load. The circuit of Figure 4.11.3 is set up for $A_V = 1 + R_1/R_2 = 101$, and it is worth noting the unusually low values of the feedback resistors. This can be attributed to the need for supply ripple rejection. Refering back to Figure 4.11.2, any supply voltage ripple will cause a change in the current in the 40kΩ resistor. This change is "mirrored" in Q_4 and without any external ac feedback the ripple voltage would appear attenuated by only −6dB at the output. However, if C_1 is large enough, the feedback network works to prevent any ac voltage change at the amplifier inverting input, so that the ripple appears as a current in the feedack resistor R_1. To a first order approximation therefore, the ripple at the output is given by the ratio of R_1 to the internal 40kΩ resistor.

By using a 220Ω feedback resistor, the ripple rejection ratio obtained is better than 40dB. Although low resistor values mean that more power will be dissipated in the feedback network, the dc voltage across R_2 is 0.7V, and that across R_1 is typically 6.5V, giving the power ratings shown.

The non-inverting input has a relatively high input impedance, but a large input coupling capacitor is recommended. Before turn-on, both the input capacitor and the feedback capacitor are at ground potential. At turn-on, the feedback capacitor can charge up more quickly than the internal bias resistor can charge the input coupling capacitor. This prevents the output from rapidly going up to the positive supply rail and producing a "pop" in the speaker.

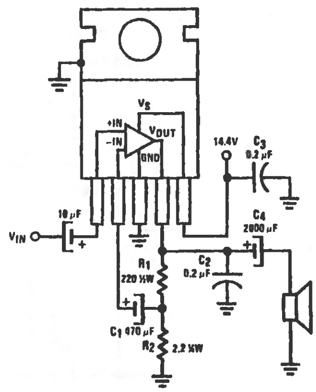


FIGURE 4.11.3 Non Inverting Amplifier (Av ≈ 40dB)

4.11.4 Layout, Ground Loops and Supply Bypassing

The very high output current capability of the LM383 means that careful attention should be paid the p.c.b. layout. A suitable component layout is shown in Figure 4.11.4. Parts worth noting are:

1) Supply decoupling capacitor (C_3) is located right at the supply pin.
2) A 0.2μF capacitor (C_2) is located at the output pin to prevent negative swing parasitic oscillation — *there is no damping resistor in series with this capacitor.*
3) The input ground is returned to the center pin of the I/C — the output or power ground is through the tab via the heatsink.

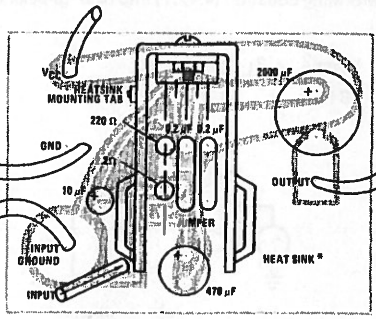


FIGURE 4.11.4 LM383 Board Layout

4.11.5 Output Power and Heat Sinking.

Device power dissipation vs. power output is indicated in Figure 4.11.6 for 4Ω loads and 4.11.7 for 2Ω loads. The ability of the LM383 to sustain these dissipation levels is given by Figure 4.11.5. For example, when driving a 4Ω load, the circuit of Figure 4.11.3 will have a maximum device dissipation of 3.5 Watts. This can be comfortably handled by a 13°C/W heatsink, such as the Staver V-5. If the load resistance is 2Ω, considerably more heatsink capability is required since the maximum device dissipation is now over 6 watts. In this instance a heatsink equivalent to the Staver V3-3-2 would be suitable.

For most applications, since the tab of the LM383 package is grounded, the device can be bolted to the chassis to provide adequate heat sinking.

4.11.6 High Voltage Operation

The LM383 has a maximum supply voltage rating of 20V, above which the amplifier will shut down. The LM383A selection will withstand momentary peak supply voltages of 40V caused by supply-line transients. In an automotive application, a worst case transient is usually caused by alternator "load-dump" or loss of the battery charging load. When a 50 amp alternator loses the load, a peak output voltage of about 120V is generated and the transient on the supply line lasts for many milliseconds. Fortunately for the radio, this transient is clamped by an "A" line L-C filter to between 35-40 volts. The LM383A is rated to withstand 40 volts for 50mSecs.

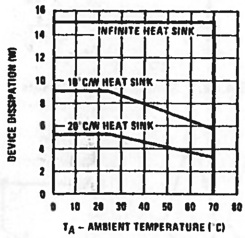


FIGURE 4.11.5 Device Dissipation vs. Ambient Temperature

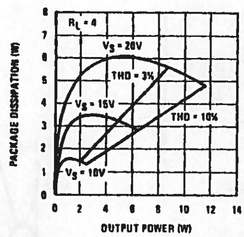


FIGURE 4.11.6 Power Dissipation vs. Output Power

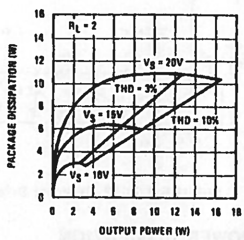


FIGURE 4.11.7 Power Dissipation vs. Output Power

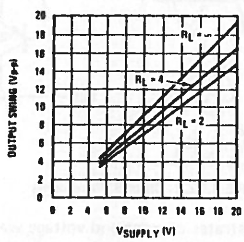


FIGURE 4.11.8 Output Swing vs. Supply Voltage

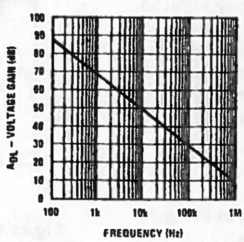


FIGURE 4.11.9 Open Loop Gain vs. Frequency

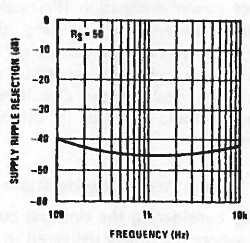


FIGURE 4.11.10 Supply Ripple Rejection vs. Frequency

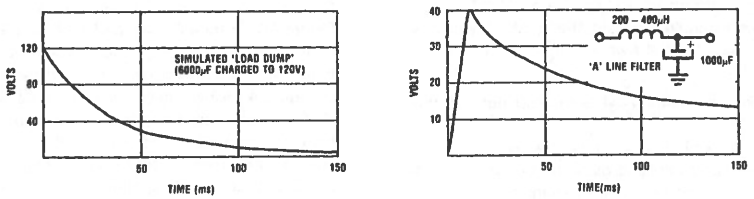


FIGURE 4.11.11 Effect of "A" Line Filter on Automotive Transient

Where more power into a given load is required, two LM383's can be used in a bridge configuration (Figure 4.11.12). The power output and device dissipation per amplifier may be estimated by assuming each amplifier is driving a load of $R_L/2$, in this case 2Ω. On a 14.4V supply, each amplifier can deliver 8W at the 10% THD level for a total output power of 16W (Figure 4.11.7). Each amplifier is dissipating a maximum of 6 watts requiring a *total* heat-sink capability of 5°C/W for the complete system (Figure 4.11.5).

A 100Ω potentiometer is used to trim out the differences in individual LM383 dc output levels since, with a direct connected load, substantial dc power consumption can result if the quiescent output levels are not matched. Although the LM383 is designed to be proof against ac short circuits, this will not be the case for the circuit of Figure 4.11.12. If there is a chance that either side of the load could be shorted to ground, coupling capacitors should be included in each output.

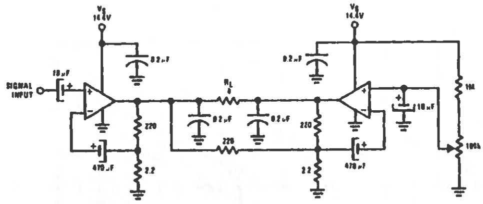


FIGURE 4.11.12 16 Watt Bridge Amplifier

4.12 POWER DISSIPATION

Power dissipation within the integrated circuit package is a very important parameter requiring a thorough understanding if optimum power output is to be obtained. An incorrect power dissipation (P_D) calculation may result in inadequate heatsinking, causing thermal shutdown to operate and limit the output power. All of National's line of audio power amplifiers use class B output stages. Analysis of a typical (ideal) output circuit results in a simple and accurate formula for use in calculating package power dissipation.

4.12.1 Class B Power Considerations

Begin by considering the simplest audio circuit as in Figure 4.12.1, where the power delivered to the load is:

$$P_o = \frac{V_O^2}{R_L} = I_O^2 R_L \qquad (4.12.1)$$

where: P_o = power output

V_O = RMS output voltage

I_O = RMS output current

Transforming Equation (4.12.1) into peak-to-peak quantities gives:

$$P_o = \frac{V_{OPP}^2}{8 R_L} = \frac{R_L I_{OPP}^2}{8} \qquad (4.12.2)$$

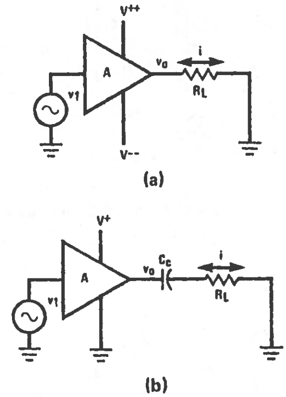


FIGURE 4.12.1 Simple Audio Circuits

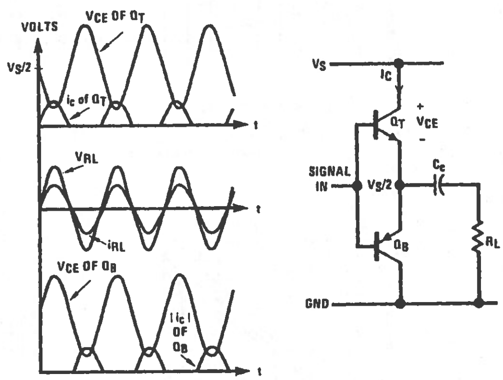


FIGURE 4.12.2 Class B Waveforms

Figure 4.12.2 illustrates current and voltage waveforms in a typical class B output. Dissipation in the top transistor Q_T is the product of collector-emitter voltage and current, as shown on the top axis. Certainly Q_T dissipates zero power when the output voltage is not swinging, since the collector current is zero. On the other hand, if the output waveform is overdriven to a square wave (delivering maximum power to the load, R_L) Q_T delivers large currents, but the voltage across it is zero — again resulting in zero power. In the range of output powers between these extremes, Q_T goes through a point of maximum dissipation. This point always occurs when the peak-to-peak output voltage is 0.637 times the power supply. At that level, assuming all class B power is dissipated in the two output transistors, the chip dissipation is:

$$\max P_D = \frac{V_s^2}{2\pi^2 R_L} \approx \frac{V_s^2}{20 R_L} \qquad (4.12.3)$$

Inserting the applicable supply voltage and load impedance into Equation (4.12.3) gives the information needed to size the heat sink for worst case conditions.

4.12.2 Derivation of Max P_D

The derivation of Equation (4.12.3) for maximum power dissipation follows from examination of Figure 4.12.2 and application of standard power formulas:

Neglect X_{C_c} and let $V_L{}'$ = voltage across the load (resistive)

then

$$V_L{}' = V_L \sin \omega t$$

$$V_{CE} = V_s - \left(\frac{V_s}{2} + V_L \sin \omega t\right) = \frac{V_s}{2} - V_L \sin \omega t$$

$$I_C = \frac{V_L \sin \omega t}{R_L}$$

since

$$\overline{P_D} = \frac{1}{2\pi}\underbrace{(2)}_{\text{two transistors operated Class B (since both transistors are in the same IC package)}}\int_0^{\pi} p_d \, d(\omega t)$$

where: $\overline{P_D}$ = average power

p_d = instantaneous power

then

$$\overline{P_D} = \frac{1}{\pi}\int_0^{\pi}\left(\frac{V_s}{2} - V_L \sin \omega t\right)\left(\frac{V_L \sin \omega t}{R_L}\right) d(\omega t)$$

$$= \frac{V_s V_L}{2\pi R_L}\int_0^{\pi} \sin \omega t \, d(\omega t) - \frac{V_L{}^2}{2\pi R_L}\int_0^{\pi}(1 - \cos 2\omega t)\, d(\omega t)$$

$$= \frac{V_s V_L}{2\pi R_L}(2) - \frac{V_L{}^2}{2\pi R_L}(\pi)$$

$$= \frac{V_s V_L}{\pi R_L} - \frac{V_L{}^2}{2 R_L} \qquad (4.12.4)$$

Equation (4.12.4) is the average power dissipated; the *maximum* average power dissipated will occur for the value of V_L that makes the first derivative of Equation (4.12.4) equal to zero:

$$\frac{d(\overline{P_D})}{d(V_L)} = \frac{V_s}{\pi R_L} - \frac{V_L}{R_L} = 0 \text{ at maximum}$$

$$\therefore \; V_{Lp} = \frac{V_s}{\pi} \qquad (4.12.5)$$

Equation (4.12.5) is the peak value of V_L that results in max P_D; multiplying by two yields the peak-to-peak value for max P_D:

$$V_{Lp\text{-}p} = \frac{2 V_s}{\pi} = 0.637 V_s \qquad (4.12.6)$$

Substitution of Equation (4.12.5) into Equation (4.12.4) gives the final value for max P_D:

$$\max \overline{P_D} = \frac{V_s^2}{2\pi^2 R_L} \approx \frac{V_s^2}{20 R_L} \qquad (4.12.7)$$

Another useful form of Equation (4.12.7) is obtained by substitution of Equation (4.12.2):

$$\max \overline{P_D} = \frac{4}{\pi^2} P_O(\max) \qquad (4.12.8)$$

4.12.3 Application of Max P_D

Max P_D determines the necessity and degree of external heatsinking, as will be discussed in Section 4.14.

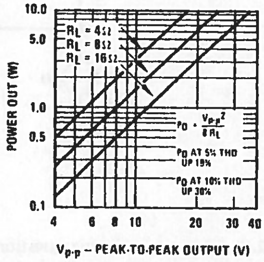


FIGURE 4.12.3 Power Out

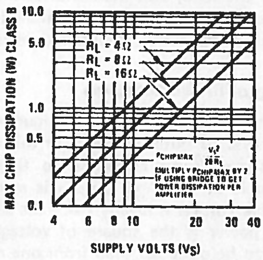
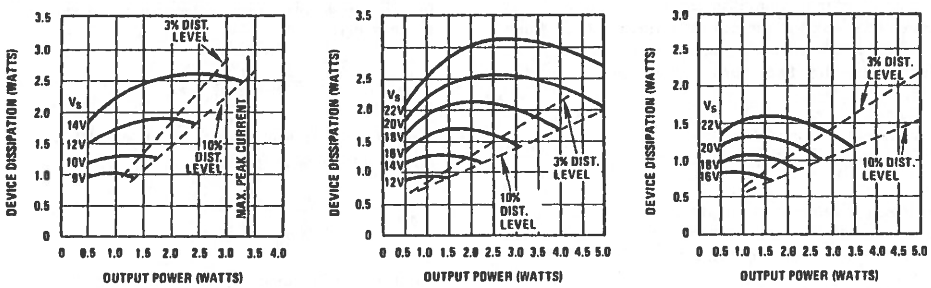


FIGURE 4.12.4 Max Chip Dissipation

The nomographs of Figures 4.12.3 and 4.12.4 make it easy to determine package power dissipation as well as output VI characteristics for popular conditions. Since part of the audio amplifier specmanship game involves juggling output power ratings given at differing distortion levels, it is useful to know that:

P_O increases by 19% at 5% THD
P_O increases by 30% at 10% THD



Device Dissipation vs. Output Power – 4Ω Load

Device Dissipation vs. Output Power – 8Ω Load

Device Dissipation vs. Output Power – 16Ω Load

FIGURE 4.12.5 Data Power Curves as Shown on Many Data Sheets

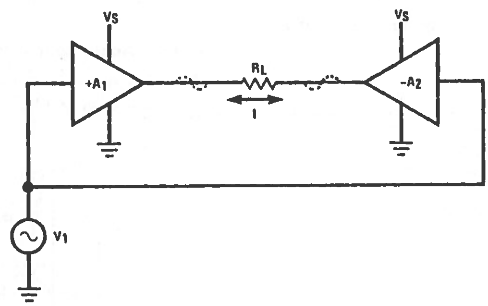


FIGURE 4.12.6 Bridge Audio

Equation (4.12.6) raises an intriguing question: If max P_D occurs at peak-to-peak output voltages equal to 0.637 times the power supply, will P_D go *down* if the output swing is *increased*? The answer is yes – indeed if an amplifier runs at $0.637V_s$ to the load, and then is driven harder, say to $0.8V_s$, it will cool off, a phenomenon implied in the power curves given on many audio amplifier data sheets (Figure 4.12.5).

4.12.4 Max P_D of Bridge Amplifiers

Bridge connecting two amplifiers as in Figure 4.12.6 results in a large increase of output power. In this configuration the amplifiers are driven antiphase so that when A_1's output voltage is at V_s, A_2's output is at ground. Thus the peak-to-peak voltage is ideally twice the supply voltage. Since output power is the square of voltage, four times more power can be obtained than from one of these same amplifiers run alone. Note, however, that since the peak voltage across the bridged load is twice that run as a single, the amplifiers must be capable of twice the peak currents. This, along with the fact that no real power amplifier can swing its output completely to V_s and ground, explains why actual bridge circuits never fully realize four times their single circuit output power.

Power dissipation in a bridge is calculated by noting that the voltage at the center of the load does not move. Thus, Equation (4.12.3) can be applied to half the load resistor:

$$P_{A1 \text{ or } A2} = \frac{V_s^2}{\pi^2 R_L} = \frac{V_s^2}{10 R_L} \qquad (4.12.9)$$

4.13 BOOSTED POWER AMPS

4.13.1 Introduction

When output power requirements exceed the limits of available monolithic devices, boosting of the output with two external transistors may be done to obtain higher power levels. The simplest approach involves adding a complementary emitter follower output stage within the feedback loop. The limiting factor is the limitation upon output voltage swing imposed by the B-E drop from the driver's output. Such designs cannot swing closer to the rail voltages than about one volt less than the IC's swing.

4.13.2 Output Boost with Emitter Followers

The simple booster circuit of Figure 4.11.1 allows power output of 10W/channel when driven from the LM378. The circuit is exceptionally simple, and the output exhibits lower levels of crossover distortion than does the LM378 alone. This is due to the inclusion of the booster transistors within the feedback loop. At signal levels below 20mW, the LM378 supplies the load directly through the 5Ω resistor to about 100mA peak current. Above this level, the booster transistors are biased ON by the load current through the same 5Ω resistor.

The response of the 10W boosted amplifier is indicated in Figure 4.13.2 for power levels below clipping. Distortion is below 2% from about 50Hz to 30kHz. 15W RMS power is available at 10% distortion; however, this represents extreme clipping. Although the LM378 delivers little power, its heat sink must be adequate for about 3W package dissipation. The output transistors must also have an adequate heat sink.

The circuit of Figure 4.13.3 achieves about 12W/channel output prior to clipping. Power output is increased because there is no power loss due to effective series resistance and capacitive reactance of the output coupling capacitor required in the single supply circuit. At power up to 10W/channel, the output is extremely clean, containing less than 0.2% THD midband at 10W. The bandwidth is also improved due to absence of the output coupling capacitor. The frequency response and distortion are plotted in Figures 4.13.4 and 4.13.5 for low and high power levels. Note that the input coupling capacitor is still required, even though the input may be ground referenced, in order to isolate and balance the DC input offset due to input bias current. The feedback coupling capacitor, C_1, maintains DC loop gain at unity to insure zero DC output voltage and zero DC load current. Capacitors C_1 and C_2 both contribute to decreasing gain at low frequencies. Either or both may be increased for better low frequency bandwidth. C_3 and the 27k resistor provide increased high frequency feedback for improved high frequency distortion characteristics. C_4 and C_5 are low inductance mylar capacitors connected within 2 inches of the IC terminals to ensure high frequency stability. R_1 and R_f are made equal to maintain $V_{OUTDC} = 0$. The output should be within 10 to 20mV of zero volts DC. The internal bias is unused; pin 1 should be open circuit. When experimenting with this circuit, use the amplifier connected to terminals 8, 9 and 13. If using only the amplifier on terminals 6, 7 and 2, connect terminals 8 and 9 to ground (split supply) to cause the internal bias circuits to disconnect.

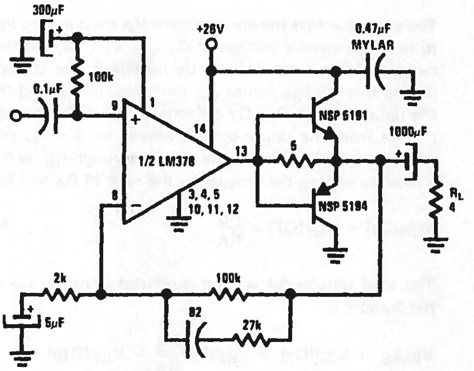


FIGURE 4.13.1 10-Watt Power Amplifier

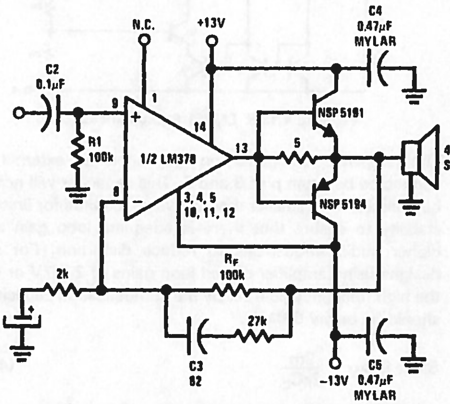


FIGURE 4.13.3 12-Watt Low-Distortion Power Amplifier

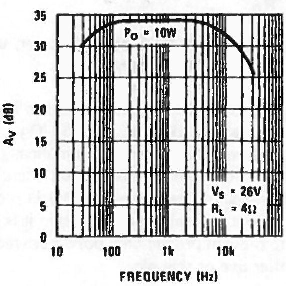


FIGURE 4.13.2 10-Watt Boosted Amplifier, Frequency Response

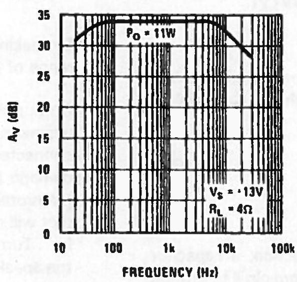


FIGURE 4.13.4 Response for Amplifier of Figure 4.13.3

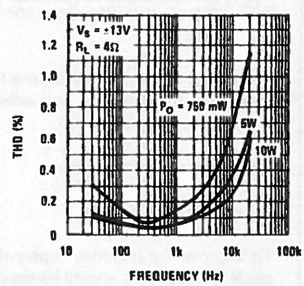


FIGURE 4.13.5 Distortion for Amplifier of Figure 4.13.3

4.13.3 Power Drivers

Using external transistors to boost the output of monolithic power amplifiers does increase the output power available, but with the limitation that the supply voltage cannot be increased beyond the maximum rating of the I/C. Also the output swing on that supply voltage is always less than the I/C output swing. To accomodate higher voltages and larger output swings, National has developed the Power Driver I/C's, the LM391 and LM2000 series. The LM391 is designed to drive power transistors in 10 watt to 50 watt power amplifiers operating from split supplies as high as ±50 volts. The LM2000 and LM2001 are designed for battery operated use to obtain 4 watts in 4Ω on a 12 volt supply and 2 watts in 2Ω on 6 volt supplies.

4.13.4 LM391 Circuit Description

An equivalent schematic for the LM391 is shown in Figure 4.13.6. A PNP differential input stage is used with emitter degeneration provided by 5kΩ resistors to give a good slew rate and a large linear input voltage range (see Section 4.1.2).

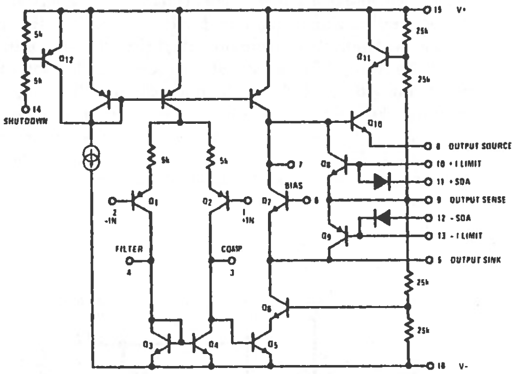


FIGURE 4.13.6 LM391 Equivalent Circuit

The amplifier compensation capacitor is external and connected between pins 3 and 5. This capacitor will normally be selected to be smaller than the value required for unity gain stability to ensure that there is adequate loop gain at the higher audio frequencies to reduce distortion. For stable designs using amplifier closed loop gains of 20V/V or more, the high frequency pole set by the compensation capacitor C_C should be below 500kHz.

$$\text{Since } f_H A_V = \frac{g_m}{2\pi C_C} \tag{4.13.1}$$

$$g_m = \text{transconductance of input stage} \approx \frac{1}{5.5 \times 10^3}$$

$$\therefore C_C \geqslant \frac{1}{2\pi 5.5 \times 10^3 \times 20 \times 500 \times 10^3} \quad (A_V = 20V/V)$$

$$\geqslant 3pF$$

The size of C_C will also determine the maximum possible slew rate. Since the largest current available to charge C_C is 100μA (input stage fully switched),

$$\text{Slew Rate} = \frac{I}{C_C} \leqslant \frac{100 \times 10^{-6}}{3 \times 10^{-12}} \tag{4.13.2}$$

i.e. Slew Rate ⩽ 33V/μS

To improve the negative supply ripple rejection, a capacitor equal in value to C_C should be connected from pin 4 to ground (Figure 4.13.9).

In order to accomodate different output stage configurations, the AB Bias circuit of the LM391 can be programmed externally. This type of bias circuit is known as a "V_{BE} Multiplier" and is set up as shown if Figure 4.13.10.

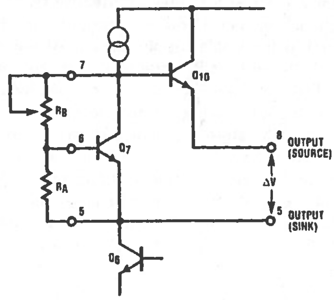


FIGURE 4.13.10 AB Bias Current Circuit

The voltage across the lower resistor R_A must always be equal to the base-emitter voltage of Q_7 . . . V_{BE} (Q_7). If the base current of Q_7 is assumed to be negligible, the current producing this voltage across R_A must also be flowing through the upper resistor R_B. Q_7 collector will absorb the additional current from the upper current source so that Q_7 collector base voltage is defined by the current through R_B, and will be a multiple of V_{BE} determined by the ratio of R_A and R_B.

$$V_{CB}(Q_7) = V_{BE}(Q7) \times \frac{R_B}{R_A} \tag{4.13.3}$$

The total differential voltage produced between the output pin 5 and 8 is

$$V_{BIAS} = V_{BE}(Q_7) + V_{BE}(Q_7)\frac{R_B}{R_A} - V_{BE}(Q_{10})$$

$$\text{If } V_{BE}(Q_7) = V_{BE}(Q_{10})$$

$$V_{BIAS} = V_{BE}(Q_7)\frac{R_B}{R_A} \tag{4.13.4}$$

By making R_B 3.9kΩ and R_A a 10kΩ potentiometer, a wide range of AB bias voltages can be obtained.

Output sink and source currents are guaranteed to be 5mA minimum, and the protection devices Q_8 and Q_9 can be connected to reduce these drive currents automatically when damage to the external output transistors could occur. Also for protection a thermal shut-down pin (Pin 14) is provided that will remove all output current capability when it is pulled low. Turn-on delay to prevent preamplifier pops from reaching the speakers is another use of this pin.

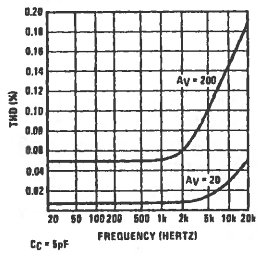


FIGURE 4.13.7 Total Harmonic Distortion $R_L = 8$

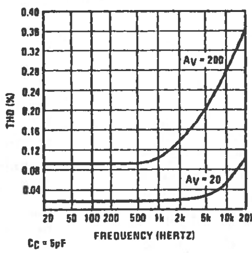


FIGURE 4.13.8 Total Harmonic Distortion $R_L = 4$

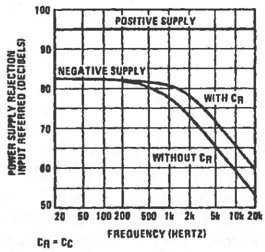


FIGURE 4.13.9 Input Referred Power Supply Rejection

4.13.5 Non-Inverting Amplifier Application

A typical amplifier set-up, without using the protection circuitry, is shown in Figure 4.13.11. R_{IN} provides a dc bias path for the input stage and sets the amplifier input resistance. A good value is 100kΩ. If very high resistor values are used for R_{IN}, layout induced oscillations will become probable and there will be larger dc offset voltages produced at the output. To minimize input bias currents producing output offsets, the feedback resistor R_{f1}, should be made equal to R_{IN}. R_{f1}, together with R_{f2} and C_F will set the amplifier mid-band gain A_V,

$$A_V = 1 + \frac{R_{f1}}{R_{f2}} \tag{4.13.5}$$

C_F reduces the amplifier gain to unity at dc for minimum offset voltage and gives a low frequency pole with R_{f2},

$$f_L = \frac{1}{2\pi R_{f2} C_F} \tag{4.13.6}$$

For amplifier gains of 20V/V and above, the recommended value for the compensation capacitor C_c is 5pF which, by rearrangement of Equation 4.13.1, will give an amplifier closed loop bandwidth of 320kHz. If C_R is used for improved ripple rejection, this also should be 5pF.

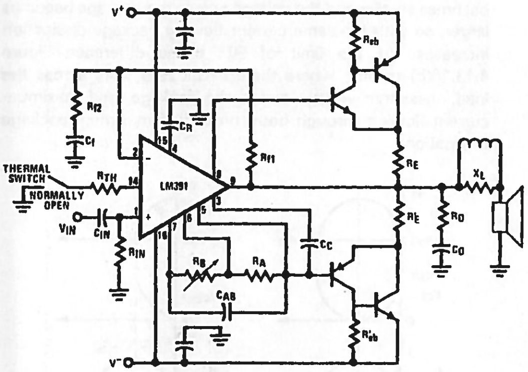


FIGURE 4.13.11 LM391 With External Components-Protection Circuitry not Shown

Connecting a 0.1μF capacitor across the AB Bias network will improve the transient response of the amplifier and reduce distortion at high frequencies. The AB bias current in the output stages should be set by R_B to above 20mA to ensure low THD (Figure 4.13.12)

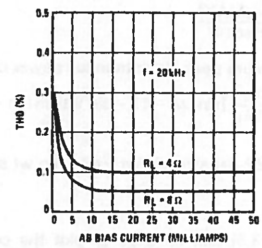


FIGURE 4.13.12 THD vs AB Bias Current

The resistor R_{TH} in series with pin 14 and a thermal shutdown switch determines the amount of current pulled from pin 14 during shut-down. Since this current shold not exceed 1mA, the value of R_{TH} is given by,

$$R_{TH} \geq \frac{V^+}{1 \times 10^{-3}} \tag{4.13.7}$$

4.13.6 Output Stage Stability (External Components)

The output stage of Figure 4.13.11 is a composite NPN/PNP arrangement. A resistor-capacitor network $R_o C_o$ is used to compensate the output power devices and a resistor R_{eb} is included to "bleed" off stored charge in these output devices caused by their large input diffusion and depletion capacitance. Between the output devices and the speaker an inductor X_L is placed to protect the amplifier from instabilities caused by driving capacitance loads.

Both output devices have a low valued resistor R_E to help maintain thermal stability of the AB bias current. When a power transistor goes through a power cycle, the chip temperature can change by many degrees with a corresponding change in the base-emitter voltage (the temperature coefficient k of the base-emitter voltage is typically −2mV/°C). It is unlikely that the V_{BE} multiplier providing the output bias voltage will be able to track the temperature change of the output device and the drop in output transistor V_{BE} will cause an increase in the AB Bias current and an increase in power dissipation. At very low frequencies these changes can occur during a single cycle of the output swing contributing to increased distortion levels. A more serious problem is that the AB Bias transistor junction (Q_7) can cool more rapidly than the output transistor junction following a period of sustained power output. Thermal instability can result if the then increased bias current causes a higher power dissipation level than that previously being sustained. Using R_E helps to maintain thermal stability since an increased bias current in the output stage will cause an additional voltage drop across R_E to compensate for the decrease in V_{BE}.

For an output stage of the type shown in Figure 4.13.11, a simple expression can be developed for the least value of R_E that will ensure thermal stability of the output stage,

$$R_E \geq \frac{\theta_{JA} V_{MAX} K}{(\beta + 1)} \tag{4.13.8}$$

where: θ_{JA} = Thermal Resistance of Driver Transistor (Junction to Ambient)

β = Minimum beta of the *output* device

V_{MAX} = Maximum supply voltage (V^+ or V^-)

K = 2mV/°C

4.13.7 Output Device S.O.A. Protection

For higher power output amplifiers, the Safe Operating Area (S.O.A.) of the output devices becomes important. Operating within the power-temperature ratings and avoiding thermal run away will not guarantee circuit reliability. To avoid failure the operating load line must be maintained within safe voltage-current limits and, particularly in the case of second breakdown which is energy dependent, time limits must be observed.

For a given device, reliable areas of operation are specified by an S.O.A. chart, of which Figure 4.13.13 is a typical example. The S.O.A. boundary can be defined by a few specific limits, current, power, second breakdown and voltage. Note that for high voltage and medium currents the available power is limited primarily by second breakdown considerations. A second important point to note is that the DC operating curve is based on a 25°C case temperature. Therefore the dc or low frequency operation is usually thermally limited to limits less than shown by this chart.

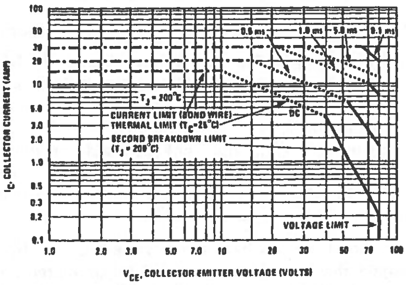


FIGURE 4.13.13 Active Region Safe Operating Area

Clearly the choice of output device as the power amplifier rating is increased in not a simple matter. Also, to keep the output stage compensation straight forward, the driver device is usually chosen to have a good frequency response, yet as the output power goes up the S.O.A. of these devices will come into play.

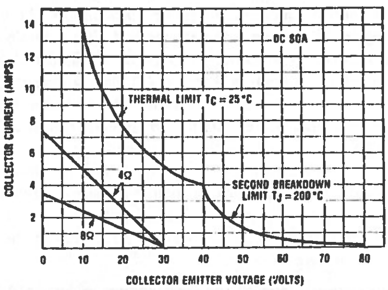


FIGURE 4.13.14 Amplifier Load Lines On S.O.A. Chart

If we add 8Ω and 4Ω load lines to the S.O.A. chart (Figure 4.13.14 — note the change in axes scale) which would represent *practical* 40 watt and 60 watt amplifiers respectively, we can see that the load lines are all safely within the S.O.A. of the device. So where is the problem, since a number of devices with a lot less S.O.A. than the example (2N5884/86) given here would do the job? The problem is that real world speakers are anything but resistive and this will cause a substantial increase in the heatsink and S.O.A. requirements than is indicated by a simple resistive load.

4.13.8 Effect of Speaker Loads

Figure 4.13.15 shows an impedance curve for a typical dynamic loudspeaker. As can be seen, there is a wide variation in impedance between 20 Hz and 20 kHz. The impedance at the resonant frequency can commonly measure five times or more the rated impedance. Indeed, many speakers will only display their rated impedance at one frequency (typically 400 Hz). The actual impedance is a complex value of dc resistance, inductive reactance of the voice coil, coupling capacitor reactance, crossover network impedance and frequency. In general, though, loudspeakers appear inductive with a worst case phase angle of 60°. This means that the voltage across the speaker leads the current by 60°.

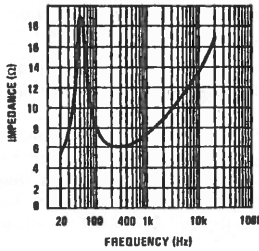


FIGURE 4.13.15 Impedance Curve for a Typical Dynamic Loudspeaker

An intuitive approach to what phase angle does to maximum average power dissipation produces the realization that the worst case load for power dissipation is purely reactive, i.e., 90° phase angle. This becomes clear by considering the resistive case of zero phase angle depicted in Figure 4.13.16(a) where the maximum voltage across the load, V_L, results in maximum current, I_L; but since they are in phase there exists zero volts across the device and no package disipation results. Now, holding everything constant while introducing a phase angle causes the voltage waveforms to shift position in time, while the current stays the same. The voltage across the load becomes smaller and the voltage across the package becomes larger, so with the same current flowing package dissipation increases. At the limit of 90° phase difference Figure 4.13.16(b) results, where there exists zero volts across the load, maximum voltage across the package, and maximum current flowing through both, producing maximum package dissipation.

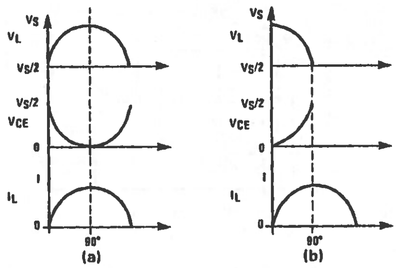


FIGURE 4.13.16 Phase Angle Relationship Between Voltage and Current (a) 0° and (b) 90°

If we consider the effect of load phase angle Φ, the formula obtained in Section 4.12.2 must be modified.
Equation (4.12.8) becomes:

$$\max \overline{P_D} = \frac{4P_{O(MAX)}}{\pi^2 \cos\Phi} \qquad (4.13.9)$$

and the maximum peak instantaneous power dissipation,

$$Pd_{(MAX)} = \frac{V_S^2}{4R_L}\,[\sin(wt-\Phi) - \sin wt \sin(wt-\Phi)]$$

or,

$$Pd_{(MAX)} = 2P_{O(MAX)}\,[\sin(wt-\Phi) - \sin wt \sin(wt-\Phi)] \qquad (4.13.10)$$

Equation (4.13.9) can be used to plot the curve shown in Figure 4.13.17 which gives the output stage maximum average power dissipation for reactive loads up to 60° phase angle. The importance of Figure 4.13.17 is seen by comparing the power ratio at zero degrees (0.405) with that at 60° (0.812) ... double! This means that the maximum Class B output stage dissipation can be *twice as much* for a speaker load as for a resistive load.

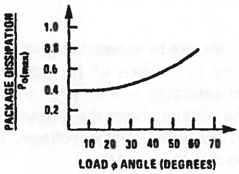


FIGURE 4.13.17 Class B Package Dissipation for Reactive Loads

The maximum power dissipation in *each* output device will be half that given by Equation (4.13.9)

$$\overline{P}_{D(MAX)} = \frac{2P_{o(MAX)}}{\pi^2 \cos\Phi} \quad (4.13.11)$$

and the maximum power dissipation for *each* driver transistor in a composite output stage is given by,

$$\overline{P}_{D(Driver)} = \frac{\overline{P}_{o(MAX)}}{\beta_{MIN}} \quad (4.13.12)$$

($\beta_{(MIN)}$ is that of the *output* device)

Equations (4.3.11) and (4.3.12) can be used to determine the heatsinking requirements for *each* device in the output stage.

$$\theta_{JA} \leqslant \frac{T_{JMAX} - T_{AMAX}}{P_{DMAX}} \quad (4.13.13)$$

The heatsink thermal resistance is given by,

$$\theta_{SA} \leqslant \theta_{JA} - \theta_{JC} - \theta_{CS} \quad (4.13.14)$$

where: T_{JMAX} is maximum transistor junction temperature

T_{AMAX} is maximum ambient temperature

θ_{JA} is thermal resistance junction to ambient

θ_{SA} is thermal resistance sink to ambient

θ_{JC} is thermal resistance junction to case

θ_{CS} is thermal resistance case to sink

For driver devices mounted on a common heatsink, the thermal resistance given by Equation (4.13.14) should be divided by the number of driver devices for the total heatsink requirement. A similar calculation can be made for the output transistors but it is worth noting that the output heatsinks may depend more on the ability or need for the amplifier to withstand a continuously shorted output (see Section 4.13.9).

The effect of a reactive load on the heatsink is easy to calculate from Equations (4.13.11) and (4.13.12). To understand the effect of reactive loads on the output device S.O.A. requirements we need to refer to Figure 4.13.18. This shows normalized curves for several reactive load lines up to 60° phase angle, and the locus of the maximum peak instantaneous power dissipation in the output transistors [given by Equation (4.13.10)]. Now it is apparent that with a 60° phase angle load, the load line will approach very closely the S.O.A. limits. In fact, if the S.O.A. limits for the 60 watt/4Ω amplifier are superimposed on Figure 4.13.18 (dashed line), the dc S.O.A. will actually *be exceeded* if the load angle increases beyond 40°!

The locus of $pd_{(MAX)}$ on Figure 4.13.18 can be used to specify the required device S.O.A. for the worst reactive load for which the amplifier will be designed. For example, with a 60° load angle the S.O.A. should be specified at a voltage of 0.59 (V^+ + V^-) with a current of 0.77 I_{MAX}. It is instructive to contrast this with 0.25 (V^+ + V^-) and 0.5 I_{MAX} for a purely resistive load!

4.13.9 Protection Circuits

If we assume that an output device can be selected with a sufficient S.O.A. to accomodate the worst load angle condition designed for, the next step is to protect this output device from abnormal or transient conditions that could place operation outside the S.O.A. — overload or short circuited outputs for example.

The simplest form of protection — current limiting — has already been made possible by the inclusion of the resistors R_E in the output stage.

During normal operation, the load current flowing through R_E does not produce a large enough voltage drop to turn on either Q_8 (positive side) or Q_9 (negative side). The limit current is reached when the voltage across R_E exceeds V_{BE} (Q_8 or Q_9). Then the protection transistors turn on and begin to bleed the available base current from the output stages to hold the output current at the limit level. Although the normal "on"

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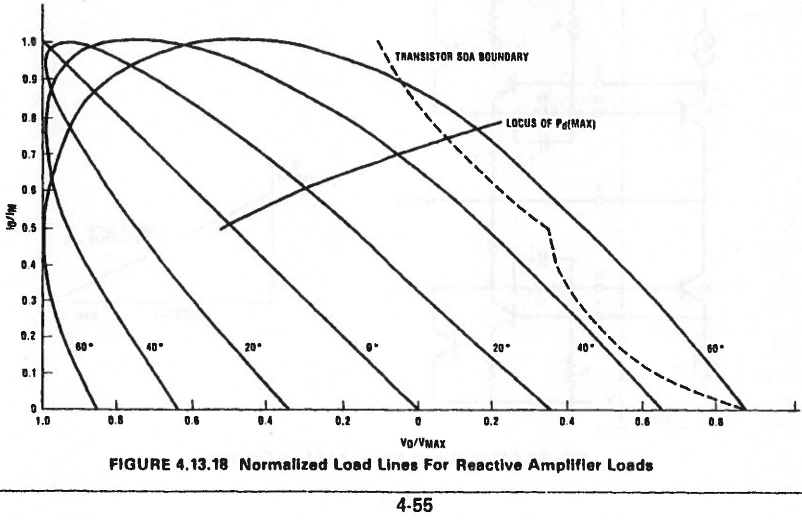

FIGURE 4.13.18 Normalized Load Lines For Reactive Amplifier Loads

voltage for Q_8 is 710mV and for Q_9 is 660mV at 25°C, equal turn-on voltages are assumed of 650mV at 55°. The load current limit is given by,

$$I_L = \frac{R_E}{0.65} \text{ AMPS} \tag{4.13.15}$$

Note that R_E also provides thermal run away protection for the AB Bias current. When R_E has been chosen for a given current limit a check should be made to see that R_E exceeds the value given by Equation (4.13.8).

Simple current limiting will not necessarily prevent the output devices from failing if a transient causes operation outside the device S.O.A. Also, during shorted conditions the average power being dissipated in each output transistor is given by,

$$\overline{P}_{D(SHORT)} = \tfrac{1}{2} I_{LIMIT} V_{CE}$$

or

$$\overline{P}_{D(SHORT)} = \frac{I_{LIMIT}(V^+ + V^-)}{4} \tag{4.13.16}$$

This power dissipation is substantially more than that obtained during normal operation and the heatsinks may not be able to handle this long term.

An improvement over current limiting is "single slope" load line protection (Figure 4.13.20). Two more resistors, R_1 and R_2 have been added to each output with a compensation capacitor C connected across R_2. Now the voltage across the output stage as well as the current through it is being monitored.

Now Q_8 and Q_9 operate to reduce the output current from the limit value set by the choice of R_E (obtained when either output device is saturated — i.e. V_{CE} = 0V) down to zero current when the voltage across the output device reaches the maximum rated collector-emitter voltage. In order for the current to be zero at $V_{CE} = V_M$,

$$V_{BE}(Q_8) = \frac{R_2}{(R_1 + R_2)} V_M$$

or

$$R_1 = \frac{R_2}{0.65}(V_M - 0.65) \tag{4.13.17}$$

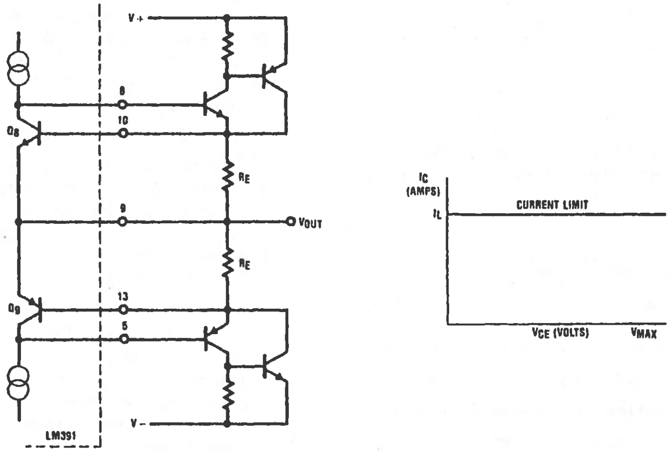

FIGURE 4.13.19 Use of LM391 To Limit Output Current

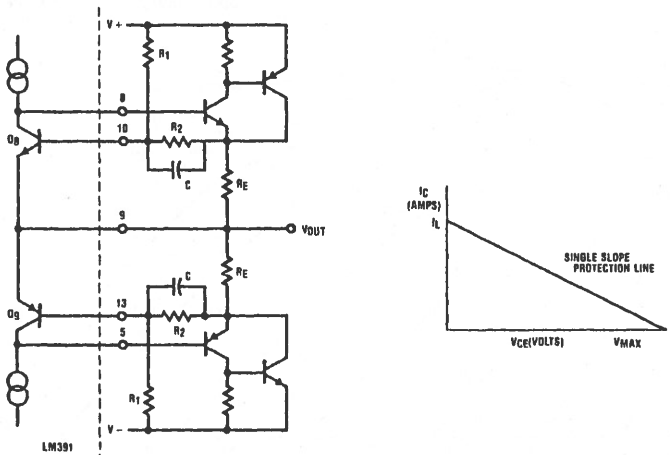

FIGURE 4.13.20 Single Slope Load Line Protection

With a shorted output, the average power dissipation in the output stage will be about half that obtained when only current limiting is used. However, if I_L is set close to the maximum load current for rated output (I_M), inspection of the reactive load lines given in Figure 4.13.18 will show that there is a good chance that the protection circuit will be activated even though the S.O.A. is not being exceeded. I_L can be increased above I_M until the protection line is asymptotic to the nearest point of the S.O.A. boundary (or to the maximum current rating of the device in some instances). This will cause a corresponding increase in the short circuit power dissipation.

R_2 is usually arbitrarily chosen to be 1kΩ leaving R_1 to be defined by Equation (4.13.17). A good choice for C is 1000pF.

To permit operation over most of the output transistor S.O.A. without activating protection circuits, dual slope load line protection is recommended, Figure 4.13.21. The corresponding protection lines superimposed on a typical transistor S.O.A. chart are shown in Figure 4.13.22.

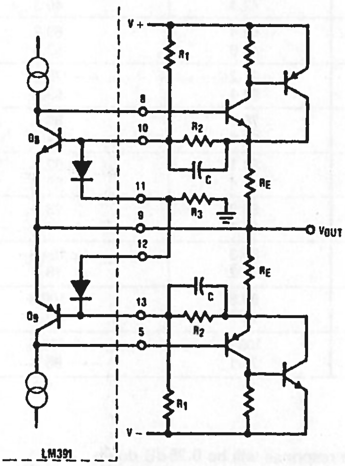

FIGURE 4.13.21 Dual Slope Load Line Protection

The internal diodes on the protection transistor bases are connected through a resistor R_3 to ground. R_E, R_1, R_2 and C are selected as before in the single slope protection circuit and with R_3 connected to ground, the break point where the protection line changes slope will be at the midpoint between the supply rails (equal to V^+ or V^- as far as the output devices are concerned).

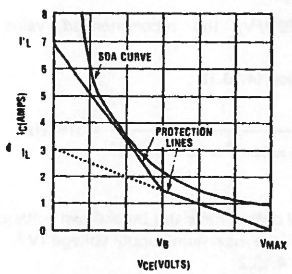

FIGURE 4.13.22 Dual Slope Load Lines on S.O.A. Chart

The design formula can be determined by assuming the output transistor is saturated and delivering the new upper limit current I'_L. The voltage across R_2, R_3 and the corresponding diode will be approximately V^+ (or V^-) and the current through R_2 is given by,

$$i = \frac{I'_L R_E - V_{BE}}{R_2}$$

Therefore,

$$R_3 = R_2\left(\frac{V+}{I'_L R_E - 0.65} - 1\right) \quad (4.13.18)$$

Assuming $V+ \gg V_{SAT}$

$V+ \gg V_{BE}$

Again the short circuit power dissipation will be half that obtained with simple current limiting, but the device can now operate over most of the S.O.A.

For convenience, Table 4.13.1 summarizes the necessary formula to determine component values for any degree of protection.

4.13.10 Power Supply Requirements

The power supply voltage and current capability depends on the amplifier rated ouput and load impedance. For a given power output P_o and load R_L,

$$V_{OPEAK} = \sqrt{2R_L P_O} \quad (4.13.19)$$

$$I_{OPEAK} = \frac{\sqrt{2P_O}}{R_L} \quad (4.13.20)$$

To obtain these output swings the power supply voltage will have to be higher to allow for transistor saturation voltage drops. Since, in a large number of cases, an unregulated supply will be used for economic reasons, the unloaded supply voltage will be about 15% higher than when delivering the rated current output. If we allow an additional 10% for high line conditions, the maximum supply voltage is given by,

$$MAXV_{SUPPLY} = \pm(V_{oPEAK} + V_{SAT}) \times 1.15 \times 1.1 \quad (4.13.21)$$

The requirement that rated power output be obtained with low line conditions will add another 10% to the number obtained from Equation (4.13.21). Assuming that transistor saturation and protection circuit voltage drops total 5V (per side), Table 4.13.2 lists the voltage and current requirements for 20 watt through 100 watt amplifiers operating on unregulated supplies. The final column of Table 4.13.2 lists the collector breakdown voltage requirement for the output transistors (including the drivers). This column will also define the breakdown voltage required of the LM391.

4.13.11 Amplifier Design

Although the preceding text may imply that power amplifier design is fraught with pitfalls, the following examples should illustrate how an amplifier can be designed using the LM391 in a very straightforward fashion.

Example 4.13.1

Design an amplifier capable of delivering an average power of 20 watts into 8Ω and 30 watts into 4Ω. The input sensitivity should be lower than $1V_{MAX}$ with an input impedance of 100kΩ or more. A 20Hz to 20kHz ± 0.25dB bandwidth is required.

4

TABLE 4.13.1 Protection Circuit Formula

TYPE OF PROTECTION	R_E	R_1	R_2	R_3	C
CURRENT LIMIT	$R_E = \frac{0.65}{I_L}$		SHORT		
SINGLE SLOPE PROTECTION LINE	$R_E = \frac{0.65}{I_L}$	$R_1 = \frac{R_2(Vm - 0.65)}{0.65}$	1kΩ		1000PF
DUAL SLOPE PROTECTION LINE	$R_E = \frac{0.65}{I_L}$	$R_1 = \frac{R_2(Vm - 0.65)}{0.65}$	1kΩ	$R_3 = R_2\left\{\left(\frac{V+}{I'_L R_E - 0.65}\right) - 1\right\}$	1000pF

TABLE 4.13.2 Power Supply Requirements For 20 Watt To 100 Watt Amplifiers

P_o POWER(WATTS)	LOAD (Ω)	IpAMPS $(\sqrt{2P_o/R_L})$	OUTPUT SWING $(\pm\sqrt{2P_oR_L})$ Volts	TOTAL SUPPLY VOLTAGE REGULATED	TOTAL SUPPLY VOLTAGE UNREGULATED (NO LOAD)	TRANSISTOR V_{CEO} (SUS) (UNREGULATED SUPPLY VOLTAGE)
20	8	2.24	±17.9	47.1	54.1	59.6
	4	3.16	±12.6	36.6	42.1	46.3
30	8	2.74	±21.9	55.1	63.4	69.7
	4	3.87	±15.5	42.3	48.6	53.5
40	8	3.16	±25.3	61.9	71.2	78.3
	4	4.47	±17.9	47.1	54.1	59.6
50	8	3.54	±28.3	67.9	78.1	85.9
	4	5.00	±20.0	51.3	59.0	64.9
60	8	3.87	±31.0	73.1	84.1	92.5
	4	5.48	±21.9	55.1	63.4	69.2
70	8	4.18	±33.5	78.2	89.9	98.9
	4	5.92	±23.7	58.6	67.4	74.2
80	8	4.47	±35.8	82.9	95.3	104.8
	4	6.32	±25.3	61.9	71.2	78.3
90	8	4.74	±37.9	85.9	98.8	108.6
	4	6.71	±26.8	65.0	74.7	82.2
100	8	5.00	±40.0	91.3	100.5	115.5
	4	7.07	±28.3	67.9	78.1	85.9

Solution.

1. From Table 4.13.2:

   Voltage swing for rated power in 8Ω = ±17.9 volts.

   Peak current for rated power in 4Ω = 3.87 amps

   For an unregulated supply, no load voltage

   V_{SUPPLY} = 54 volts or ±27 volts.

2. $A_V \geq \frac{17.9/\sqrt{2}}{1} = 12.66$

   If we use an LM391 with a gain of $A_V = 20V/V$ the resulting sensitivity is 630mVrms which is well within the required specification.

3. Letting $R_{IN} = 100k$ gives the required input impedance, and to ensure low dc offset voltages, $R_{F1} = 100k$. From Equation (4.13.5)

   $$A_V = 20 = 1 + \frac{R_{f1}}{R_{f2}}$$

   $\therefore R_{f2} = 5.26k\Omega$

   Put $R_{f2} = 5.1k\Omega$

4. Two octaves below the high frequency pole (f−3dB), the amplitude response will be 0.25dB down,

   i.e. $f_H \geq 20 \times 10^3 \times 4 = 80kHz$

   Similarly two octaves above the low frequency pole, the response will again be −0.25dB,

   i.e. $f_L \leq \frac{20}{4} = 5Hz$

   From Equation (4.13.6)

   $$C_F \geq \frac{1}{2\pi f_L R_{f2}} \geq 6.2\mu F$$

   Use $C_F = 10\mu F$

   For $A_V = 20V/V$, the recommended value for C_c is 5pF

   From Equation (4.13.1)

   $$f_H = \frac{1}{2\pi \times 5 \times 10^{-12} \times 20 \times 5 \times 10^3} = 318kHz$$

5. The I/C and output transistor breakdown voltages must be greater than the maximum supply voltage $(V^+ + V^-)$. From Table 4.13.2

   V_{MAX} = 59.6 volts (unregulated high line)

   Use LM391N-60

6. A suitable power transistor complementary pair is the National BD346 and BD347 (2N6487, 2N6490) with a V_{CEO} (sus) of 60 volts and a minimum beta of 30 at 4 amps. Since the guaranteed minimum drive current from the LM391-60 is 5mA, the driver transistors must have a minimum beta given by,

$$\text{Driver } \beta_{MIN} \geq \frac{3.87}{30 \times 5 \times 10^{-3}} \approx 26 \text{ (@ 130mA)}$$

The National complementary pair BD344, 345 (MJE171, MJE181) are 60V devices with a minimum beta of 40 at 200mA.

7. For each output transistor the maximum average power dissipation is given by Equation (4.13.11)

$$P_{d(MAX)} = \frac{2P_{D(MAX)}}{\pi^2 \cos\Phi}$$

Assume Φ = 60°MAX

$$P_{d(MAX)} = \frac{2 \times 30}{\pi^2 \times 0.05} = 12.2 \text{ watts}$$

8. From Equation (4.13.13)

$$\theta_{JA} \geq \frac{150°C - 55°C}{12.2W} \text{ for } T_A = 55°C$$

$$= 7.8°C/W$$

From Equation (4.13.14)

$$\theta_{SA} \leq 7.8 - 2.1 - 1.0 = 4.8°C/W$$

If both transistors for one amplifier are mounted on a single heatsink.

$$\theta_{SA} \leq 2.4°C/W$$

9. The driver maximum average power dissipation is given by Equation (4.13.12)

$$P_{D(Driver)} = \frac{\overline{P}_{DMAX}}{\beta_{MIN}} = \frac{12.2W}{30} = 410mW$$

Using Equation (4.13.13) again

$$\theta_{JA} \leq \frac{155°C - 55°C}{0.410} = 244°C/W$$

The free-air thermal resistance of the BD344, 345 is 100°C/W so that no additional heatsinking is required.

10. The least value of R_E to prevent AB bias thermal runaway is obtained from Equation 4.13.8

$$R_E \geq \frac{100(30) \times 2 \times 10^{-3}}{(30 + 1)} = 0.19\Omega$$

11. Figure 4.13.23 is the S.O.A. chart for the BD346 and BD347 transistors. Also shown are the 4Ω and 8Ω load lines and desired protection lines for dual slope protection. From Figure 4.13.23

V_m = 60 volts; V_B = 23 volts; I_L = 3 amps;

I'_L = 7 amps

From Table 4.13.1

$$R_E = \frac{0.65}{3} = 0.22\Omega$$

(This value of R_E also satisfies Equation 4.13.8, see Step 10)

The completed amplifier schematic is shown in Figure 4.13.24. One final point, the heatsink capability calculated in Steps 8 and 9 were for continuous operation into a 4Ω load. If that load is inadvertently shorted, then the average power being dissipated is given by Equation (4.13.16)

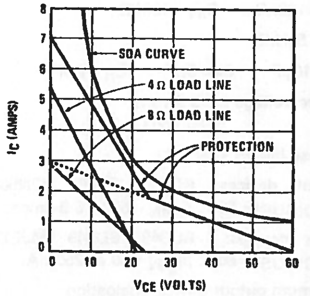

FIGURE 4.13.23 DC-S.O.A. for BD346 and BD347

$$P_{D(SHORT)} = \frac{I_{LIMIT}(V^+ - V^-)}{4}$$

I_{LIMIT} is obtained from Figure 4.13.23 and in this case is 1.8 amps.

$$\overline{P}_{D(SHORT)} = \frac{1.8(56)}{4} = 25.2 \text{ watts}$$

Sustained operation under shorted conditions will require a much larger output device heatsink or a thermal sensor to pull down Pin 14 of the LM391 when the original heatsink temperature exceeds 111°C.

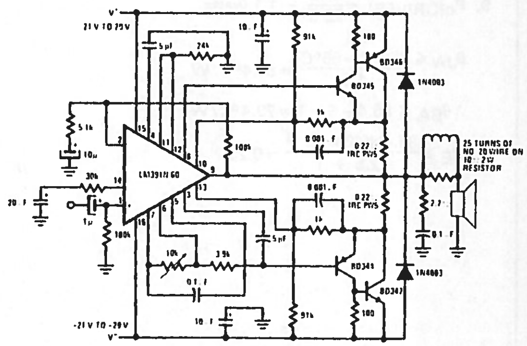

FIGURE 4.13.24 20W-8Ω, 30W-4Ω Amplifier

Example 4.13.2

Design an audio power amplifier with an input sensitivity of $1V_{rms}$, to drive 8Ω and 4Ω loads to power levels of 40 watts and 60 watts respectively. The maximum load phase angle is 60° and the design should include S.O.A. protection for the output stage.

Solution

Following the same steps as in the previous example:

1. I_{PEAK} = 5.48 amps

V_{PEAK} = ±25.3 volts

V_{SUPPLY} = 71.2 volts (No Load)
= ±31 volts (Full load)

2. Put A_V = 20V/V

Input sensitivity = $900mV_{rms}$

3. $R_{f1} = 100k\Omega$ $\quad R_{IN} = 100k\Omega$

$R_{f2} = 5.1k\Omega$

4. $C_F = 10\mu F$ $\quad C_c = 5pF$ $\quad C_R = 5pF$

5. Device voltage $\geq$ 78.3 volts
   = 80 volts

   Choose LM391N-80

6. Output devices: BD350, BD351 (2N5880, 2N5882), $V_{CEO(SUS)} = 80V$, $\beta_{MIN} = 50$ at 5.5 amps.

   Driver devices: BD348, BD349 (MJE172, MJE182), $V_{CEO(SUS)} = 80V$, $\beta_{MIN} = 50$ at 250mA.

7. Maximum output device dissipation

   $P_{o(MAX)}$ = 24.3 watts

8. $\theta_{JA} \leq \dfrac{200°C - 55°C}{24.3} = 6°C/W$

   $\theta_{SA} \leq 6 - 1.1 - 1.0 = 3.9°C/W$

   For both devices on a common heatsink

   $\theta_{SA} \leq 2.0°C/W$

   It is worth noting and comparing the heatsinking for this amplifier with that of the previous example. Using TO-3 case style transistor with higher junction temperature and lower thermal resistance has kept the heatsink size down more than might be expected. (Example 4.13.1 Specified Case Style TO-220)

9. $\overline{P}_{o(DRIVER)} = \dfrac{24.3}{22.5} = 1.1$ watts

   $\theta_{JA} \leq \dfrac{150°C - 55°C}{1.1} = 86.4°C/W$

   $\therefore \theta_{SA} \leq 86.4 - 6 - 1 = 79.4°C/W$

10. $R_E \geq \dfrac{86.4(40)2 \times 10^{-3}}{22.5 + 1} = 0.29\Omega$

11. Using Figure 4.13.25

   V_m = 80 volts, V_B = 47 volts, I_L = 3 amps, I'_L = 11 amps

   For I_L = 3 amps, from Table 4.13.1

   $R_E = \dfrac{0.65}{3} = 0.22\Omega$

This does not simultaneously satisfy Equation 4.13.8 — Step 10. Recalculating driver device heatsink from Equation 4.13.8

$\theta_{JA} \leq \dfrac{0.22(22.5 + 1)}{40(2 \times 10^{-3})} = 65°C/W$

$R_2 = 1k\Omega$ (arbitrary)

$R_1 = 1 \times 10^3 \dfrac{(80 - 0.65)}{0.65} \approx 120k\Omega$

Since to obtain the desired protection lines, V_B is not centered between the supply rails R_3 is replaced with a resistive divider (R_AR_B) between the positive supply and ground for the lower output device (Pin 12). A similar divider is connected at Pin 11 between the negative supply and ground for protection of the upper output device.

Now $R_A \parallel P_B = R_3$

$\therefore R_A \parallel R_B = 10^3 \left\{\dfrac{47}{11(0.22) - 0.65}\right\} - 1 = 25.55k\Omega$

Since V_B is 17 volts away from the center of the output swing (with supply loaded to ±31 volts)

$\dfrac{R_A\ 31}{(R_A + R_B)} = 17$

$\therefore R_B = 0.82\ R_A$

a) guess R_A = 62k
   $\therefore R_B$ = 51k, $R_A \parallel R_B$ = 28kΩ

b) guess R_A = 56k
   $\therefore P_B$ = 45.92 Put R_B = 47kΩ
   $\therefore R_A \parallel R_B$ = 25.55k which is close enough.

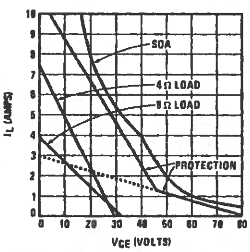

FIGURE 4.13.25 S.O.A. for BD350, and BD351

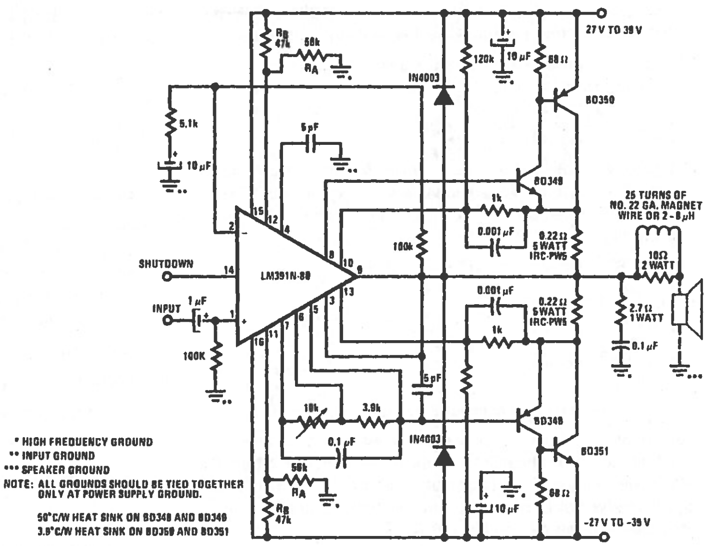

FIGURE 4.13.26 40W-8Ω, 60W-4Ω Amplifier

The complete amplifier is shown in Figure 4.13.26. Two additional diodes are shown clamping Pin 9 to within a diode drop greater than either supply. This is to prevent the output devices being damaged when the output voltage exceeds either supply — which can occur if the protection circuitry is activated while the load appears inductive.

Under shorted output conditions the limit current will be 4.5amps and the short circuit power dissipation is given by,

$$\bar{P}_{D(SHORT)} = \frac{4.5 \times 60}{4} = 67.5 \text{ watts}$$

Again, the heatsink capability caculated in Steps 8 and 11 will not permit a continuous short on the output. However, for normal music inputs into typical speaker loads, these heatsinks are actually quite conservative and will have the thermal capacity to ride out intermittent shorts of limited duration. Where a designer has a specific knowledge of the load and operating conditions for his amplifier, smaller heatsinks may be used for economic reasons. Even so, especially when smaller output devices are selected by the same reasoning, any amplifier should be thoroughly tested for reliability under actual operating conditions.

4.13.12 Oscillations and Grounding

Most power amplifiers will work the first time they are turned on. They also tend to oscillate, sometimes with catastrophic results, and have excess THD. The majority of oscillation problems are caused by inadequate power supply bypassing and by ground loops (see Section 2.2.2 and 2.2.3). 10µF capacitors on the supply leads, close to the circuit rather than to the power supply, will stop supply related oscillations. If the signal ground is used for these bypass capacitors, the THD will probably be further increased. To avoid this, the signal ground must return to the power supply alone, as should the output load or speaker ground. The bypass capacitor, output R-C and protection grounds can be connected together. Figure 4.13.26 shows the recommended grounding arrangement for power amplifiers using the LM391.

4.13.13 Turn-On Delay

It is often desirable to delay the turn-on of the power amplifier so that turn-on pops generated in the pre-amplifier section do not go to the speakers. This can be achieved with the LM391 simply by using the shutdown pin (Pin 14). A series capacitor-resistor combination is used to set the turn-on delay (Figure 4.13.27). At turn-on, the capacitor is at ground potential, holding Pin 14 low through the resistor and there is no current drive available for the output stage. After approximately two time constants, the capacitor has charged sufficiently that output drive current is enabled and normal amplifier operation can take place. The minimum value for the resistor is given by Equation (4.13.7)

$$R = \frac{V_+(\text{max})}{10^{-3}} \tag{4.13.7}$$

For an amplifier with ±30 volts supplies

$$R = \frac{30}{10^{-3}} = 30\text{k}$$

Turn-on delay in seconds is given by

$$T = 2RC \tag{4.13.22}$$

If we use a 33kΩ resistor

$$C = \frac{1}{2 \times 33 \times 10^3} = 15\mu\text{F}$$

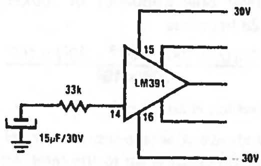

FIGURE 4.13.27 Turn-On Circuit Delay

4.13.14 Transient Distortion

The topic of transient distortion is one that is still subject to a great deal of discussion at this time. Nevetheless several design criteria have been evolved for the avoidance of distortion effects produced by transients in the program material. This section will discuss two of these criteria with respect to LM391 amplifiers.

Slew rate limits have been mentioned several times with respect to the compensation capacitors used in monolithic audio amplifiers (Sections 1.2.1, 4.1.1, etc.). Simple expressions have been developed for the frequency at which slew rate limiting will occur for a given output voltage swing and amplifier slew rate. What has not been emphasized is that at that frequency, distortion of the signal has already begun — of the order of 1% to 3% THD. To minimize distortion, the frequency at which slew rate limiting occurs (with maximum output swing) should be well above the audio bandwidth. Current practice indicates that a slew rate of 0.5V/µS *per output peak volt* is acceptable, with 1V/µS per output peak volt being conservative. The LM391 has a slew rate of 20V/µS with a 5pF compensation capacitor — reference to Table 4.13.2 shows that this is adequate for amplifiers up to 100 watts.

It may not be possible to avoid slew induced distortion simply by being able to slew at frequencies substantially above the audio bandwidth. If an input transient causes the amplifier input stage to overload, the output will be in slew limiting until the feedback loop responds. This can be prevented by using a low pass filter at the input stage. The cut-off frequency of this filter must be above the audio bandwidth, and how far above will depend on the amplifier input stage dynamic range and transconductance, and the open loop pole frequency of the amplifier. A detailed paper by Peter Garde in the May, 1978 *Journal of the Audio Engineering Society* derives the following criterion to prevent input stage overload.

$$\frac{I}{g_m} \geqslant V_{IN}\frac{(2f_f - f_o)K}{f_c} \tag{4.13.23}$$

where: V_{IN} = maximum peak input voltage to the amplifier

I = input stage maximum current

g_m = input stage transconductance

f_o = open loop pole frequency

f_c = closed loop pole frequency

f_f = input filter pole frequency

K = constant(dependent on ratio of f_c/f_f)

For the LM391; $f_o = 1\text{kHz}$, $g_m = \dfrac{1}{5.5 \times 10^3}$, $I = 100\mu\text{A}$. If the closed loop gain is 20V/V with C_c = 5pF as in the previous design examples, the closed loop pole frequency is 300kHz. In a 40 watt, 8Ω amplifier, the maximum input voltage

$$V_{IN} = \frac{25.3}{20} = 1.3 \text{ volts peak}$$

For an input filter pole frequency of 100kHz, K = 0.6 and Equation 4.13.23 becomes

$$\frac{100 \times 10^{-6}}{0.18 \times 10^{-3}} \geqslant \frac{1.3(2 \times 100 \times 10^{-3} - 10^{3}) \times 0.6}{300 \times 10^{3}}$$

or $0.56 \geqslant 0.52$ which is so.

Figure 4.13.28 shows a simple input filter for the LM391 to prevent input stage clipping up to the rated amplifier output.

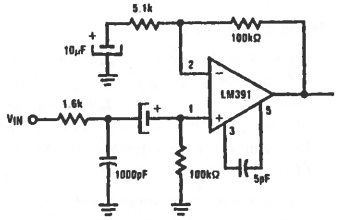

FIGURE 4.13.28 LM391 Input Filter

4.13.15 Low Voltage Power Drivers

High voltage power amplifiers require careful selection of the output transistors in order for them to be able to handle the power requirements. At the other end of the scale with low voltage, battery operated equipment, the concern is to obtain sufficient output swing into the available load impedance. The LM2000 and LM2001 amplifiers are designed to drive low cost external transistors to within a collector saturation voltage drop of the supply rails. For operation from 12 volts down to 2.5 volts the LM2000 is recommended, and for operation below 6 volts down to 1.8 volts the LM2001 is the device to use. Both have similar circuit configurations except for slight differences in the output stage. The output stage gain setting resistors are external for the LM2000 because of the higher levels of power dissipation.

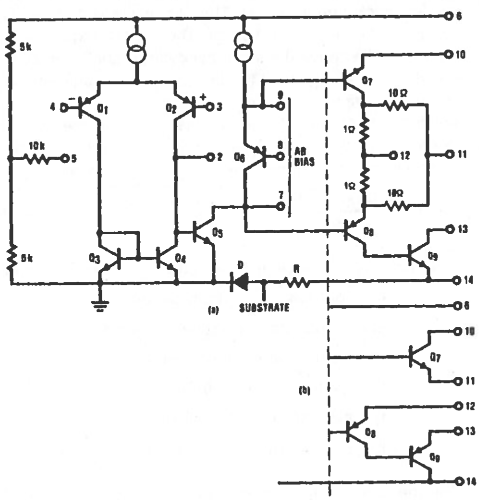

FIGURE 4.13.29 LM2000/2001 Equivalent Schematic
(a) LM2001
(b) LM2000

An equivalent schematic for both amplifiers is shown in Figure 4.13.29 (a) & (b). A fully differential PNP input stage with active load is used and a half supply voltage bias point (Pin 5) is provided to set up the output midway between the positive supply and ground. The external compensation around Q_5 (Pins 2 and 7) and a V_{BE} multiplier Q_6 for the output stage AB Bias (Pins 9, 8 and 7) are similar to those already described for the LM391 (Section 4.13.1). Both amplifiers have output driver stages designed for voltage gain and current drive to the external transistors.

4.13.16 Output Stage Operation — Upper Side

Referring to Figure 4.13.30, the external potentiometer R_B in the collector circuit of Q_6 is adjusted to set the current level in the driver transistor Q_7. The local feedback resistors R_2 and R_1 set the output stage voltage gain at,

$$A_{V(OUTPUT)} = 1 + \frac{R_2}{R_1} \tag{4.13.24}$$

This voltage gain allows Q_{10} to be driven into saturation without Q_7, Q_6 or the bias current source also saturating. A_V is determined externally for the LM2000 and fixed at 11 internally for the LM2001. Notice that there is no bleed resistor in the Q_{10} base circuit so that the collector current of Q_7 (I_{C7}), set by R_B, is also the base current of the output device Q_{10}. This means that the AB Bias current is $\beta \times I_{C7}$ where β is the dc beta of Q_{10}. Now the AB Bias current thermal stability depends on the T.C. of the output device β rather than its V_{BE}.

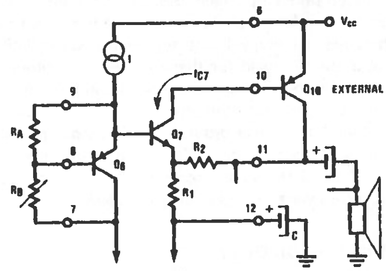

FIGURE 4.13.30 LM2001 Upper Side Driver Stage

4.13.17 Output Stage Operation — Lower Side

To enable the lower external transistor to be driven into saturation, the output load coupling capacitor is used to bootstrap the lower side output driver, which is isolated by the resistor R from the substrate (Diode D clamps the substrate when the output swings above ground). Figure 4.13.31.

4.13.18 Inverting Amplifier Applications

Figure 4.13.32 demonstrates a typical use of the LM2001 driving a 2Ω load through two PNP output transistors. The mid-band voltage gain is set externally by the ratio of R_{f1} and R_{f2} to 101, with C_F reducing the dc gain to unity to minimize output voltage offsets and giving a low frequency pole at $1/2\pi \times C_F R_{f2}$. R_{IN} connected to Pin 5 establishes the output dc bias at half supply and sets the input resistance of the amplifier. R_B adjusts the output stage AB Bias current to about 15mA.

The compensation capacitor C_C is selected on the basis of Figure 4.13.33 which has curves of the ±3dB bandwidth for values of C_C and amplifier closed loop gain. Since the bandwidth and C_C are inversely proportional (Equation 4.13.1), for a given closed loop gain, other curves for different bandwidths are easily extrapolated. For example, for a gain of 100, and a bandwidth of 50kHz, C_C will be 20/50 × 120pF, or approximately 50pF.

An additional resistor R_C can be added between Pin 8 and the inverting input to improve the amplifier response when the output stage is recovering from clipping. If the input level is sufficiently high that the output external transistors are driven into clipping, the amplifier will momentarily be open loop. The internal node at Pin 8 will come out of overload before the external devices (since the output driver stage has a gain of 11) and the feedback loop via R_C allows the output to make a much smoother transition from clipping than it would otherwise.

A large valued capacitor is connected at Pin 12 to enable the LM2001 driver transistors to sink large currents at low frequencies. At least 100mA sink current guarantees that over 1amp can be delivered to the load with external transistors having a forced beta of 10 or better.

In the case of the LM2000 a similar circuit hook-up is used with the output driver stage gain set resistors external. Operation to 12V supplies is permissible (see Figure 4.13.34).

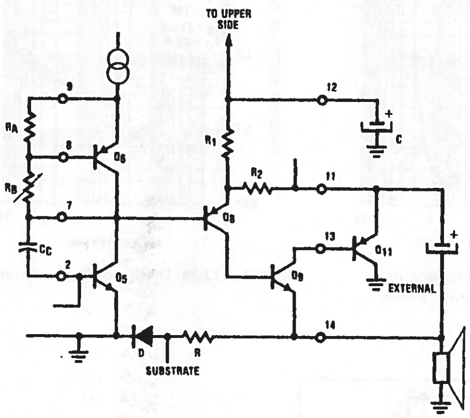

FIGURE 4.13.31 LM2001 Lower Side Driver Stage

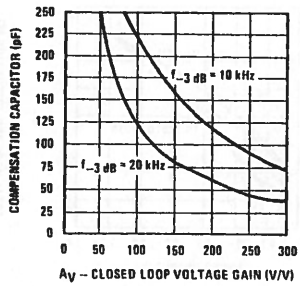

FIGURE 4.13.33 Compensation Capacitance and Closed Loop Bandwidth

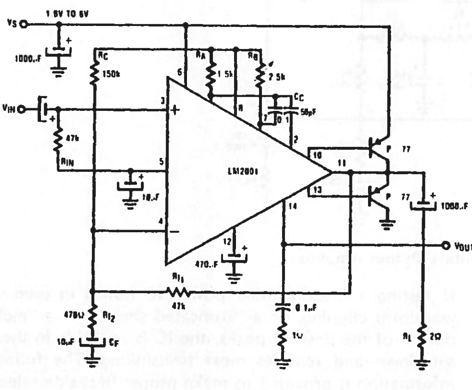

FIGURE 4.13.32 LM2001 Inverting Amplifier

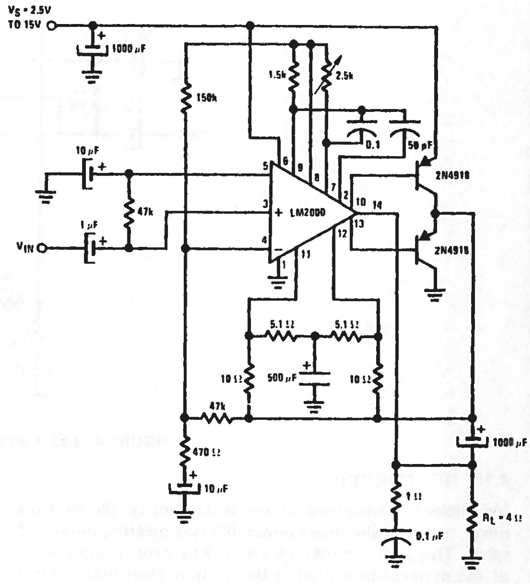

FIGURE 4.13.34 LM2000 Power Amplifier

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Typical performance characteristics for the amplifiers are summarized in Figures 4.13.35 through 4.13.38.

4.13.19 Complementary Output Stages

Both amplifiers can be used to drive complementary external transistors by reconfiguring the lower side output driver stage (Figure 4.13.39). The NPN output device is driven from the emitter of Q_9 with Q_9 collector being connected to the positive supply rail.

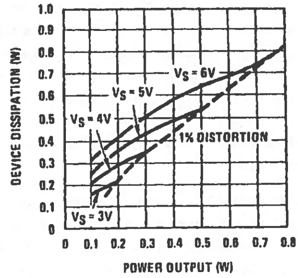


FIGURE 4.13.35 Device Dissipation — 4Ω Load LM2001 Only

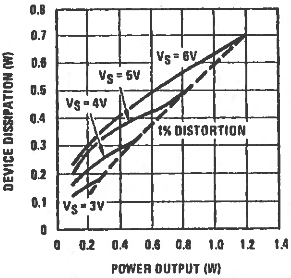


FIGURE 4.13.36 Device Dissipation — 2Ω Load LM2001 Only

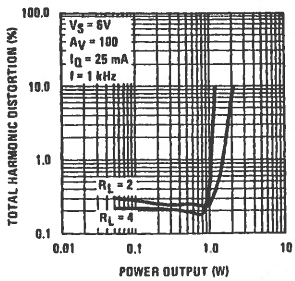


FIGURE 4.13.37 Distortion vs. Output Power

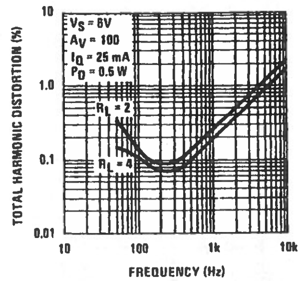


FIGURE 4.13.38 Distortion vs. Frequency

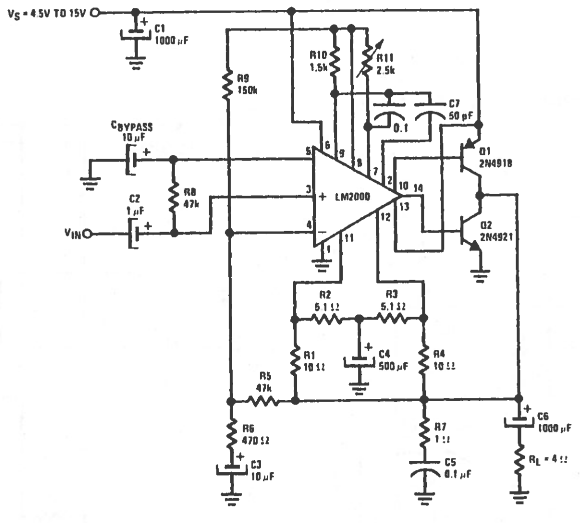


FIGURE 4.13.39 Complementary Output Amplifier

4.14 HEATSINKING

Insufficient heatsinking accounts for many phone calls made to complain about power ICs not meeting published specs. This problem may be avoided by proper application of the material presented in this section. Heatsinking is not difficult, although the first time through it may seem confusing.

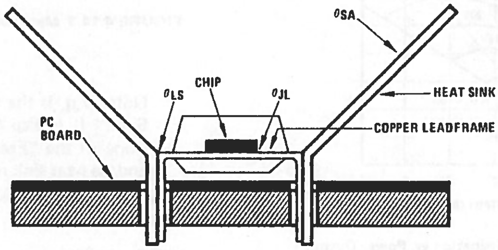
If testing a breadboarded power IC results in premature waveform clipping, or a "truncated shape," or a "melting down" of the positive peaks, the IC is probably in thermal shutdown and requires more heatsinking. The following information is provided to make proper heat sink selection easier and help take the "black magic" out of package power dissipation.

4.14.1 Heat Flow

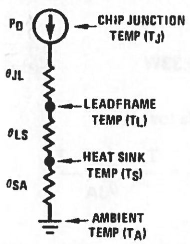
Heat can be transferred from the IC package by three methods, as described and characterized in Table 4.14.1.

TABLE 4.14.1 Methods of Heat Flow

METHOD	DESCRIBING PARAMETERS
Conduction is the heat transfer method most effective in moving heat from junction to case and case to heat sink.	Thermal resistance θ_{JL} and θ_{LS}. Cross section, length and temperature difference across the conducting medium.
Convection is the effective method of heat transfer from case to ambient and heat sink to ambient.	Thermal resistance θ_{SA} and θ_{LA}. Surface condition, type of convecting fluid, velocity and character of the fluid flow (e.g., turbulent or laminar), and temperature difference between surface and fluid.
Radiation is important in transferring heat from cooling fins.	Surface emissivity and area. Temperature difference between radiating and adjacent objects or space. See Table 4.14.2 for values of emissivity.



(a) Mechanical Diagram



(b) Electrical Equivalent

Symbols and Definitions

θ = Thermal Resistance (°C/W)
θ_{JL} = Junction to Leadframe
θ_{LS} = Leadframe to Heat Sink
θ_{SA} = Heat Sink to Ambient
θ_{JS} = Junction to Heat Sink = $\theta_{JL} + \theta_{LS}$
θ_{JA} = Junction to Ambient = $\theta_{JL} + \theta_{LS} + \theta_{SA}$
T_J = Junction Temperature (maximum) (°C)
T_A = Ambient Temperature
P_D = Power Dissipated (W)

(c) Symbols and Definitions

FIGURE 4.14.1 Heat Flow Model

4.14.2 Thermal Resistance

Thermal resistance is nothing more than a useful figure-of-merit for heat transfer. It is simply temperature drop divided by power dissipated, under steady state conditions. The units are usually °C/W and the symbol most used is θ_{AB}. (Subscripts denote heat flowing from A to B.)

The thermal resistance between two points of a conductive system is expressed as:

$$\theta_{12} = \frac{T_1 - T_2}{P_D} \text{ °C/W} \tag{4.14.1}$$

4.14.3 Modeling Heat Flow

An analogy may be made between thermal characteristics and electrical characteristics which makes modeling straightforward:

T – temperature differential is analogous to V (voltage)
θ – thermal resistance is analogous to R (resistance)
P – power dissipated is analogous to I (current)

Observe that just as R = V/I, so is its analog θ = T/P. The model follows from this analog.

A simplified heat transfer circuit for a power IC and heat sink system is shown in Figure 4.14.1. The circuit is valid only if the system is in thermal equilibrium (constant heat flow) and there are, indeed, single specific temperatures T_J, T_L, and T_S (no temperature distribution in junction, case, or heat sink). Nevertheless, this is a reasonable approximation of actual performance.

4.14.4 Where to Find Parameters

P_D

Package dissipation is read directly from the "Power Dissipation vs. Power Output" curves that are found on all of the audio amp data sheets. Most data sheets provide separate curves for either 4, 8 or 16Ω loads. Figure 4.14.2 shows the 8Ω characteristics of the LM378.

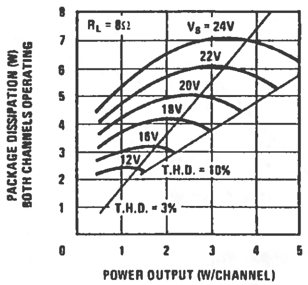


FIGURE 4.14.2 Power Dissipation vs. Power Output

> Note: For P_o = 2W and V_s = 18V, $P_{D(max)}$ = 4.1W, while the same P_o with V_s = 24V gives $P_{D(max)}$ = 6.5W — 50% greater! This point cannot be stressed too strongly: *For minimum P_D, V_s must be selected for the minimum value necessary to give the required power out.*

For loads other than those covered by the data sheet curves, max power dissipation may be calculated from Equation (4.14.2). (See Section 4.12.)

$$P_{D(max)} = \frac{V_s{}^2}{20\,R_L} \qquad (4.14.2)$$

Equation (4.14.2) is for each channel when applied to duals.

When used for bridge configurations, package dissipation will be twice that found from Figure 4.14.2

θ_{LS}

The thermal resistance between lead frame and heat sink is a function of how close the bond can be made. For the D.I.P., soldering to the ground pins with 60/40 solder is recommended. When soldered, θ_{LS} may be neglected or a value of θ_{LS} = 0.25°C/W may be used. Where the package style permits bolting to the heat sink, θ_{LS} will depend on whether a heat sink compound and/or an insulating washer is used. For a TO-3 case style 0.1°C/W is obtained with compound, increasing to 0.4°C/W with a 3 mil mica washer. The TO-220 case style used by the LM383 has corresponding values for θ_{LS} between 1.6°C/W and 2.6°C/W.

$T_{J(max)}$

Maximum junction temperature for each device is 150°C.

θ_{JL}

Thermal resistance between junction to lead frame (or junction to heat sink if θ_{LS} is ignored) is read, directly from the "Maximum Dissipation vs. Ambient Temperature" curve found on the data sheet. Figure 4.14.3 shows a typical curve for the LM378.

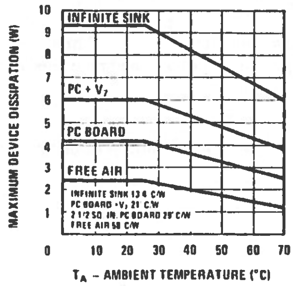


FIGURE 4.14.3 Maximum Dissipation vs. Ambient Temperature

> Note: θ_{JL} is the slope of the curve labeled "Infinite Sink." It is also $\theta_{JA(best)}$, while $\theta_{JA(worst)}$ is the slope of the "Free Air" curve, i.e., infinite heat sink and no heat sink respectively.

So, what does it mean? Simply that with no heat sink you can only dissipate

$$\frac{150°C - 25°C}{58°C/W} = 2.16W.$$

And with the best heat sink possible, the maximum dissipation is

$$\frac{150°C - 25°C}{13.4°C/W} = 9.33W$$

Or, for you formula lovers:

$$\text{Max Allowable } P_D = \frac{T_{J(max)} - T_A}{\theta_{JA}} \qquad (4.14.3)$$

4.14.5 Procedure for Selecting Heat Sink

1. Determine $P_{D(max)}$ from curve or Equation (4.14.2).
2. Neglect θ_{LS} if soldering; if not, θ_{LS} must be considered.
3. Determine θ_{JL} from curve.
4. Calculate θ_{JA} from Equation (4.14.3)
5. Calculate θ_{SA} for necessary heat sink by subtracting (2) and (3) from (4) above, i.e., $\theta_{SA} = \theta_{JA} - \theta_{JL} - \theta_{LS}$ (4.14.4)

For example, calculate heat sink required for an LM378 used with V_s=24V, R_L=8Ω, P_o=4W/channel and T_A=25°C:

1. From Figure 4.14.2, P_D=7W.
2. Heat sink will be soldered, so θ_{LS} is neglected.
3. From Figure 4.14.3, θ_{JL}=13.4°C/W.
4. From Equation (4.14.3),

$$\theta_{JA} = \frac{150°C - 25°C}{7W} = 17.9°C/W.$$

5. From Equation (4.14.4),

$$\theta_{SA} = 17.9°C/W - 13.4°C/W = 4.5°C/W.$$

Therefore, a heat sink with a thermal resistance of 4.5°C/W is required. Examination of Figure 4.14.3 shows this to be substantial heatsinking, requiring forethought as to board space, sink cost, etc.

Results modeled:

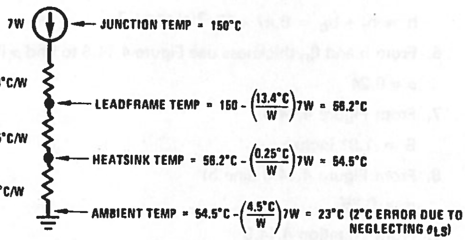


FIGURE 4.14.4 Heat Flow Model for LM378 Example

4.14.6 Custom Heat Sink Design

The required θ_{SA} was determined in Section 4.14.5. Even though many heat sinks are commercially available, it is sometimes more practical, more convenient, or more economical to mount the device to chassis, to an aluminum extrusion, or to a custom heat sink. In such cases, design a simple heat sink.

Simple Rules

1. Mount cooling fin vertically where practical for best conductive heat flow.
2. Anodize, oxidize, or paint the fin surface for better radiation heat flow; see Table 4.14.2 for emissivity data. However, note that although paint increases the emissivity of a surface, the paint itself has a high thermal resistance and should be removed where the semiconductor device is attached to the heat sink. (This will also apply to anodized and oxidized surfaces.)
3. Use 1/16" or thicker fins to provide low thermal resistance at the IC mounting where total fin cross-section is least.

Fin Thermal Resistance

The heat sink-to-ambient thermal resistance of a vertically mounted symmetrical square or round fin (see Figure 4.4.5) in still air is:

$$\theta_{SA} = \frac{1}{2H^2\eta(h_C + h_r)} \ °C/W \qquad (4.14.5)$$

where: H = height of vertical plate in inches

η = fin effectiveness factor

h_C = convection heat transfer coefficient

h_r = radiation heat transfer coefficient

$$h_C = 2.21 \times 10^{-3}\left(\frac{T_S - T_A}{H}\right)^{1/4} \ W/in^2°C \qquad (4.14.6)$$

$$h_r = 1.47 \times 10^{-10} E\left(\frac{T_S + T_A}{2} + 273\right)^3 \ W/in^2°C \qquad (4.14.7)$$

where: T_S = temperature of heat sink at IC mounting, in °C

T_A = ambient temperature in °C

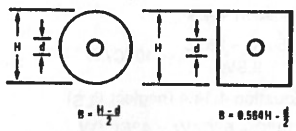
E = surface emissivity (see Table 4.14.2)

Fin effectiveness factor η includes the effects of fin thickness, shape, thermal conduction, etc. It may be determined from the nomogram of Figure 4.14.6.

TABLE 4.14.2 Emissivity Values for Various Surface Treatments

SURFACE	EMISSIVITY, E
Polished Aluminum	0.05
Polished Copper	0.07
Rolled Sheet Steel	0.66
Oxidized Copper	0.70
Black Anodized Aluminum	0.7 - 0.9
Black Air Drying Enamel	0.85 - 0.91
Dark Varnish	0.89 - 0.93
Black Oil Paint	0.92 - 0.96

For untreated copper and aluminum surfaces, E can be approximated to about 0.2.



Note: For H >> d, using B = H/2 is a satisfactory approximation for either square or round fins.

FIGURE 4.14.5 Symmetrical Fin Shapes

The procedure for use of the nomogram of Figure 4.14.6 is as follows:

1. Specify fin height H as first approximation.
2. Calculate $h = h_r + h_C$ from Equations (4.14.6) and (4.14.7).
3. Determine α from values of h and fin thickness x (line a).
4. Determine η from values of B (from Figure 4.14.5) and α (line b).

The value of η thus determined is valid for vertically mounted symmetrical square or round fins (with $H \gg d$) in still air. For other conditions, η must be modified as follows:

Horizontal mounting – multiply h_C by 0.7.

Horizontal mounting where only one side is effective – multiply η by 0.5 and h_C by 0.94.

For 2:1 rectangular fins – multiply h by 0.8.

For non-symmetrical fins where the IC is mounted at the bottom of a vertical fin – multiply η by 0.7.

Fin Design

1. Establish initial conditions, T_A and desired θ_{SA} as determined in Section 4.14.5.
2. Determine T_S at contact point with the IC by rewriting Equation (4.14.1):

$$\theta_{JL} + \theta_{LS} = \frac{T_J - T_S}{P_D} \qquad (4.14.8)$$

$$T_S = T_J - (\theta_{JL} + \theta_{LS})(P_D) \qquad (4.14.9)$$

$$\approx T_J - \theta_{JL} P_D$$

3. Select fin thickness, $x > 0.0625''$ and fin height, H.
4. Determine h_c and h_r from Equations (4.14.6) and (4.14.7).
5. Find fin effectiveness factor η from Figure 4.14.6.
6. Calculate θ_{SA} from Equation (4.14.5).
7. If θ_{SA} is too large or unnecessarily small, choose a different height and repeat steps (3) through (6).

Design Example

Design a symmetrical square vertical fin of 1/16" thick black anodized aluminum to be bolted onto an LM379 delivering a maximum power of 4W/Ch into 8Ω from a 28V supply.

1. LM379 operating conditions are:

   $T_J = 150°C_{(MAX)}$, $T_A = 55°C_{(MAX)}$

   From Figure 4.4.9, $\theta_{JL} = 6°C/W$

   From Figure 4.4.8, $P_{D(MAX)} = 9.5W$

2. From Equation 4.4.3

$$\theta_{JA} = \frac{150°C - 55°C}{9.5W} = 10°C/W$$

   From Equation 4.14.4 (neglect θ_{LS})

   $\theta_{SA} = 10°C - 6°C/W = 4°CC/W$

3. $T_S = 150°C - 6°C/W(9.5W) = 93°C$

4. $X = 0.0625''$ from initial conditions

   $E = 0.9$ from Table 4.14.2.

   Select H = 3.5" for first trial (experience will simplify this step).

5. From Equation 4.14.6

$$h_c = 2.21 \times 10^{-3} \frac{93 - 60}{3.5}^{1/4}$$

$$= 3.87 \times 10^{-3}\ W/°C\ in^2$$

   From Equation 4.14.7

$$h_r = 1.47 \times 10^{-10} \times 0.9 \frac{93 + 60 + 273}{2}^{3}$$

$$= 5.6 \times 10^{-3} W/°C\ in^2$$

$$h = h_r + h_c = 9.47 \times 10^{-3} W/°C\ in^2$$

6. From h and f_{in} thickness use Figure 4.14.6 to find α (line a)

   $\alpha = 0.24$

7. From Figure 4.14.5

   B = 1.91 inches

8. From Figure 4.14.6 (line b)

   $\eta = 0.85$

9. From Equation 4.14.5

$$\theta_{SA} = \frac{10^3}{2 \times 12.25 \times 0.85 \times 9.46} = 5.1°C/W$$

   Since the required heatsink thermal resistance is 4°C/W a larger f_{in} will be needed. A 4.25" square will increase the area by about 40% and a new calculation is made.

5'. $h_c = 3.7 \times 10^{-3} W/°C\ in^2$

   $h_r = 5.6 \times 10^{-3} W/°C\ in^2$

   $h = 9.3 \times 10^{-3} W/°C\ in^2$

6'. $\alpha = 0.24$

7'. B = 2.4

8'. $\eta = 0.73$

9'. $\theta_{SA} = 4.08°C/W$ which is satisfactory.

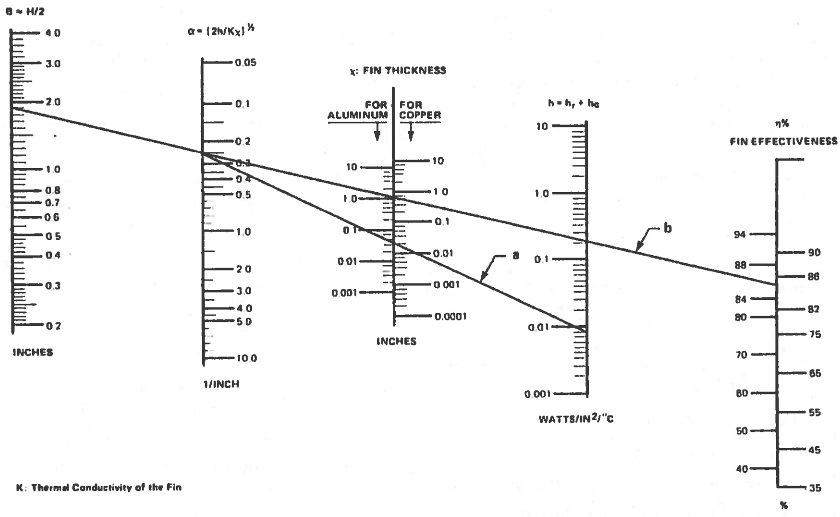


FIGURE 4.14.6 Fin Effectiveness Nomogram for Symmetrical, Flat, Uniformly-Thick, Vertically Mounted Fins

Although the above design procedure will specify the dimensions of the required heatsink, any design should be thoroughly tested under actual operating conditions to ensure that the maximum device case temperature does not exceed the rating for worst case thermal and load conditions.

4.14.7 Heatsinking with PC Board Foil

National Semiconductor's use of copper leadframes in packaging power ICs, where the center three pins on either side of the device are used for heatsinking, allows for economical heat sinks via the copper foil that exists on the printed circuit board. Adequate heatsinking may be obtained for many designs from single-sided boards constructed with 2 oz. copper. Other, more stringent, designs may require two-sided boards, where the top side is used entirely for heatsinking. Figure 4.14.7 allows easy design of PC board heat sinks once the desired thermal resistance has been calculated from Section 4.14.5.

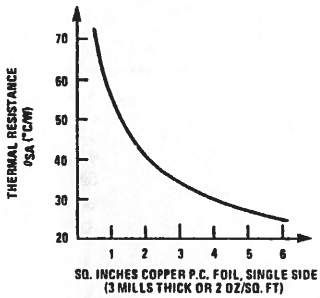


FIGURE 4.14.7 Thermal Resistance vs. Square Inches of Copper Foil

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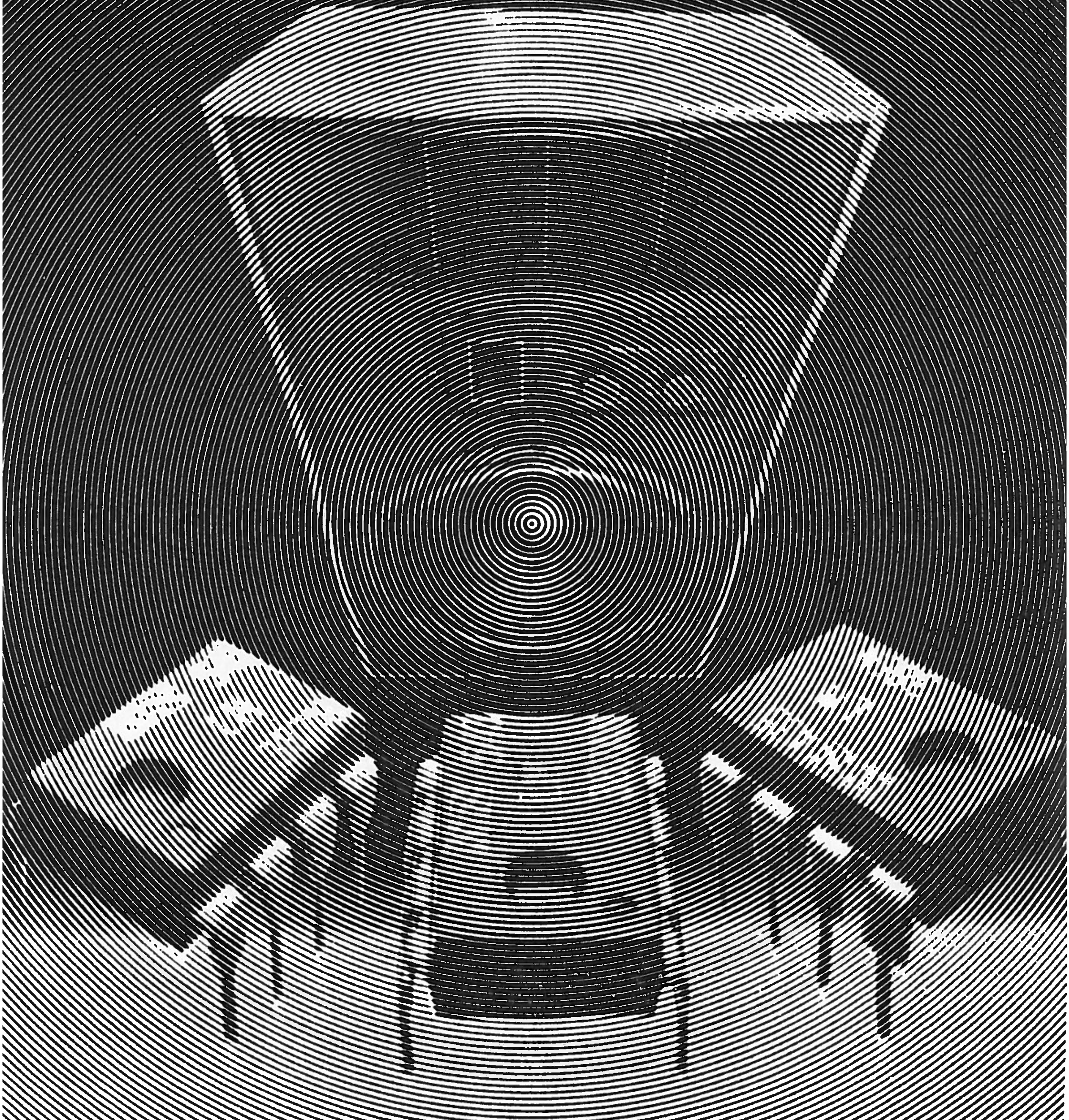
Section 5.0
Floobydust



5.0 Floobydust

5.1 BIAMPLIFICATION

The most common method of amplifying the output of a preamplifier into the large signal required to drive a speaker system is with one large wideband amplifier having a flat frequency response over the entire audio band. An alternate method is to employ two amplifiers, or biamplification, where each amplifier is committed to amplifying only one part of the frequency spectrum. Biamping requires splitting up the audio band into two sections and routing these signals to each amplifier. This process is accomplished by using an active crossover network as discussed in the next section.

The most common application of biamping is found in conjunction with speaker systems. Due to the difficulty of manufacturing a single speaker capable of reproducing the entire audio band, multiple speakers are used, where each speaker is designed only to reproduce one section of frequencies. In conventional systems using one power amplifier the separation of the audio signal is done by passive high and low pass filters located within the speaker enclosure as diagrammed in Figure 5.1.1. These filters must be capable of processing high power signals and are therefore troublesome to design, requiring large inductors and capacitors.

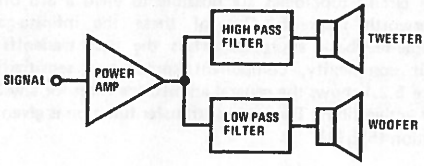


FIGURE 5.1.1 Passive Crossover, Single Amp System

Biamping with active crossover networks (Figure 5.1.2) allows a more flexible and easier design. It also sounds better. Listening tests demonstrate that biamped systems have audibly lower distortion.[4] This is due chiefly to two effects. The first results from the consequence of bass transient clipping. Low frequency signals tend to have much higher transient amplitudes than do high frequencies, so amplifier overloading normally occurs for bass signals. By separating the spectrum one immediately cleans up half of it and greatly improves the other half, in that the low frequency speaker will not allow high frequency components generated by transient clipping of the bass amplifier to pass, resulting in cleaner sound. Second is a high frequency masking effect, where the low level high frequency distortion components of a clipped low frequency signal are "covered up" (i.e., masked) by high level undistorted high frequencies. The final advantage of biamping is allowing the use of smaller power amplifiers to achieve the same sound pressure levels.

5.2 ACTIVE CROSSOVER NETWORKS

An active crossover network is a system of active filters (usually two) used to divide the audio frequency band into separate sections for individual signal processing by biamped systems. Active crossovers are audibly desirable because they give better speaker damping and improved transient response, and minimize midrange modulation distortion.

5.2.1 Filter Choice

The choice of filter type is based upon the need for good transient and frequency response. Bessel filters offer excellent phase and transient response but suffer from frequency response change in the crossover region, being too slow for easy speaker reproduction. Chebyshev filters have excellent frequency division but possess unacceptable instabilities in their transient response. Butterworth characteristics fall between Bessel and Chebyshev and offer the best compromise for active crossover design.

5.2.2 Number of Poles (Filter Order)

Intuitively it is reasonable that if the audio spectrum is split into two sections, their sum should exactly equal the original signal, i.e., without change in phase or magnitude (vector sum must equal unity). This is known as a constant voltage design. Also it is reasonable to want the same power delivered to each of the drivers (speakers). This is known as constant power design. What is required, therefore, is a filter that exhibits constant voltage and constant power. Having decided upon a Butterworth filter, it remains to

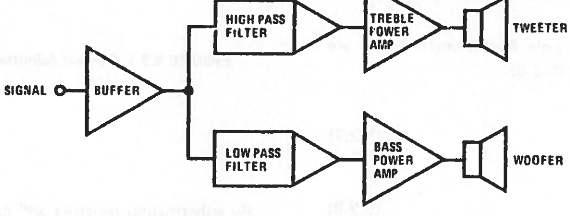


FIGURE 5.1.2 Active Crossover, Biamp System

determine an optimum order of the filter (the number of poles found in its transfer function) satisfying constant voltage and constant power.

Both active and passive realizations of a Butterworth filter have identical transfer functions, so a good place to start is with conventional passive crossover networks. Passive crossovers exhibit a single pole (1st order) response and have a transfer function given by Equations (5.2.1) and (5.2.2) (normalized to $\omega_o = 1$).

$$T_L(S) = \frac{1}{S+1} \tag{5.2.1}$$

$$T_H(S) = \frac{S}{S+1} \tag{5.2.2}$$

where $T_L(S)$ equals low pass transfer function and $T_H(S)$ equals high pass transfer function. This filter exhibits constant voltage (hence, constant power) as follows:

$$\text{require } T_L(S) + T_H(S) = 1 \tag{5.2.3}$$

Inspection of Equations (5.2.1) and (5.2.2) shows this to be true.

The problem with a single pole system is that the rolloff beyond the crossover point is only −6dB/octave and requires the speakers to operate linearly for two additional octaves if distortion is to be avoided.[6]

The 2nd order system exhibits transfer functions:

$$T_L(S) = \frac{1}{S^2 + \sqrt{2}S + 1} \tag{5.2.4}$$

$$T_H(S) = \frac{S^2}{S^2 + \sqrt{2}S + 1} \tag{5.2.5}$$

These transfer functions exhibit constant power but not constant voltage. This is demonstrated by applying Equation (5.2.3), yielding:

$$T_L(S) + T_H(S) = \frac{S^2 + 1}{S^2 + \sqrt{2}S + 1} \tag{5.2.6}$$

At crossover, $S = -j\omega_o = -j$ (since $\omega_o = 1$); substitution into Equation (5.2.6) equals zero. This means that at the crossover frequency there exists a "hole," or a frequency that is not reproduced by either speaker. Ashley[1] demonstrated that this hole is audible. A commonly seen solution to this problem is to invert the polarity of one speaker in the system. Mathematically this changes the sign of the transfer function and effectively subtracts the two terms rather than adds them. This does eliminate the hole, but it creates a new problem of severe phase shifting at the crossover point which Ashley also demonstrated to be audible, making consideration of 3rd order Butterworth filters necessary.

The transfer functions for 3 pole Butterworth filters are given as Equations (5.2.7) and (5.2.8).

$$T_L(S) = \frac{1}{S^3 + 2S^2 + 2S + 1} \tag{5.2.7}$$

$$T_H(S) = \frac{S^3}{S^3 + 2S^2 + 2S + 1} \tag{5.2.8}$$

Applying Equation (5.2.3) yields:

$$T_L(S) + T_H(S) = \frac{S^3 + 1}{S^3 + 2S^2 + 2S + 1} \tag{5.2.9}$$

which at $S = -j\omega_o$ gives

$$T_L(-j\omega_o) + T_H(-j\omega_o) = -1 \tag{5.2.10}$$

Equation (5.2.10) satisfies constant voltage and constant power with one nagging annoyance — the phase has been inverted. Examination of the phase characteristics of Equation (5.2.9) shows that there is a gradual phase shift from 0° to −360° as the frequency is swept through the filter sections, being −180° at ω_o. Is it audible? Ashley[2] demonstrated that the ear cannot detect this gradual phase shift when it is not accompanied by ripple in the magnitude characteristic. (It turns out that all odd ordered Butterworth filters exhibit this effect with increasing amounts of phase shift, e.g., 5th order gives 0 to −720°, etc.)

The conclusion is that the best compromise is to use a 3rd order Butterworth filter. It will exhibit maximally flat magnitude response, i.e., no peaking (which minimizes the work required by the speakers); it has sharp cutoff characteristics of −18dB/octave (which minimizes speakers being required to reproduce beyond the crossover point); and it has flat voltage and power frequency response with a gradual change in phase across the band.

5.2.3 Design Procedure for 3rd Order Butterworth Active Crossovers

Many circuit topologies are possible to yield a 3rd order Butterworth response. Out of these the infinite-gain, multiple-feedback approach offers the best tradeoffs in circuit complexity, component spread and sensitivities. Figure 5.2.1 shows the general admittance form for any 3rd order active filter. The general transfer function is given by Equation (5.2.11).

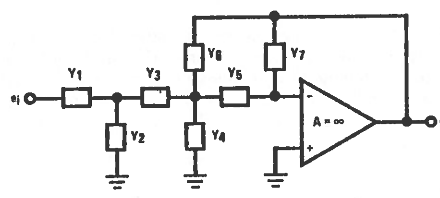


FIGURE 5.2.1 General Admittance Form for 3rd Order Filter

By substituting resistors and capacitors for the admittances per Figures 5.2.2 and 5.2.3, low and high pass active filters are created.

$$\frac{e_o}{e_i} = -\frac{Y_1Y_3Y_5}{(Y_5Y_6 + Y_3Y_7 + Y_4Y_7 + Y_5Y_7 + Y_6Y_7)(Y_1 + Y_2 + Y_3) - Y_7Y_3^2} \tag{5.2.11}$$

Low Pass:

$$\frac{e_{oL}}{e_i} = -\frac{\dfrac{1}{R_1R_3R_5C_2C_4C_7}}{S^3 + \left(\dfrac{R_5R_6 + R_3R_6 + R_3R_5}{R_3R_5R_6C_4} + \dfrac{R_1 + R_3}{R_1R_3C_2}\right)S^2 + \left(\dfrac{1}{R_5R_6C_4C_7} + \dfrac{R_5R_6 + R_3R_6 + R_3R_5 + R_1R_5 + R_1R_6}{R_1R_3R_5R_6C_2C_4}\right)S + \dfrac{R_1 + R_3}{R_1R_3R_5R_6C_2C_4C_7}} \tag{5.2.12}$$

High Pass:

$$\frac{e_{oH}}{e_i} = -\frac{\dfrac{C_1C_3}{C_6(C_1 + C_3)}S^3}{S^3 + \left(\dfrac{C_1(C_3 + C_5 + C_6) + C_3(C_5 + C_6)}{R_7C_5C_6(C_1 + C_3)} + \dfrac{1}{(C_1 + C_3)R_2}\right)S^2 + \left(\dfrac{1}{C_5C_6R_4R_7} + \dfrac{C_3 + C_5 + C_6}{C_5C_6(C_1 + C_3)R_2R_7}\right)S + \dfrac{1}{C_5C_6(C_1 + C_3)R_2R_4R_7}} \tag{5.2.13}$$

Substitution of the appropriate admittances shown in Figures 5.2.2 and 5.2.3 into Equation 5.2.11 gives the general equation for a 3rd order low pass (Equation (5.2.12)) and for a 3rd order high pass (Equation (5.2.13)):

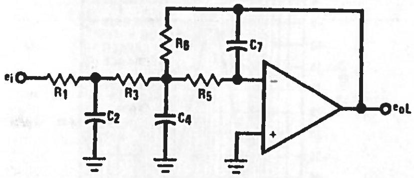


FIGURE 5.2.2 General 3rd Order Low Pass Active Filter

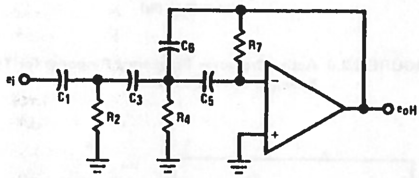


FIGURE 5.2.3 General 3rd Order High Pass Active Filter

Equation (5.2.12) is of form

$$\frac{K\omega_o^3}{S^3 + aS^2 + bS + \omega_o^3}$$

where: K = passband gain = 1

Letting $a = b = 2$ and normalizing $\omega_o^3 = 1$ gives the 3rd order Butterworth response of Equation (5.2.7).

Similarly, Equation (5.2.13) is of form

$$\frac{KS^3}{S^3 + aS^2 + bS + \omega_o^3}$$

and corresponds to Equation (5.2.8).

By letting $R_1 = R_3 = R_5 = R$ and $R_6 = 2R$ and equating coefficients between Equations (5.2.12) and (5.2.7), it is possible to solve for the capacitor values in terms of R. Doing so yields the relationships shown in Figure 5.2.4. For the high pass section, let $C_1 = C_3 = C_5 = C$ and $C_6 = C/2$ and equate coefficients to get the resistor values in terms of C. The high pass results also appear in Figure 5.2.4, which shows the complete 3rd order Butterworth crossover network.

Example 5.2.1

Design an active crossover network with −18dB/octave rolloff (3rd order), maximally flat (Butterworth) characteristics having a crossover frequency equal to 500Hz.

1. Let: R = 10k (1%)

2. Calculate C_2, C_4 and C_7 from Figure 5.2.4:

$$C_2 = \frac{2.4553}{(2\pi)(500)(10K)} = 7.82 \times 10^{-8}$$

Use $C_2 = 0.082\mu F$, 2%.

$$C_4 = \frac{2.1089}{(2\pi)(500)(10K)} = 6.71 \times 10^{-8}$$

Use $C_4 = 0.068\mu F$, 2%.

$$C_7 = \frac{0.1931}{(2\pi)(500)(10K)} = 6.51 \times 10^{-9}$$

Use $C_7 = 0.0056\mu F$, 2%.

3. Select C for high pass section to have same impedance level as R_{IN} for low pass, i.e., 20kΩ:

Let C = 0.015μF, 2%

C/2 = 0.0082μF, 2%.

4. Calculate R_2, R_4 and R_7 from Figure 5.2.4:

$$R_2 = \frac{0.4074}{(2\pi)(500)(1.592 \times 10^{-8}} = 8148$$

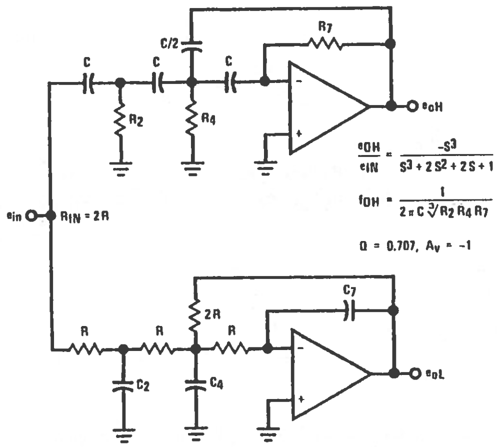
Use $R_2 = 8.06K$, 1%.

$$R_4 = \frac{0.4742}{(2\pi)(500)(1.592 \times 10^{-8})} = 9484$$

Use $R_4 = 9.53K$, 1%.

$$R_7 = \frac{5.1766}{(2\pi)(500)(1.592 \times 10^{-8})} = 103532$$

Use $R_7 = 102K$, 1%.



$$C_2 = \frac{2.4553}{2\pi f_{OL} R} \qquad R_2 = \frac{0.4074}{2\pi f_{OH} C} \qquad \frac{e_{OL}}{e_{IN}} = \frac{-1}{S^3 + 2S^2 + 2S + 1}$$

$$C_4 = \frac{2.1089}{2\pi f_{OL} R} \qquad R_4 = \frac{0.4742}{2\pi f_{OH} C} \qquad f_{OL} = \frac{1}{2\pi R \sqrt[3]{C_2 C_4 C_7}}$$

$$C_7 = \frac{0.1931}{2\pi f_{OL} R} \qquad R_7 = \frac{5.1766}{2\pi f_{OH} C} \qquad Q = 0.707,\ A_V = -1$$

FIGURE 5.2.4 Complete 3rd Order Butterworth Crossover Network

The completed design is shown in Figure 5.2.5 using LF356 op amps for the active devices. LF356 devices were chosen for their very high input impedances, fast slew and extremely stable operation into capacitive loads. A buffer is used to drive the crossover network for two reasons: it guarantees low driving impedance which active filters require, and it gives another phase inversion so that the outputs are in phase with the inputs. Power supplies are ±15V, decoupled with 0.1 ceramic capacitors located close to the integrated circuits (not shown). Figure 5.2.6 gives the frequency response of Figure 5.2.5.

Figure 5.2.7 can be used to "look up" values for standard crossover frequencies of 100Hz to 5kHz.

5.2.4 Alternate Design for Active Crossovers

The example of Figure 5.2.5 is known as a symmetrical filter since both high and low pass sections are symmetrical about the crossover point (see Figure 5.2.6). An interesting alternate design is known as the asymmetrical filter (since the high and low pass sections are asymmetrical about the crossover point). This design is based upon the simple realization that if the output of a high pass filter is subtracted from the original signal then the result is a low pass.[3] Constant voltage is assured since the sum of low and high pass are always equal to unity (with no phase funnies). But, as always, there are tradeoffs and this time they are not obvious.

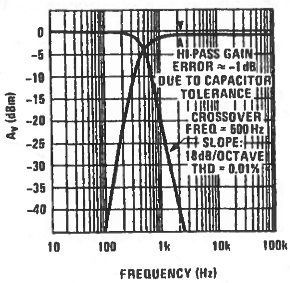


FIGURE 5.2.6 Active Crossover Frequency Response for Typical Example of Figure 5.2.5

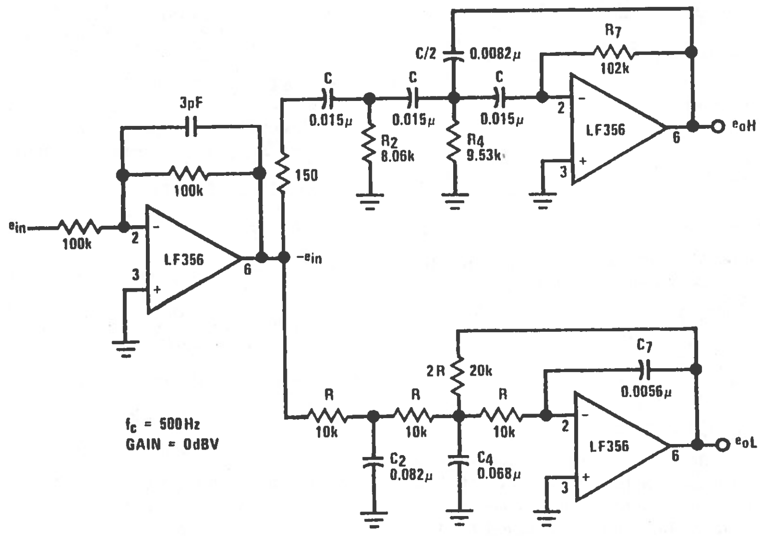


FIGURE 5.2.5 Typical Active Crossover Network Example

f_c Hz	C μF	R_2 Ω	R_4 Ω	R_7 Ω	C_2 μF	C_4 μF	C_7 μF
100	0.080	8148	9484	103532	0.391	0.336	0.0307
200	0.040				0.195	0.168	0.0154
300	0.027				0.130	0.112	0.0102
400	0.020				0.0977	0.0839	0.00768
500	0.016				0.0782	0.0671	0.00615
600	0.013				0.0651	0.0559	0.00512
700	0.011				0.0558	0.0479	0.00439
800	0.010				0.0488	0.0420	0.00384
900	0.0088				0.0434	0.0373	0.00341
1k	0.008				0.0391	0.0336	0.00307
2k	0.004				0.0195	0.0168	0.00154
3k	0.0027				0.0130	0.0112	0.00102
4k	0.002				0.00977	0.00839	768pF
5k	0.0016				0.00782	0.00671	615pF

* Assumes R = 10k

FIGURE 5.2.7 Precomputed Values for Active Crossover Circuit Shown in Figure 5.2.4 (Use nearest available value.)

Referring back to Equation (5.2.8) for the transfer function of a 3rd order high pass and subtracting it from the original signal yields the following:

$$T_L(S) = 1 - T_H(S) \qquad (5.2.14)$$

$$T_L(S) = 1 - \frac{S^3}{S^3 + 2S^2 + 2S + 1}$$

$$T_L(S) = \frac{2S^2 + 2S + 1}{S^3 + 2S^2 + 2S + 1} \qquad (5.2.15)$$

Analysis of Equation (5.2.15) shows it has two zeros and three poles. The two zeros are in close proximity to two of the poles and near cancellation occurs. The net result is a low pass filter that exhibits only −6dB rolloff and rather severe peaking (~ +4dB) at the crossover point. For low frequency drivers with extended frequency response, this is an attractive design offering lower parts count, easy adjustment, no crossover hole and without gradual phase shift.

Figure 5.2.8 shows the circuit design for an asymmetrical filter, and Figure 5.2.9 gives its frequency response.

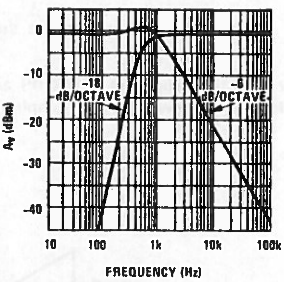


FIGURE 5.2.9 Frequency Response of Asymmetrical Filter Shown in Figure 5.2.8

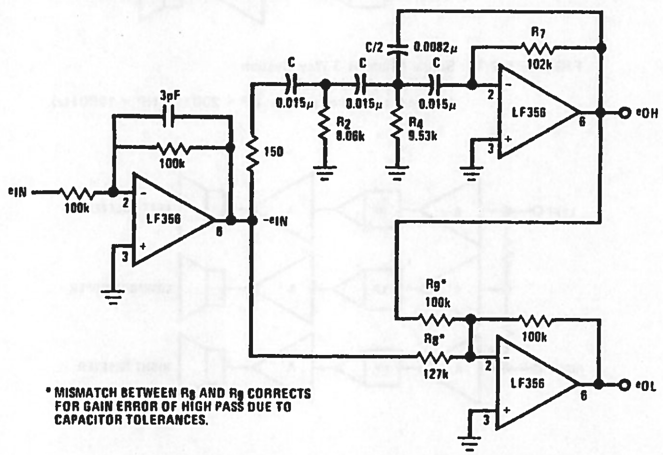


FIGURE 5.2.8 Asymmetrical 3rd Order Butterworth Active Crossover Network

5.2.5 Use of Crossover Networks and Biamping

Symbolically, Figure 5.2.5 can be represented as shown in Figure 5.2.10:

Figures 5.2.11-5.2.14 use Figure 5.2.10 to show several speaker systems employing active crossover networks and biamping.

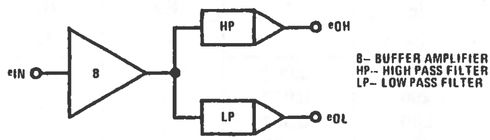


FIGURE 5.2.10 Symbolic Representation of Figure 5.2.5

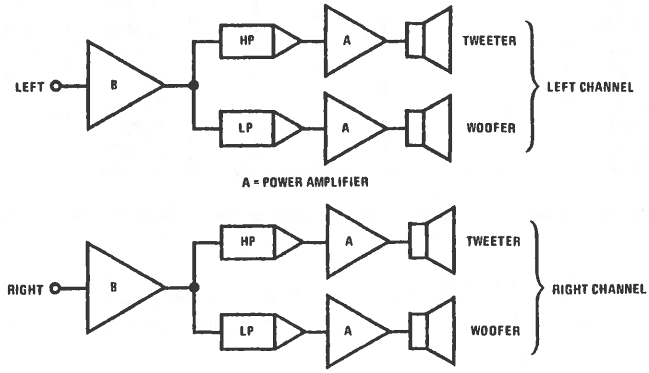


FIGURE 5.2.11 Stereo 2-Way System (Typical crossover points from 800 to 1600 Hz)

Cascading low pass (LP) and high pass (HP) active filters creates a bandpass and allows system triamping as follows:

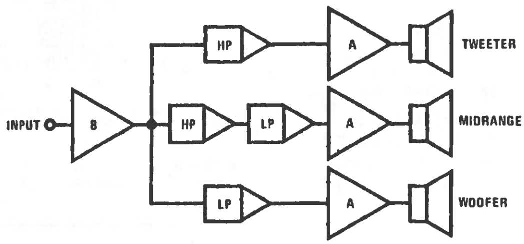


FIGURE 5.2.12 Single Channel 3-Way System (Duplicate for Stereo) (Typical crossover points: LP = 200 Hz, HP = 1200 Hz)

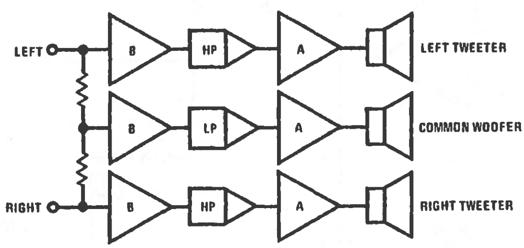


FIGURE 5.2.13 Common Woofer 2-Way Stereo System[5] (Stereo-to-mono crossover point typically 150 Hz)

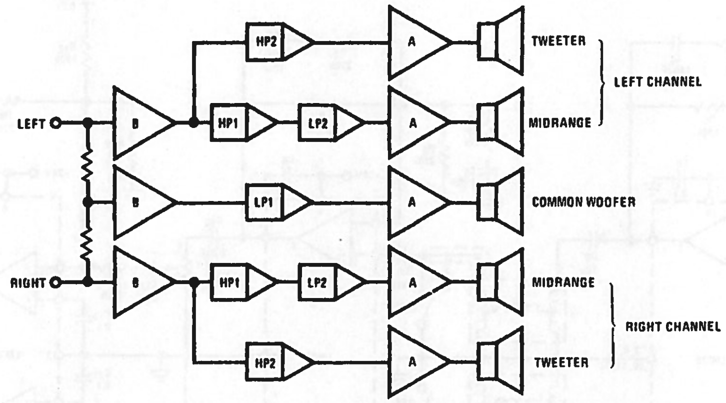


FIGURE 5.2.14 Common Woofer 3-Way Stereo System (Typically LP1 = HP1 = 150 Hz, LP2 = HP2 = 2500 Hz)

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5.3 REVERBERATION

Reverberation is the name applied to the echo effect associated with a sound after it has stopped being generated. It is due to the reflection and re-reflection of the sound off the walls, floor and ceiling of a listening environment and under certain conditions will act to enhance the sound. It is the main ingredient of concert hall ambient sound and accounts for the richness of "live" versus "canned" music. By using electro-mechanical devices, it is possible to add artificial reverberation to existing music systems and enhance their performance. The most common reverberation units use two precise springs that act as mechanical delay lines, each delaying the audio signal at slightly different rates. (Typical delay times are ~ 30 milliseconds for one spring and ~ 40 milliseconds for the other, with total decay times being around 2 seconds.) The electrical signal is applied to the input transducer where it is translated into a torsional force via two small cylindrical magnets attached to the springs. This "twisting" of one end of each spring slowly propogates along the length of the unit until it arrives at the other end, where similar magnets convert it back into an electrical signal. (Reflection also occurs, which creates the long decay time, relative to the delay time.)

5.3.1 Design Considerations for Driver and Recovery Amplifiers

Since the reverb driver is applying an electrical signal to a coil, its load is essentially inductive and as such has a rising impedance vs. frequency characteristic of +6dB/octave. Further, since the spring assembly operates best at a fixed value of ampere/turns (independent of frequency), it becomes desirable to drive the transducer with constant current. Constant current can be achieved in two ways: (1) by incorporating the input transducer as part of the negative feedback network, or (2) by creating a rising output voltage response as a function of frequency to follow the corresponding rise in output impedance. Method (1) precludes the use of grounded input transducers, which tend to be quieter and less susceptible to noise transients. (While grounded load, constant current sources exist, they require more parts to implement.) For this reason method (2) is preferred and will be used as a typical design example.

A high slew rate (~ 2V/μs) amplifier should be used since the rising amplitude characteristic necessitates full output swing at the maximum frequency of interest (typical spring assemblies have a frequency response of 100Hz-5kHz), thereby allowing enough headroom to prevent transient clipping. It is also advisable to roll the amplifier off at high frequencies as a further aid in headroom. "Booming" at low frequencies is controlled by rolling off low frequencies below 100Hz.

The requirements of the recovery amplifier are determined by the recovered signal. Typical voltage levels at the transducer output are in the range of 1-5mV, therefore requiring a low noise, high gain preamp. Hum and noise need to be minimized by using shielding cable, mounting the reverb assembly and preamp away from the power supply transformer, and using good single point ground techniques to avoid ground loops. Equalization is not necessary if a constant current drive amplifier is used since the output voltage is constant with frequency.

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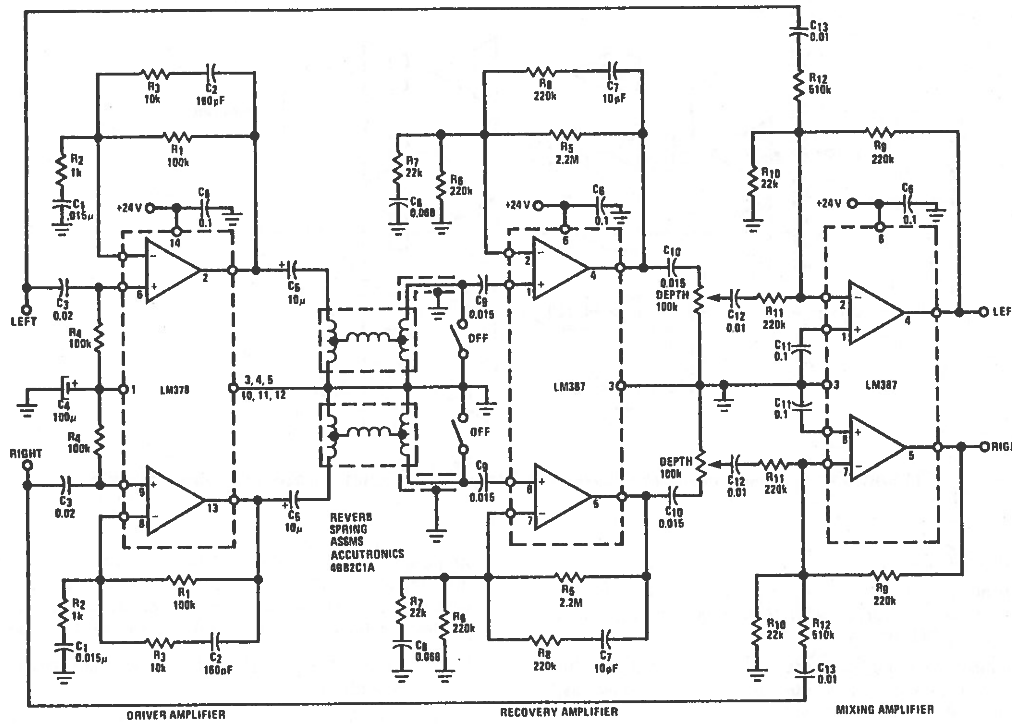

FIGURE 5.3.1 Stereo Reverb System

5.3.2 Stereo Reverb System

A complete stereo reverb system is shown in Figure 5.3.1, with its idealized "straightline" frequency response appearing as Figure 5.3.2.

The LM378 dual power amplifier is used as the spring driver because of its ability to deliver large currents into inductive loads. Some reverb assemblies have input transducer impedance as low as 8Ω and require drive currents of ~ 30mA. (There is a preference among certain users of reverbs to drive the inputs with as much as several hundred milliamps.) The recovery amplifier is easily done by using the LM387 low noise dual preamplifier which gives better than 75dB signal-to-noise performance at 1kHz (10mV recovered signal). Mixing of the delayed signal with the original is done with another LM387 used in an inverting summing configuration.

Figure 5.3.2 shows the desired frequency shaping for the driver and recovery amplifiers. The overall low frequency response is set by f_o and occurs when the reactance of the coupling capacitors equals the input impedance of the next stage. For example, the driver stage low frequency corner f_o is found from Equation (5.3.1).

$$f_o = \frac{1}{2\pi R_4 C_3} \approx 80\text{Hz (as shown)} \tag{5.3.1}$$

The +6dB/octave response is achieved by proper selection of R_1, R_2 and C_1 as follows:

$$f_1 = \frac{1}{2\pi (R_1 + R_2) C_1} \approx 100\text{Hz (as shown)} \tag{5.3.2}$$

$$f_2 = \frac{1}{2\pi R_2 C_1} \approx 10\text{kHz (as shown)} \tag{5.3.3}$$

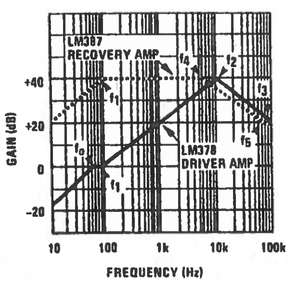

FIGURE 5.3.2 Straightline Frequency Response of Reverb Driver and Recovery Amplifiers

Ultimate gain is given by the ratio of R_2 and R_1:

$$A_O = 1 + \frac{R_2}{R_1} \text{ (gain beyond } f_2 \text{ corner)} \tag{5.3.4}$$

High frequency rolloff is accomplished with R_3 and C_2, beginning at f_2 and stopping at f_3.

$$f_2 = \frac{1}{2\pi R_1 C_2} \approx 10\,\text{kHz (as shown)} \tag{5.3.5}$$

$$f_3 = \frac{1}{2\pi R_3 C_2} \approx 100\,\text{kHz (as shown)} \tag{5.3.6}$$

Stopping high frequency rolloff at f_3 is necessary so the gain of the amplifier does not drop lower than 20dB, thereby preserving stability. (LM378 is not unity gain stable.) Resistors R_5 and R_6 are selected to bias the output of the LM387 at half-supply. (See Section 2.8.) Low frequency corner f_1 is fixed by R_7 and C_8:

$$f_1 = \frac{1}{2\pi R_7 C_8} \approx 100\,\text{Hz (as shown)} \tag{5.3.7}$$

High frequency rolloff is done similar to the LM377 by R_8 and C_7:

$$f_4 = \frac{1}{2\pi R_5 C_7} \approx 7\,\text{kHz (as shown)} \tag{5.3.8}$$

$$f_5 = \frac{1}{2\pi R_8 C_7} \approx 70\,\text{kHz (as shown)} \tag{5.3.9}$$

The same stability requirements hold for the LM387 as for the LM378.

Resistors R_9 and R_{10} are used to bias the LM387 summing amplifier. The output of the summer will be the scaled sum of the original signal and the delayed signal. Scaling factors are adjusted per Equation (5.3.10).

$$-V_{OUT} = \frac{R_9}{R_{12}} V_s + \frac{R_9}{R_{11}} V_D \tag{5.3.10}$$

where: V_s = original signal
V_D = delayed signal

As shown, the output is the sum of approximately one half of the original signal and all of the delayed signal.

5.3.3 Stereo Reverb Enhancement System

The system shown in Figure 5.3.3 can be used to synthesize a stereo effect from a monaural source such as AM radio or FM-mono broadcast, or it can be added to an existing stereo (or quad) system where it produces an exciting "opening up" spacial effect that is truly impressive.

The driver and recovery sections are as in Figure 5.3.1 with the exception that only one spring assembly is required. The second half of the LM387 recovery amplifier is used as an inverter and a new LM387 is added to mix both channels together. The outputs are inverted, scaled sums of the original and delayed signals such that the left output is composed of LEFT minus DELAY and the right output is composed of RIGHT plus DELAY.

When applied to mono source material, both inputs are tied together and the two outputs become INPUT minus DELAY and INPUT plus DELAY, respectively. If the outputs are to be used to drive speakers directly (as in an automotive application, or small home systems), then the LM387 may be replaced by one of the LM1877/378/379 dual 2W/4W/6W amplifier family wired as an inverting power summer per Figure 5.3.4.

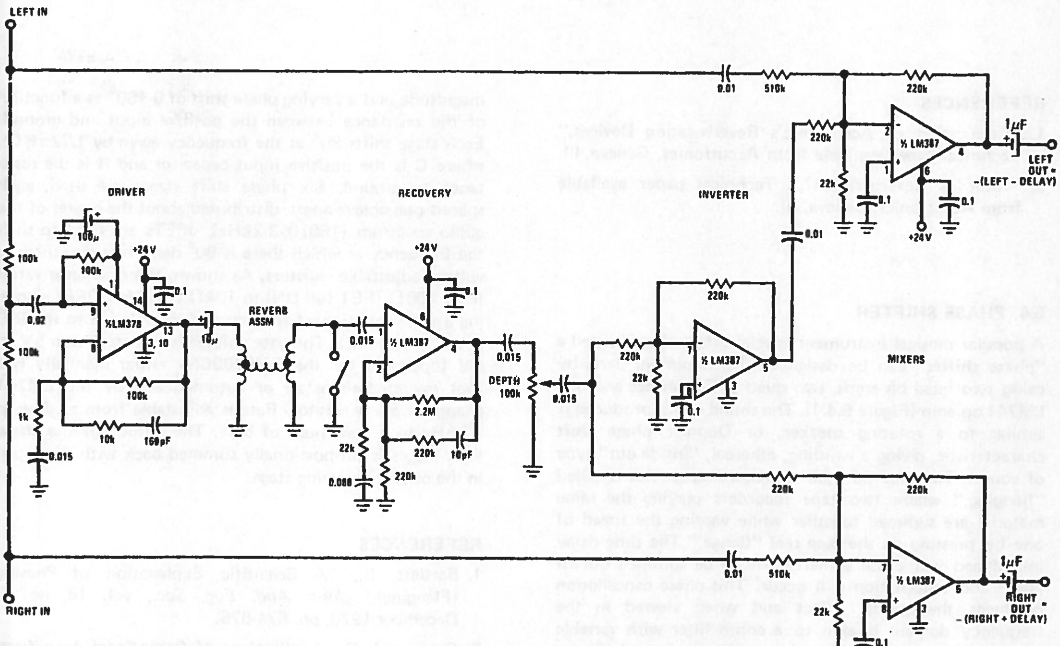

FIGURE 5.3.3 Stereo Reverb Enhancement System

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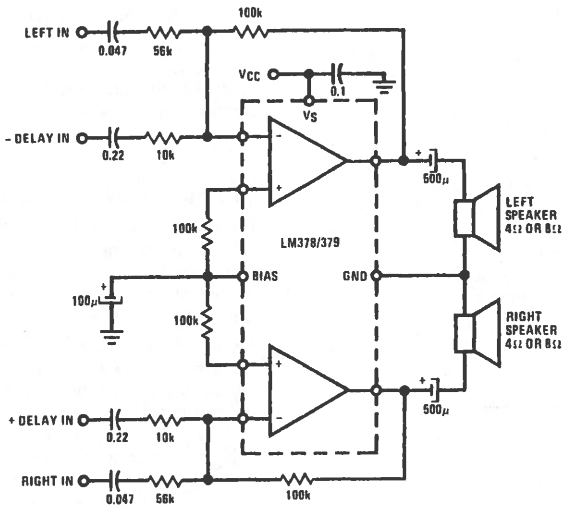

FIGURE 5.3.4 Alternate Output Stage for Driving Speakers Directly Using LM378/379 Family of Power Amplifiers

REFERENCES

1. "Application of Accutronic's Reverberation Devices," Technical paper available from Accutronics, Geneva, Ill.
2. "What Is Reverberation?," Technical paper available from Accutronics, Geneva, Ill.

5.4 PHASE SHIFTER

A popular musical instrument special effect circuit called a "phase shifter" can be designed with minimum parts by using two quad op amps, two quad JFET devices and one LM741 op amp (Figure 5.4.1). The sound effect produced is similar to a rotating speaker, or Doppler phase shift characteristic, giving a whirling, ethereal, "inside out" type of sound. The method used by recording studios is called "flanging," where two tape recorders playing the same material are summed together while varying the speed of one by pressing on the tape reel "flange." The time delay introduced will cause some signals to be summed out of phase and cancellation will occur. This phase cancellation produces the special effect and when viewed in the frequency domain is akin to a comb filter with variable rejection frequencies.[1] The phase shift stage used (Figure 5.4.1) is a standard configuration[2] displaying constant magnitude and a varying phase shift of 0-180° as a function of the resistance between the positive input and ground. Each stage shifts 90° at the frequency given by $1/(2\pi RC)$, where C is the positive input capacitor and R is the resistance to ground. Six phase shift stages are used, each spaced one octave apart, distributed about the center of the audio spectrum (160Hz-3.2kHz). JFETs are used to shift the frequency at which there is 90° delay by using them as voltage adjustable resistors. As shown, the resistance varies from 100Ω (FET full ON) to 10kΩ (FET full OFF), allowing a wide variation of frequency shift (relative to the 90° phase shift point). The gate voltage is adjusted from 5V to 8V (optimum for the AM9709CN), either manually (via foot operated rheostat) or automatically by the LM741 triangle wave generator. Rate is adjustable from as slow as 0.05Hz to a maximum of 5Hz. The output of the phase shift stages is proportionally summed back with the input in the output summing stage.

REFERENCES

1. Bartlett, B., "A Scientific Explanation of Phasing (Flanging)," *Jour. Aud. Eng. Soc.*, vol. 18, no. 6, December 1970, pp. 674-675.
2. Graeme, J. G., *Applications of Operational Amplifiers*, McGraw-Hill, New York, 1973, pp. 102-104.

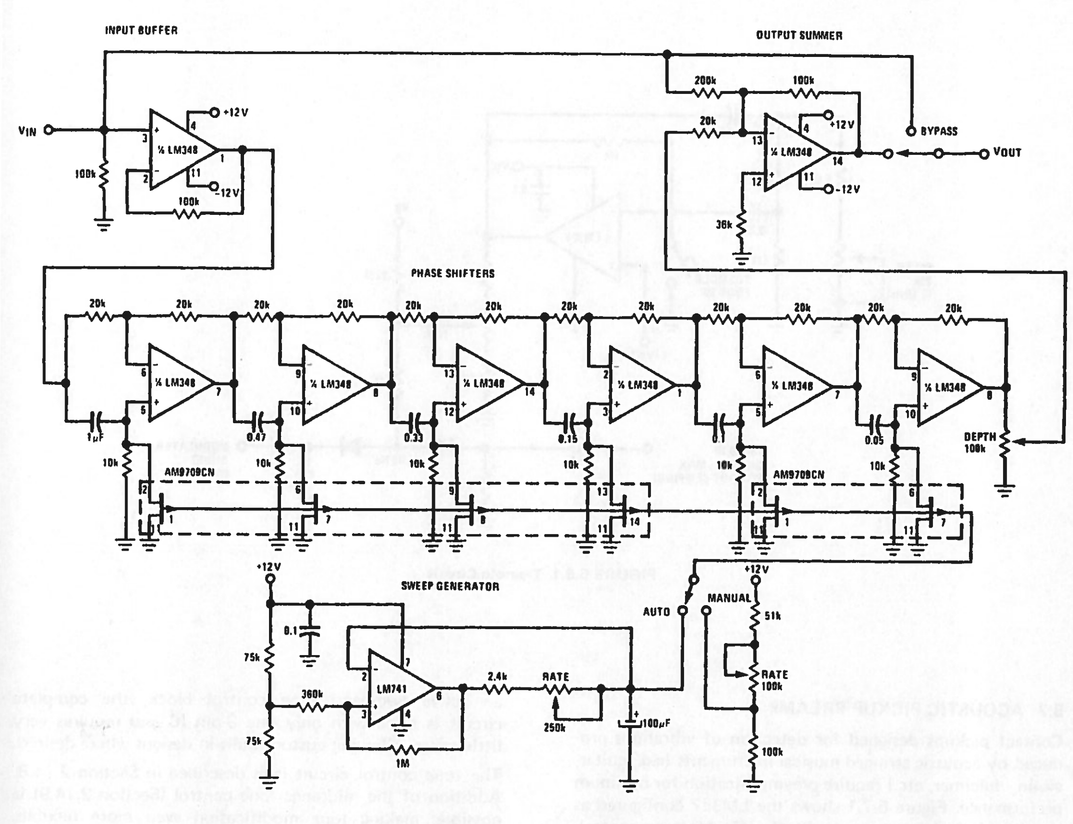

FIGURE 5.4.1 Phase Shifter

5.5 FUZZ

Two diodes in the feedback of a LM324 create the musical instrument effect known as "fuzz" (Figure 5.5.1). The diodes limit the output swing to ±0.7V by clipping the output waveform. The resultant square wave contains predominantly odd-ordered harmonics and sounds similar to a clarinet. The level at which clipping begins is controlled by the Fuzz Depth pot while the output level is determined by Fuzz Intensity.

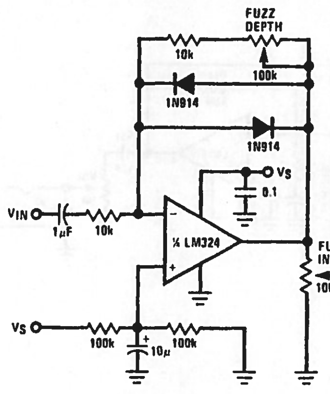

FIGURE 5.5.1 Fuzz Circuit

5.6 TREMOLO

Tremolo is amplitude modulation of the incoming signal by a low frequency oscillator. A phase shift oscillator (Figure 5.6.1) using the LM324 operates at an adjustable rate (5-10Hz) set by the SPEED pot. A portion of the oscillator output is taken from the DEPTH pot and used to modulate the "ON" resistance of two 1N914 diodes operating as voltage controlled attenuators. Care must be taken to restrict the incoming signal level to less than $0.6V_{p\text{-}p}$ or undesirable clipping will occur. (For signals greater than 25mV, THD will be high but is usually acceptable. Applications requiring low THD require the use of a light detecting resistor (LDR) or a voltage-controlled gain block. See Figure 4.8.9.)

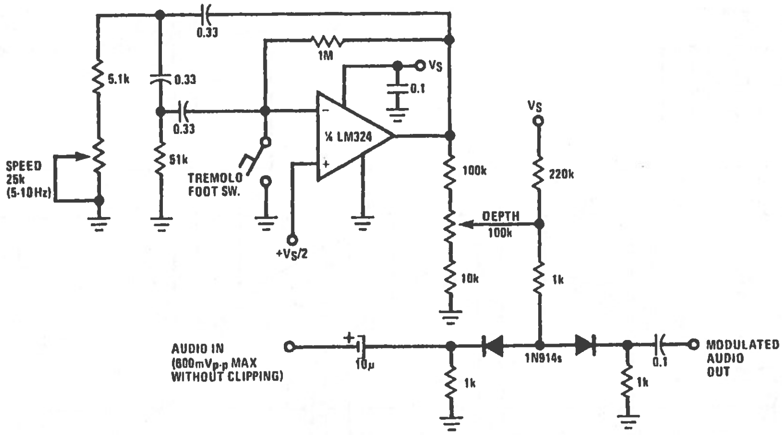

FIGURE 5.6.1 Tremolo Circuit

5.7 ACOUSTIC PICKUP PREAMP

Contact pickups designed for detection of vibrations produced by acoustic stringed musical instruments (e.g., guitar, violin, dulcimer, etc.) require preamplification for optimum performance. Figure 5.7.1 shows the LM387 configured as an acoustic pickup preamp, with Bass/Treble tone control, volume control, and switchable ±10dB gain select. The pickup used is the Ibanez "Bug," which is a flat response piezo-ceramic contact unit that is easy to use, inexpensive, and has excellent tone response. By using one half of the LM387 as the controllable gain stage and the other half as an active two-band tone control block, the complete circuit is done with only one 8-pin IC and requires very little space, allowing custom built-in designs where desired.

The tone control circuit is as described in Section 2.14.8. Addition of the midrange tone control (Section 2.14.9) is possible, making tone modification even more flexible. Switchable gain control of ±10dB is achieved using a DPDT, center off, switch to add appropriate parallelling resistors around the main gain setting resistors R_8 and R_6. Resistor R_9 is capacitively coupled (C_{14}) so as not to disturb dc conditions set up by R_8 and R_{10}.

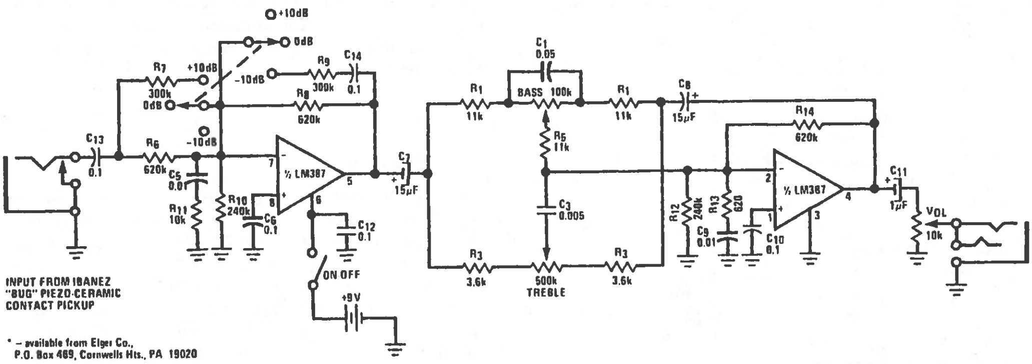

FIGURE 5.7.1 Acoustic Pickup Preamp

LM13600 — DUAL TRANSCONDUCTANCE AMPLIFIER

The LM13600 is similar to the more familiar op-amp with the major exception that the output is a current, the magnitude and polarity of which is defined by the product of the amplifier transconductance and the input voltage (i.e. $I_{OUT} = gmV_{IN}$). This output circuit is characterized as an infinite impedance current generator rather than the zero impedance voltage generator that represents the output of the conventional op-amp. The schematic for one half of the LM13600 is shown in Figure 5.8.1 and the circuit has a differential input stage with a tail current defined by the current injected into pin 1 (16). This current I_{ABC} controls the input stage transconductance, and the differential components of this current in Q_4 and Q_5 are mirrored into the output stage such that

$$I_{OUT} \doteq \frac{V_{IN} I_{ABC} q}{2KT} = gmV_{IN} \qquad (5.8.1)$$

where $gm \doteq 19.2\, I_{ABC}$ at room temperature.

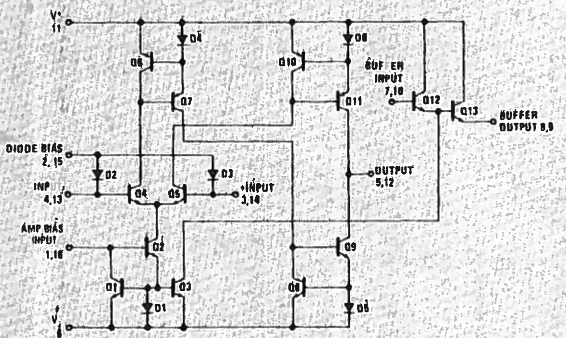

FIGURE 5.8.1 LM13600 Schematic

To use either section of the LM13600 as a low pass filter, we can configure it as shown in Figure 5.8.2

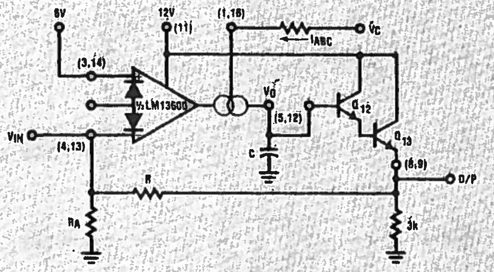

FIGURE 5.8.2 Voltage Controlled Low Pass Filter

By using the output Darlington buffer transistors Q_{12}, Q_{13}, the signal voltage V_{OUT} that appears at the capacitor C, is fed back to the amplifier inverting input, attenuated by the feedback resistors R_A and R, such that

$$V_{IN} = \frac{V_{OUT} R_A}{(R_A + R)} \qquad (5.8.2)$$

This input voltage will produce an output current I_{OUT} dependent on the control current magnitude I_{ABC}. From Equation (5.8.1)

$$I_{OUT} = gmV_{IN}$$
$$= gmV_{OUT} \frac{R_A}{(R_A + R)}$$

Therefore the amplifier output resistance R_o is given by

$$\frac{V_{OUT}}{I_{OUT}} \text{(Pin 5, 12)} = \frac{V_{OUT}}{gm\, V_{OUT}} \frac{(R_A + R)}{R_A}$$

$$\text{i.e. } R_{OUT} = \frac{(R_A + R)}{R_A gm} \qquad (5.8.3)$$

Since gm is controlled by I_{ABC}; the amplifier appears as a variable resistance R_{OUT} driving a capacitance C, which is a low pass filter configuration with a −3dB corner frequency given by

$$f_c \doteq \frac{1}{2\pi R_{OUT} C} \qquad (5.8.4)$$

As R_{OUT} is changed by I_{ABC}, the corner frequency is changed by the same amount.

5.8 NON-COMPLEMENTARY NOISE REDUCTION

One of the many contributors to the success of cassette recorders in becoming part of component hi-fi systems has been the Dolby B-type noise reduction scheme. This is a complementary system — i.e. the original material is encoded in such a way before recording that the complementary decoding process reduces the noise that can be added by the tape recorder. A weighted 9dB S/N ratio improvement is obtained without affecting the fidelity of the source. Unfortunately the Dolby B system cannot improve the S/N ratio of the original material — it simply prevents further degradation by the recorder. So what can we do about old and favorite recordings made before Dolby circuits were widely available and whose value is marred by the ever present tape hiss? Also, for many of us, FM broadcasts still leave something to be desired in the attainment of low background noise levels. In either case the alternative is a non-complementary noise reduction system that operates to remove the noise *already present* in the source.

This can be done by restricting the system bandwidth (down to about 800Hz) in the absence of programme material. For a typical cassette source this will improve the S/N ratio by about 14dB (CCIR/ARM weighted). When programme material is present, with sufficient amplitude in the appropriate frequency range to mask the noise, the system bandwidth is automatically opened up. The degree to which the bandwidth can be opened depends largely on the masking effect of the programme material which, in turn, depends on the pitch and loudness of the noise. For this reason, the detector circuit used to determine the signal amplitude for which the audio bandwidth can be opened should include frequency response shaping networks. While several such audio processing systems have been built, Ref. 1, 2, 3, with varying degrees of complexity, the introduction of a dual transconductance amplifier, the LM13600 (see box), has made the implementation of automatically variable filters both simple and economical.

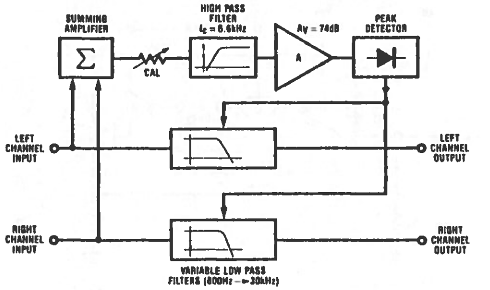

FIGURE 5.8.3 Noise Processing System Block Diagram

A block diagram containing the necessary functions is shown in Figure 5.8.3. This format is suitable for a stereo system with two unity gain current controlled filters operating with a common signal. A common control signal path, with frequency shaping, is used to prevent possible loss of stereo image which could occur if the bandwidth of the L and R channels were different.

A single LM13600 is used for both variable filters with the advantage that both channels will be inherently well matched and no set-up adjustments are required. When the control current to each filter from the peak detector is 4μA, the audio bandwidth is 800Hz, increasing to 20kHz, when the control current is 100μA. Since both filters are equivalent to single section RC low pass filters, they have a 6dB/octave roll-off slope above the cut-off frequency.

The response times of the filters for a bandwidth change are determined by the detector circuit time constants — in this case a 1msec attack time is used to obtain rapid opening of the bandwidth with programme transients, and a 50msec decay time to prevent the filters cutting off the natural reverberation following a music transient.

The control path sums the L & R inputs into a high pass filter. this filter has a corner frequency of 6.6kHz and a 12dB/octave roll off slope to ensure that proper weighting is given to the programme material in terms of its noise masking ability. A single adjustment for the system, a sensitivity control, also precedes the high pass filter and sets the summed input level such that the noise in the source (during a blank period in the programme) is just beginning to open the audio channel filters.

A practical circuit for the noise processor is shown in Figure 5.8.4 and is designed to be included in the tape monitor loop of a hi-fi system. The gain in the audio channel is unity, with an 88dB S/N ratio for a $775mV_{rms}$ input level and a 30kHz system bandwidth.

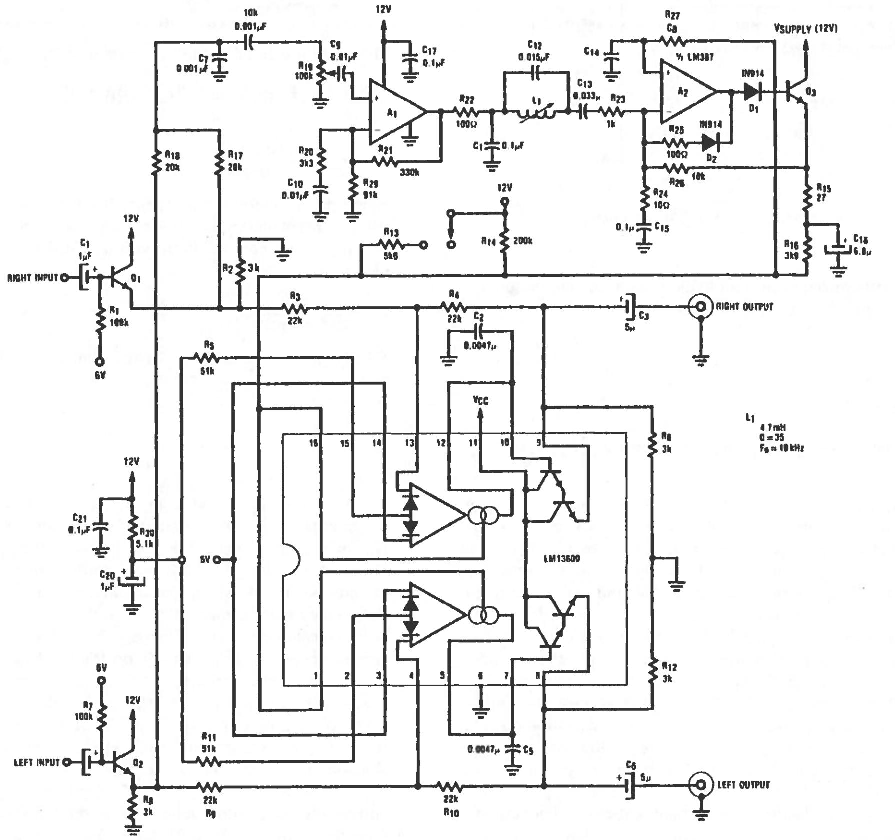

FIGURE 5.8.4 Stereo Noise Reduction Circuit

Two input buffer transistors Q_1 and Q_2 are used to provide a high input impedance and to allow the audio signals to be summed into the sensitivity potentiometer R_{19} with minimum crosstalk. To ensure that the programme signals above the noise floor are capable of opening the audio filters, an LM387 is used to give two stages of gain before the detector capacitor C_{16}. These amplifier stages also perform necessary detector signal filtering so that masking is obtained. To obtain a −12dB/octave slope, two RC sections are used, $R_{23}C_{13}$ and $R_{20}C_{10}$. For a half power frequency f_c of 6.6kHz with two stages having identical corner frequencies f_o, we use Equation (5.8.5)

$$f_c = f_o \sqrt{10^{0.3/n} - 1} \tag{5.8.5}$$

For n = 2

$f_o = 0.643 f_c = 4.24\text{kHz}$

Therefore $R_{23}C_{13} = R_{20}C_{10} = 37.5\mu S$.
Above the filter corner frequency the midband gain is obtained with stage gains of 40dB for a total of 80dB.

$$|A_{V1}| = 1 + \frac{R_{21}}{R_{20}} \tag{5.8.6}$$

$$|A_{V2}| = \frac{R_{26}}{R_{24}} \tag{5.8.7}$$

The detector time constants are set by charging C_{16} through the resistor R_{15} and discharging C_{16} through the resistor R_{16} connected to the filter control pins (pins 1 & 16) of the LM13600. To bypass the noise reduction effect, a 5.6kΩ resistor R_{13} is switched into the control path forcing the filter to a fixed B-W in excess of 200kHz.

As shown, the circuit gives a 14dB S/N ratio improvement (weighted) with a distortion level of 0.13% with the rated input level (0.775mV$_{rms}$).

To display the action of the noise processor, the circuit in Figure 5.8.5(a) can be used. When connected to the detector capacitor C_{16}, the LM3915 will illuminate successive L.E.D.s for each 3dB increase in detected signal level (floating dot mode), providing a dynamic display of the instantaneous audio bandwidth. A suitable power supply, utilizing a 200mA filament transformer, is shown in Figure 5.8.5(b).

REFERENCES:

1. Burwen, Richard S., "A Dynamic Noise Filter For Mastering," *Audio*, June 1972, page 29.
2. Hellyer, H. W., "Noise Reduction Techniques," *Audio*, October 1972, page 18.
3. Scott, H. H., "Dynamic Noise Processor," *Electronics*, December 1947, page 96.

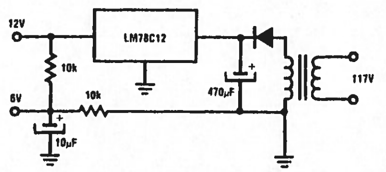

FIGURE 5.8.5(b) Power Supply

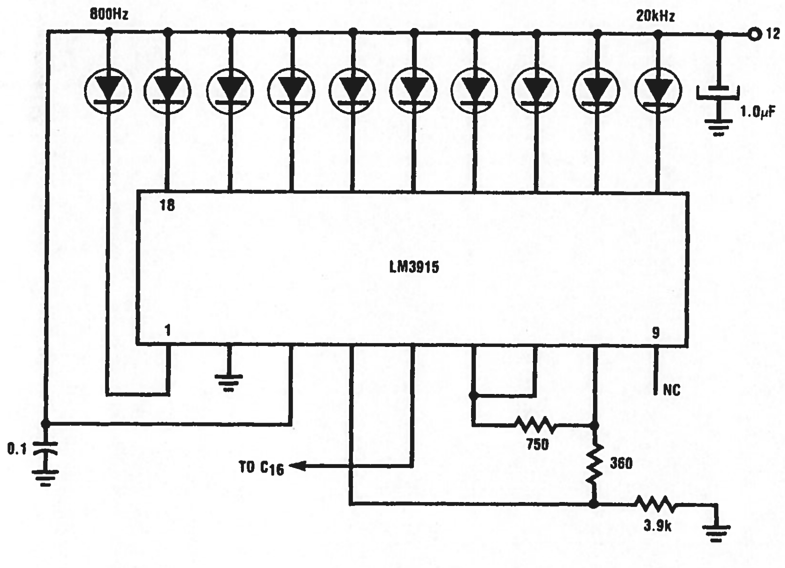

FIGURE 5.8.5(a) Audio Bandwidth Display

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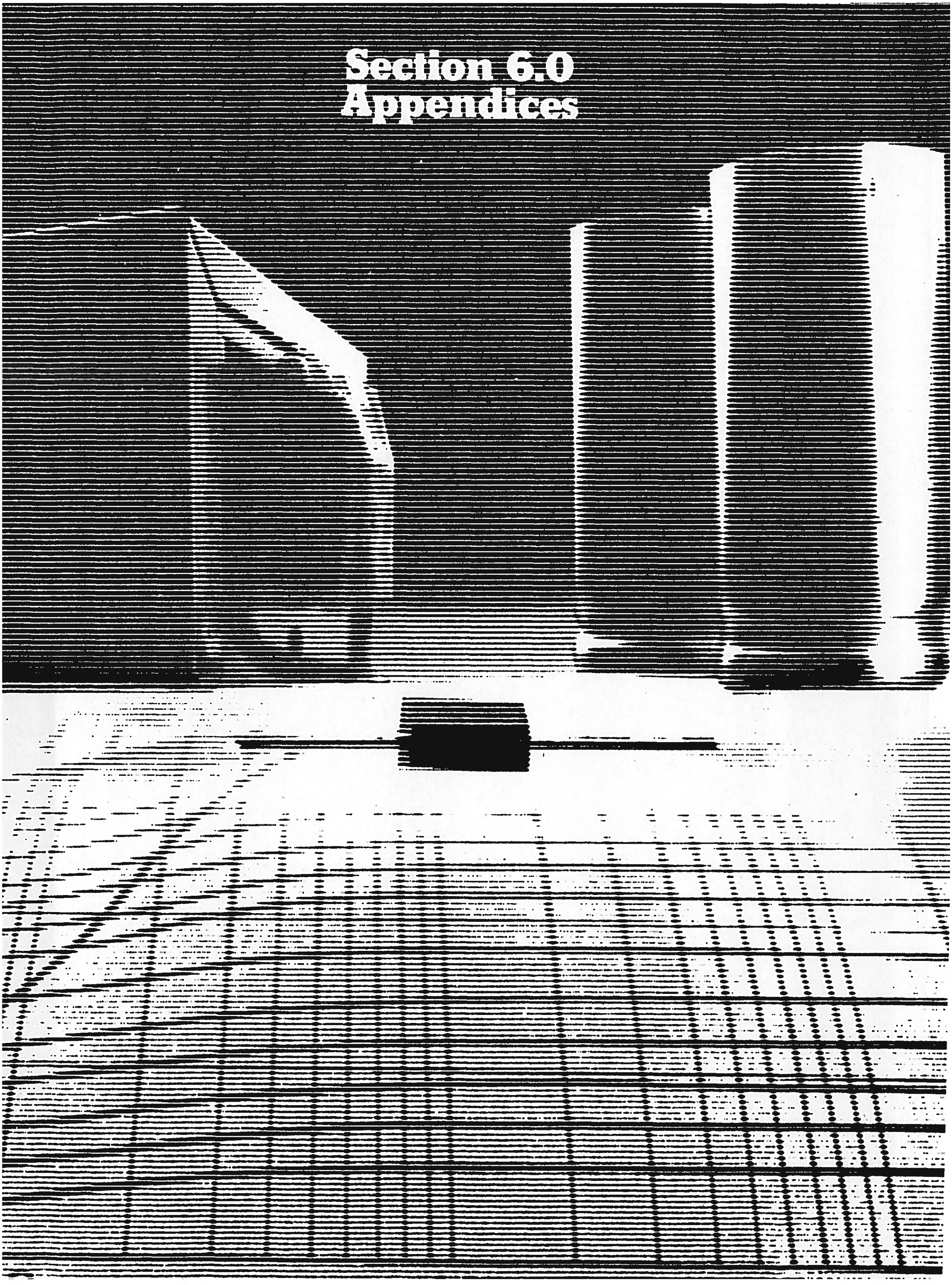
Section 6.0
Appendices



6.0 Appendices

A1.0 POWER SUPPLY DESIGN

A1.1 Introduction

One of the nebulous areas of power IC data sheets involves the interpretation of "absolute maximum ratings" as opposed to "operating conditions." The fact that parameters are specified at an operating voltage quite a few volts below the absolute maximum is not nearly so important in "garden variety" op amps as in power amps — because a key spec of any power amplifier is how much power it can deliver, a spec that is a strong function of the supply. Indeed P_o is approximately proportional to the square of supply voltage. Since many audio ICs are powered from a step down transformer off the $120V_{AC}$ line, the "absolute maximum voltage" is an attempt to spec the highest value the supply might ever reach under power company overvoltages, transformer tolerances, etc. This spec says the IC will not die if taken to its "absolute maximum rating." Operating voltage, on the other hand, should be approximately what a nominal supply will sag under load at normal power company voltages. Some audio amplifiers are improperly specified at their "absolute maximum voltages" in order to give the illusion of large output power capability. However, since few customers regulate the supply voltage in their applications of audio ICs, this sort of "specsmanship" can only be termed deceptive.

A1.2 General

This section presents supply and filter design methods and aids for half-wave, full-wave center tap, and bridge rectifier power supplies. The treatment is sufficiently detailed to allow even those unfamiliar with power supply design to specify filters, rectifier diodes and transformers for single-phase supplies. A general treatment referring to Figure A1.1 is given, followed by a design example. No attempt is made to cover multiphase circuits or voltage multipliers. For maximum applicability a regulator is included, but may be omitted where required.

A1.3 Load Requirements

The voltage, current, and ripple requirements of the load must be fully described prior to filter and supply design. Actually, so far as the filter and supply are concerned, the load requirements are those at the regulator input. (See Figure A1.1.) Therefore, V_{IN} and I_{IN} become the governing conditions, where:

$I_{IN} = I_O + I_Q$,	output current plus regulator quiescent current
$I_{IN(MAX)} \approx I_{O(MAX)}$,	full-load operating current
$I_{IN(MIN)} \leq I_Q$,	no-load or minimum operating current; could be near zero
$V_{IN(PK)} = V_M$,	maximum permissible instantaneous no-load filter output voltage equal to peak value of transformer secondary voltage at highest design line voltage V_{PRI}; limited by absolute maximum regulator input voltage
$V_{IN} > V_O$,	nominal DC voltage input to the regulator, usually 2 to 15V higher than V_O
$V_{IN(MIN)} \approx V_O + 2V$,	minimum instantaneous full-load filter output voltage including ripple voltage; limited by minimum regulator input voltage to insure satisfactory regulation ($V_O + V_{dropout}$) or minimum regulator input voltage to allow regulator start-up under full load or upon removal of a load short circuit
r_f	RMS ripple factor at filter output expressed as a percentage of V_{IN}; limited by maximum permissible ripple at load as modified by the ripple rejection characteristics of the regulator

A1.4 Filter Selection, Capacitor or Inductor-Input

For power supplies using voltage regulators, the filter will most often use capacitor input; therefore, emphasis will be placed upon that type of filter in following discussions. Notable differences between the two types of filters are that the capacitor input filter exhibits:

1. Higher DC output voltage
2. Poorer output voltage regulation with load variation
3. Higher peak to average diode forward currents

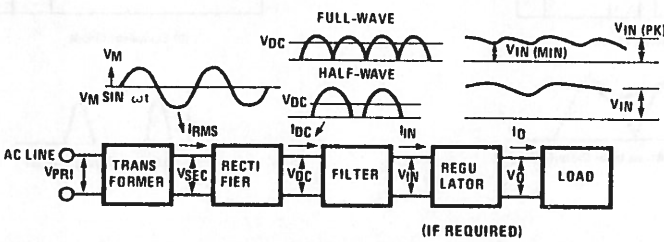
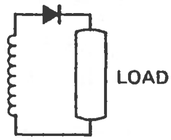
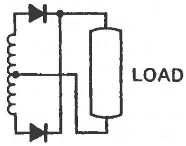
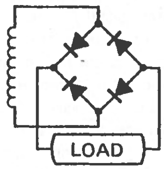
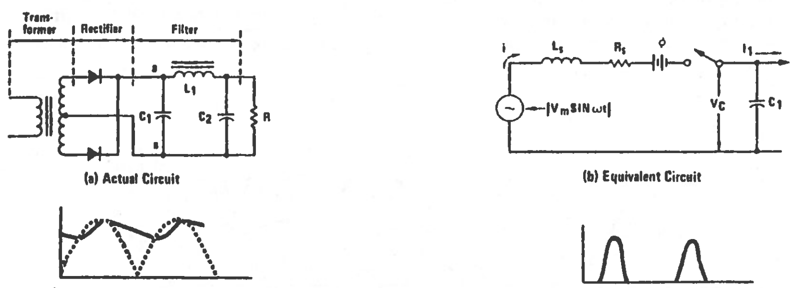


FIGURE A1.1 Power Supply Block Diagram, General Case

TABLE A1.1 Summary of Significant Rectifier Circuit Characteristics, Single Phase Circuits
Capacitive Data is for $\omega C R_L$ = 100 & R_S/R_L = 2% (higher values)
and for $\omega C R_L$ = 10 & R_S/R_L = 10% (lower values)

	Single Phase Half Wave			Single Phase Full Wave Center Tap			Single Phase Full Wave Bridge		
Rectifier Circuit Connection									
Voltage Waveshape to Load of Filter									
CHARACTERISTIC LOAD	R	L	C	R	L	C	R	L	C
Average Diode Current $I_{F(AVG)}/I_{O(DC)}$	1	1	1	0.5	0.5	0.5	0.5	0.5	0.5
Peak Diode Current $I_{FM}/I_{F(AVG)}$	3.14	-	8 5.2	3.14	2	10 6.2	3.14	2	10 6.2
Diode Current Form Factor, F = $I_{F(RMS)}/I_{F(AVG)}$	1.57	-	2.7 2	1.57	1.41	3 2.2	1.57	1.41	3 2.2
RMS Diode Current $I_{F(RMS)}/I_{O(DC)}$	1.57	-	2.7 2	0.785	0.707	1.35 1.1	0.785	0.707	1.35 1.1
RMS Input Voltage per Transformer Leg $V_{SEC}/V_{IN(DC)}$	2.22	2.22	0.707	1.11	1.11	0.707	1.11	1.11	0.707
Transformer Primary VA Rating VA/P_{DC}	3.49	-	-	1.23	1.11	-	1.23	1.11	-
Transformer Secondary VA Rating VA/P_{DC}	3.49	-	-	1.75	1.57	-	1.23	1.11	-
Total RMS Ripple %	121	-	-	48.2	-	-	48.2	-	-
Rectification Ratio (Conversion Efficiency) %	40.6	-	-	81.2	100	-	81.2	100	-



(a) Actual Circuit

(b) Equivalent Circuit

(c) Voltage Across Input Capacitor C1

(d) Current Through Diodes

FIGURE A1.2 Actual and Equivalent Circuits of Capacitor-Input Rectifier System, Together with Oscillograms of Voltage and Current for a Typical Operating Condition

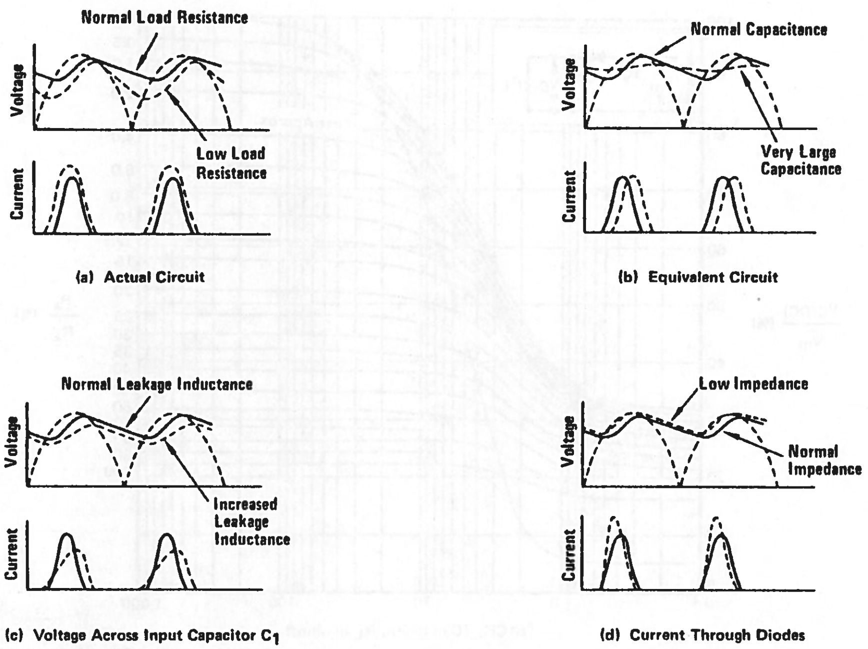


FIGURE A1.3 Effects of Circuit Constants and Operating Conditions on Behavior of Rectifier Operated with Capacitor-Input Filter

4. Lower diode PIV rating requirements
5. Very high diode surge current at turn-on
6. Higher peak to average transformer currents

The voltage regulator overcomes disadvantage (2) while semiconductor diodes of moderate price meet most of the peak and surge requirements except in supplies handling many amperes. Still, it may be necessary to balance increased diode and transformer cost against the alternative of a choke-input filter. In power supply designs employing voltage regulators, it is assumed that only moderate filter output regulation and ripple are required. Therefore, a capacitor input filter would exhibit peak currents considerably lower than indicated in the comparison of Table A1.1.

A1.5 Filter Design, Capacitor-Input

Figure A1.2 shows a full-wave, capacitor-input (filter) rectifier system with typical voltage and current waveforms. Note that ripple is inevitable as the capacitor discharges approximately linearly between voltage peaks. Figure A1.3 shows the effects on DC voltage, ripple, and peak diode current under varying conditions of load resistance, input capacitance, series diode and transformer resistance R_S, and transformer leakage inductance. The most practical design procedure for capacitor-input filters is to use the graphs of Figures A1.4-A1.7. Note, however, that these include the effects of diode dynamic resistance within R_S. Diode forward drop is not included, and must be subtracted from the transformer secondary voltage. A good rule of thumb is to subtract 0.7V from the transformer voltage and assume diode dynamic resistance is insignificant (0.02Ω at $I_F = 1A$, 0.26Ω at $I_F = 100mA$); ordinarily the transformer resistance will overshadow diode dynamic resistance.

Figures A1.4 and A1.5 show the relationship between peak AC input voltage and DC output voltage as a relation to load resistance R_L, series circuit resistance R_S, and filter input capacitance C. Figure A1.4 is for half-wave rectifiers and Figure A1.5 is for full-wave rectifiers. Note that the horizontal axis is labeled in units of $\omega C R_L$ where:

ω = AC line frequency in Hertz x 2π

C = value of input capacitor in Farads

$R_L = V_{IN}/I_{IN} \approx V_O/I_O$, equivalent load resistance in Ohms

R_S = total of diode dynamic resistance, transformer secondary resistance, reflected transformer primary resistance, and any added series surge limiting resistance

The major design trade-off encountered in designing capacitor-input filters is that between achieving good voltage regulation with low ripple and achieving low cost. Referring to Figures A1.4 and A1.5:

1. Good regulation means $\omega C R_L \approx 10$.
2. Low ripple may mean $\omega C R_L > 40$.
3. High efficiency means $R_S/R_L < 0.02$.
4. Low cost usually means low surge currents and small C.
5. Good transformer utilization means low VA ratings, best with full-wave bridge FWB circuit, followed by full-wave center tap FWCT circuit.

In most cases, a minimum capacitance accomplishing a reasonable full-load to no-load regulation is preferable for low cost. To achieve this, use an intercept with the upper

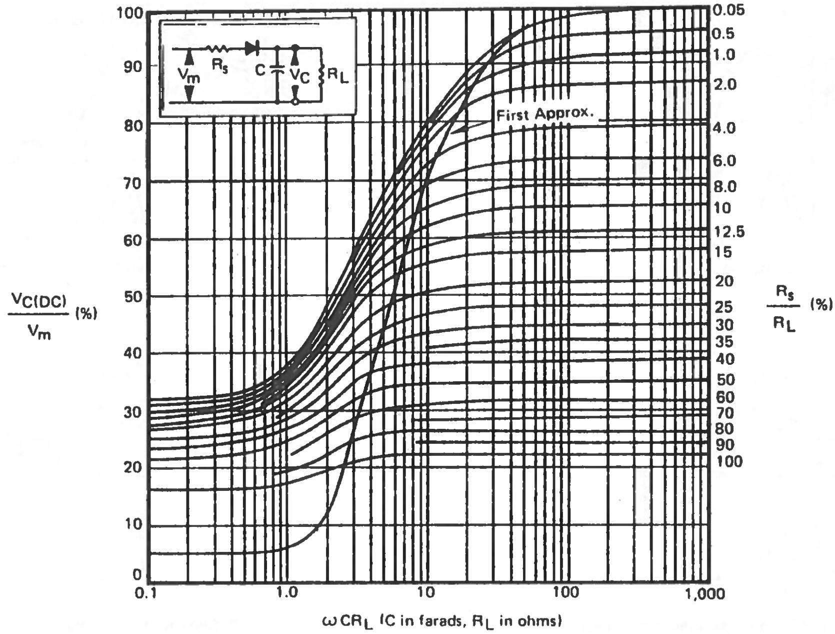


FIGURE A1.4 **Relation of Applied Alternating Peak Voltage to Direct Output Voltage in Half-Wave Capacitor-Input Circuits (From O. H. Schade, Proc. IRE, vol. 31, p. 356, 1943.)**

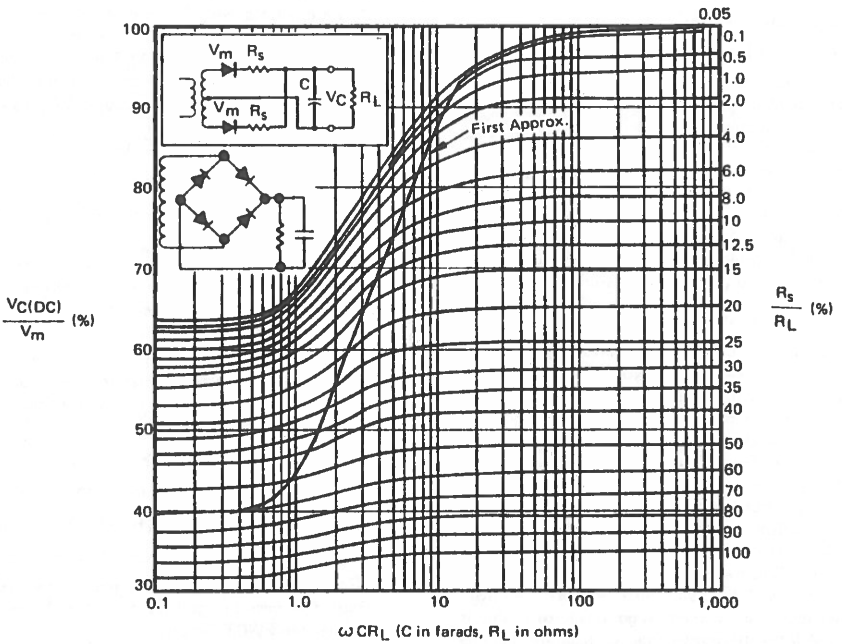


FIGURE A1.5 **Relation of Applied Alternating Peak Voltage to Direct Output Voltage in Full-Wave Capacitor-Input Circuits (From O. H. Schade, Proc. IRE, vol. 31, p. 356, 1943.)**

knee of the curves in Figures A1.4 and A1.5. Occasionally, a minimum value filter capacitor will not result in a lower cost system. For example, increasing the value of C may allow higher R_S/R_L to result in lower surge and RMS currents, thus allowing lower cost transformers and diodes. Be sure that capacitors used have adequate ripple current ratings.

Design procedure is as follows:

1. Assuming that V_O, I_O, ω, and load ripple factor r_f have been established and an appropriate voltage regulator has been selected, we know or can determine:

   $\omega = 2\pi f = 377\,rad/sec$ for 60Hz line

   $r_{f(in)} = r_{f(out)} \times$ ripple reduction factor of selected regulator

   $V_{IN(PK)} \leq$ Max V_{IN} for the selected regulator; allow for highest line voltage likely to be encountered

   $V_{IN(MIN)} \approx V_O + 2V$; allow for lowest line voltage

   $V_{IN(DC)}{}^{+} = V_{IN}$, usually 2-15V above V_O; if chosen midway between $V_{IN(PK)}$ and $V_{IN(MIN)}$ or slightly below that point, will allow for greatest ripple voltage

   $I_{IN} \approx I_O$ for full load

   $I_{IN(MIN)} = I_Q$ for open load

   $R_L = V_{IN(DC)}/I_{IN}$

   $R_{L(MIN)} = V_{IN(MIN)}/I_{IN}$

2. Set $V_M \leq V_{IN(PK)}$ and calculate $V_{IN(DC)}/V_{IN}$. Enter the graph of Figure A1.4 or A1.5 at the calculated $V_{IN(DC)}/V_M$ to intercept one of the R_S/R_L = constant lines. Either estimate R_S at this time or intercept the curve marked "First Approximation."
3. Drop vertically from the intercept of Step (2) to the horizontal axis and read $\omega C R_L$. Calculate C, allowing for usual commercial tolerance on capacitors of +100, –50%.

   If $V_{IN(DC)}$ is midway between $V_{IN(PK)}$ and $V_{IN(MIN)}$, the supply can present maximum ripple to the regulator. A low value of C is then practical. If $V_{IN(DC)}$ is near $V_{IN(MIN)}$, regulator power dissipation is low and supply efficiency is high; however, ripple must be low, requiring large C.
4. Determine ripple factor r_f from Figure A1.6. Make certain that the ripple voltage does not drop instantaneous V_{IN} below $V_{IN(MIN)}$.

   The ripple factor could determine minimum required C if ripple is the limiting factor instead of voltage regulation. Again, allow for –50% tolerance on the capacitor.

$$V_{ripple(pk)} = \sqrt{2}\,\frac{r_f}{100}\,V_{IN(DC)}$$

A1.6 Diode Specification

Find diode requirements as follows:

1. $I_{F(AVG)} = I_{IN(DC)}$ for half-wave rectification

   $= I_{IN(DC)}/2$ for full-wave rectification
2. Determine peak diode current ratio from Figure A1.7; remember to allow for highest operating line voltage and +100% capacitor tolerance.

   $I_{FM} = I_{FM}/I_{F(AVG)} \times I_{IN(DC)}$ for half-wave

   $= I_{FM}/I_{F(AVG)} \times I_{IN(DC)}/2$ for full-wave
3. Determine diode surge current requirement at turn-on of a fully discharged supply when connected at the peak of the highest expected AC line waveform. Surge current is:

$$I_{SURGE} = \frac{V_M}{R_S + ESR}$$

   where ESR = effective series resistance of capacitor.
4. Find required diode PIV rating from Figure A1.8. Actually, required PIV may be considerably more than the value thus obtained due to noise spikes on the line. See Section A1.9 for details on transient protection. Remember that the PIV for the diodes in the FWB configuration are one half that of diodes as found in FWCT or HW rectifier circuits.

The diodes may now be selected from diode manufacturers' data sheets. If calculated surge current rating or peak current ratings are impractically high, return to Step A1.5(2) and choose a higher R_S/R_L or lower C. Conversely, it may be practical to choose lower R_S/R_L or higher C if diode current ratings can be practically increased without adverse effect on transformer cost; the result will be higher supply efficiency.

A1.7 Transformer Specification

A decision may have been made at Step A1.5(2) as to using half-wave or full-wave rectification. The half-wave circuit is often all that is required for low current regulated supplies; it is rarely used at currents over 1A, as large capacitors and/or high surge currents are dictated. Transformer utilization is also quite low, meaning that higher VA rating is required of the transformer in HW circuits than in FW circuits. (See VA ratings of Table A1.1.)

Half-wave circuits are characterized by low $V_{IN(DC)}/V_M$ ratio, or very large C required (about 4 times that required for FW circuits, high ripple, high peak to average diode and transformer current ratios, and poor transformer utilization). They do, however, require only one diode.

Full-wave circuits are characterized by high $V_{IN(DC)}/V_M$ ratio, low C value required, low ripple, low peak to average diode and transformer current ratios, and good transformer utilization. They do require two diodes in the center-tap version, while the bridge configuration with its very high transformer utilization requires four diodes.

The information necessary to specify the transformer is:

1. Half-wave, full-wave CT or full-wave bridge circuit
2. Secondary V_{RMS} per transformer leg, $(V_M + 0.7^*)/\sqrt{2}$, from Section A1.5
3. Total equivalent secondary resistance including reflected primary resistance from Section A1.5
4. Peak, average, and RMS diode or winding currents from Sections A1.6(1) and -(2), and VA ratings.

*1.4 for full-wave bridge circuit.

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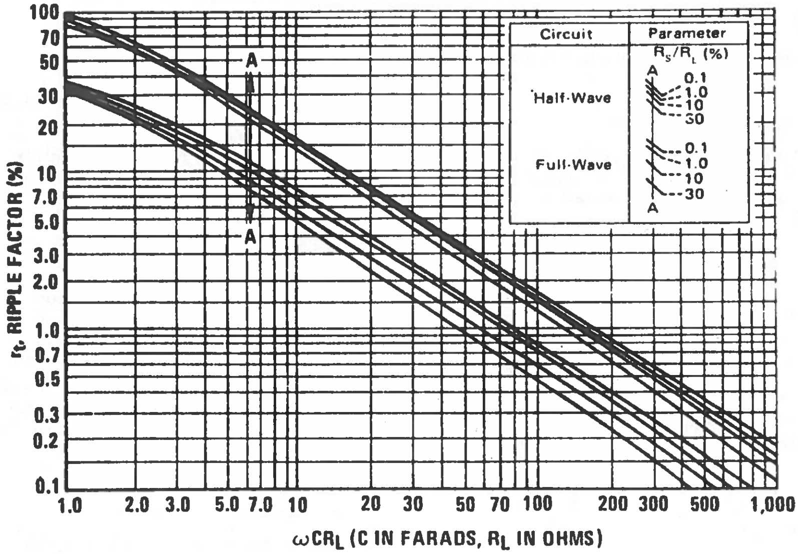

FIGURE A1.6 Root-Mean-Square Ripple Voltage for Capacitor-Input Circuits (From O. H. Schade, Proc. IRE, vol. 31, p. 356, 1943.)

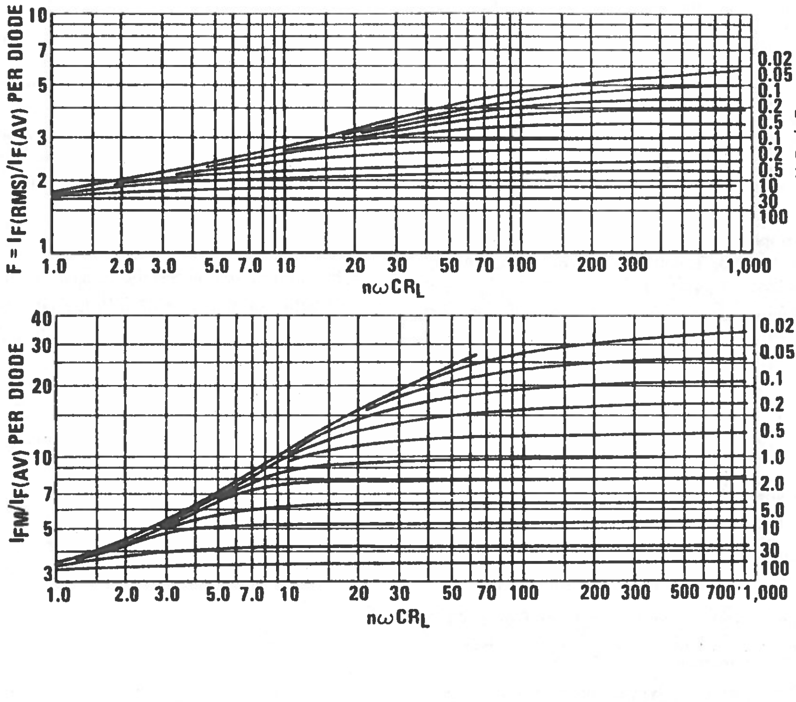

FIGURE A1.7 Relation of RMS and Peak-to-Average Diode Current in Capacitor-Input Circuits (From O. H. Schade, Proc. IRE, vol. 31, p. 356, 1943.)

Transformer VA rating and secondary current ratings are determined as follows:

		FWB	FWCT	HW
$I_{RMS(SEC)}$	=	$I_{IN(DC)}F/\sqrt{2}$	$I_{IN(DC)}F/2$	$I_{IN(DC)}F$
VA_{SEC}	=	$V_{RMS}I_{RMS}$	$2V_{RMS}I_{RMS}$	$V_{RMS}I_{RMS}$
VA_{PRI}	=	VA_{SEC}	$VA_{SEC}/\sqrt{2}$	VA_{SEC}

where: $F = I_{F(RMS)}/I_{IN(DC)}$
= form factor from Figure A1.7
V_{RMS} = secondary RMS voltage per leg

A1.8 Additional Filter Sections

Occasionally, it is desirable to add an additional filter to reduce ripple. When this is done, an LC filter section is cascaded with the single C section filter already designed. If the inductor is of low resistance, the effect on output voltage is small. The additional ripple reduction may be determined from Figure A1.9.

A1.9 Transient Protection

Often the PIV rating of the rectifier diodes must be considerably greater than the minimum value determined from Figure A1.8. This is due to the likely presence of high-voltage transients on the line. These transients may be as high as 400V on a 115V line. The transients are a result of switching inductive loads on the power line. Such loads could be motors, transformers, or could even be caused by SCR lamp dimmers or switching-type voltage regulators, or the reverse recovery transients in rectifying diodes. As the transients appearing on the transformer primary are coupled to the secondary, the rectifier diodes may see rather high peak voltages. A simple method of protecting against these transients is to use diodes with very high PIV. However, high-current diodes with very high PIV ratings can be expensive.

There are several alternate methods of protecting the rectifier diodes. All rely on the existence of some line impedance, primary transformer resistance or secondary circuit resistance. See Figure A1.10 for the system circuit.

The several methods of transient protection rely on shunting the transient around the rectifier diodes to dissipate the transient energy in the series circuit resistance and the protective device. The usual protection methods are:

1. Series resistor at the primary with shunt capacitor across the primary winding — see Figure A1.10

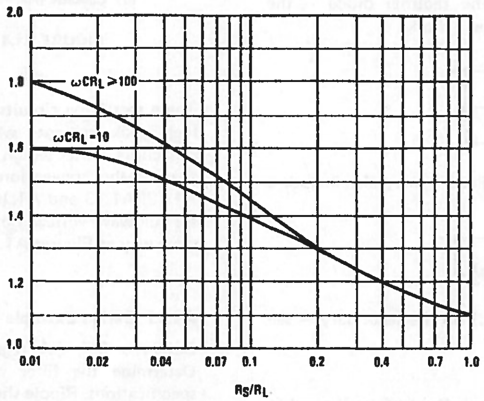

FIGURE A1.8 Ratio of Operating Peak Inverse Voltage to Peak Applied AC for Rectifiers Used in Capacitor-Input, Single-Phase, Filter Circuits

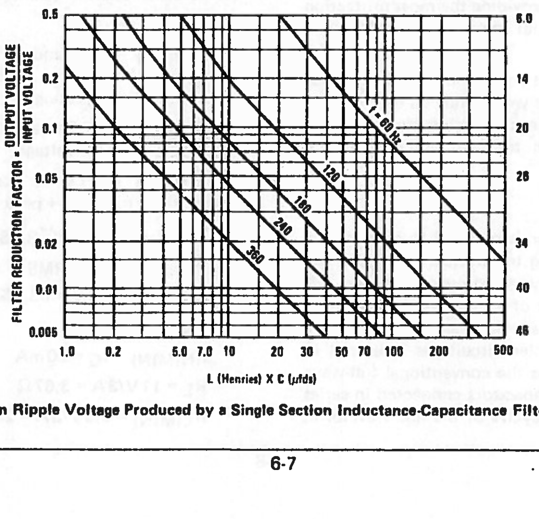

FIGURE A1.9 Reduction in Ripple Voltage Produced by a Single Section Inductance-Capacitance Filter at Various Ripple Frequencies

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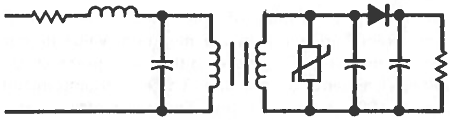
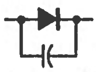
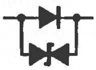


FIGURE A1.10 Transformer/Filter Circuit Showing Placement of Transient Protection Components

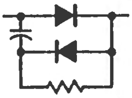
2. Series inductance at the primary, possibly with a shunt capacitor across the primary – see Figure A1.10
3. Shunt capacitor on the secondary – see Figure A1.10
4. Capacitor shunt on the rectifier diode – transient power is thus dissipated in circuit series resistance.



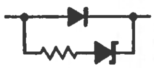
5. Surge suppression varactor shunt on the rectifier diode – this scheme is quite effective, but costly.



6. Dynamic clipper shunt on the rectifier diode – the clipper consists of an R, a C and a diode.



7. Zener shunt on the rectifier diode – may also include a series resistance.



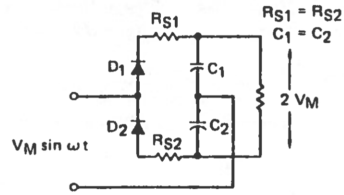
8. Shunt varistor (e.g., GE MOVs) on the secondary – see Figure A1.10.

Of the several protective circuits:

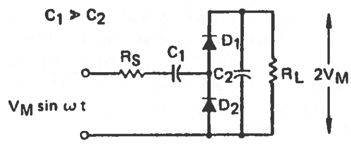
- (1), (2), (3) and (4) are least costly, but are limited in their utility to incomplete protection.
- (4) is probably the circuit providing the most protection for the money and is all that may be required in low-current regulated supplies.
- (5), (6), (7) and (8) are most costly, but provide greatest protection. Their use is most worthwhile on high current supplies where high PIV ratings on high-current diodes is costly, or where very high transient voltages are encountered.

A1.10 Voltage Doublers

Occasionally, a voltage doubler is required to increase the voltage output from an existing transformer. Although the doubler circuits will provide increased output voltage, this is accomplished at the expense of an increased component count. Specifically, two filter capacitors are required. There are two basic types of doubler circuits as indicated in Figure A1.11. Figure A1.11a is the conventional full-wave doubler circuit wherein two capacitors connected in series are charged on alternate half cycles of the line waveform. Figure A1.11b is a half-wave doubler circuit wherein C_2 is partially charged on one half cycle and then on the second half cycle the input voltage is added to provide a doubling effect. C_1 is normally considerably larger than C_2. The advantage of the half-wave circuit is that there is a common input and output terminal; disadvantages are high ripple, low I_O capability, and low V_{OUT}.



(a) Conventional Full-Wave Voltage Doubling Circuit



(b) Cascade (Half-Wave) Voltage Doubling Circuit

FIGURE A1.11 Voltage Doubler Circuits

These rectifying circuits, being capacitively loaded, exhibit high peak currents when energy is transferred to the capacitors. Filter design for the doubler circuits is similar to that of the conventional capacitor filter circuits. Figures A1.12, A1.13 and A1.14 provide the necessary design aids for full-wave voltage doubler circuits. They are used in the same way as Figures A1.5, A1.6 and A1.7.

A1.11 Design Example

Design a 5V, 3A regulated supply using an LM123K. Determine the filter values and transformer and diode specifications. Ripple should be less than $7 mV_{RMS}$. Assume 60dB ripple reduction from typical curves.

1. Establish operating conditions:

   $\omega = 377$ rad/sec

   $V_{IN(PK)} = 18V$ and 10% high line voltage; this allows some 2V headroom before reaching the 20V absolute maximum V_{IN} rating of the LM123K

   $V_{IN(MIN)} = 7.5V$ at 10% low line voltage including effects of ripple voltage

   $V_{IN(DC)} = 11V$ at nominal line voltage; chosen to exceed $V_{IN(MIN)}$ + peak ripple voltage

   $V_{ripple(out)} \leqslant 7 mV_{RMS}$

   $V_{ripple(in)} \leqslant 7 V_{RMS}$

   $r_{f(in)} \leqslant 7V/11V = 63.5\%$

   $I_{IN} = 3A$

   $I_{IN(MIN)} = I_Q = 20mA$

   $R_L = 11V/3A = 3.67\Omega$

   $R_{L(MIN)} = 7.5V/3A = 2.5\Omega$

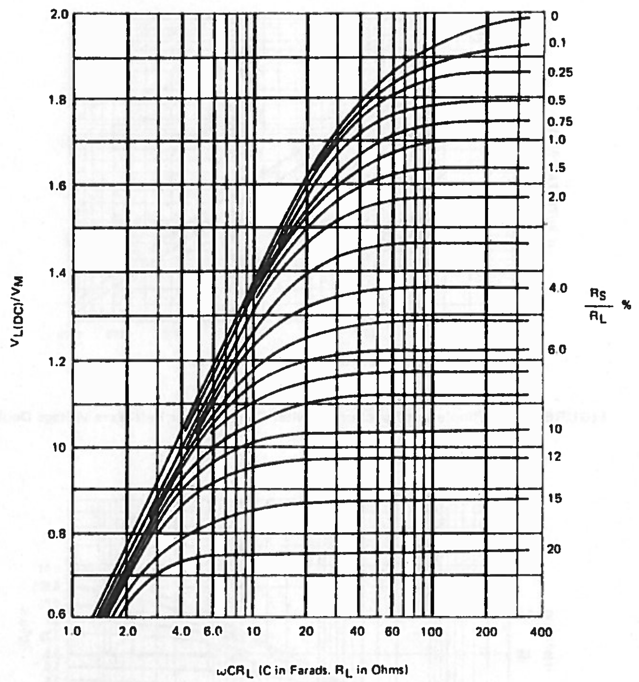


FIGURE A1.12 Output Voltage as a Function of Filter Constants for Full-Wave Voltage Doubler for Full-Wave Voltage Doubler

2. Set:

   V_M = 16.3V nominal, which is 18V – 10% line variation

   $V_{IN(DC)}/V_M = 11/16.3 = 0.67$

   Assume full-wave bridge rectification because of the high current load. Enter the graph of Figure A1.5 at $V_{IN(DC)}/V_M = 0.67$ to intercept the "First Approximation" curve.

3. Drop down to the horizontal axis to find $\omega C R_L = 3.33$. Thus, $R_S/R_L \approx 13\%$, or $R_S = 0.4\Omega$ is allowable.

   $$C = \frac{3.33}{3.67 \times 377} = 2400\mu F$$

   (4800μF allowing for –50% capacitor tolerance)

4. Ripple factor is 15% from Figure A1.6. Ripple is then

   $$V_{ripple(pk)} = \sqrt{2} \times 0.15 \times 11 = 2.33V \text{ pk.}$$

5. Checking for $V_{IN(MIN)}$:

   V_M = 16.3V or, allowing for 10% low line voltage, 14.8V

   $V_{IN(DC)} = 14.8 \times 0.67 = 9.91V$

   Subtracting peak ripple, $V_{IN(MIN)} = 9.91 - 2.33 = 7.6V$ which is within specifications

   In fact, all requirements have been met.

6. Diode specifications are:

   $$I_{F(AVG)} = \frac{I_{IN(DC)}}{2} = 1.5A \text{ for FW rectifiers}$$

   $I_{FM} = 8 \times 1.5A = 12A$, from figure A1.7, allowing C = 100% high, for commercial tolerances

   $I_{SURGE} = 18V/0.48\Omega = 37.5A$, worst case with 10% high line, neglecting capacitor ESR

   $I_{F(RMS)} = 2.1 \times 1.5A = 3.15A$, from Figure A1.7, allowing for 100% high tolerance on C

7. Transformer specifications are:

   $$V_{SEC(RMS)} = \frac{16.3 + 1.4}{\sqrt{2}} = 12.6 \text{ for FWB}$$

   (24 VCT for FWCT)

   $R_S = 0.48\Omega$ including reflected primary resistance, but subtract 2 x diode resistance

   $I_{AVG} = I_{IN(DC)} = 3A$

   $$I_{SEC(RMS)} = \frac{I_{IN(DC)} \times F}{\sqrt{2}} = \frac{3A \times 2.1}{1.414} = 4.45A$$

   VA rating = 4.45A x 12.6 = 56VA, or 62VA, allowing for 10% high line.

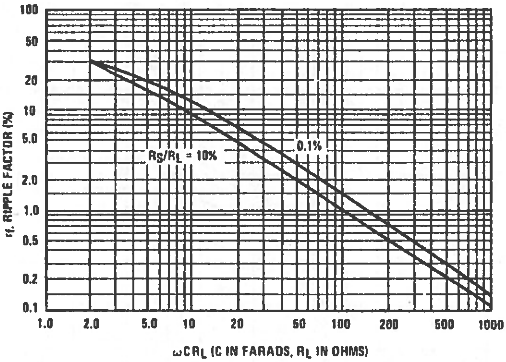
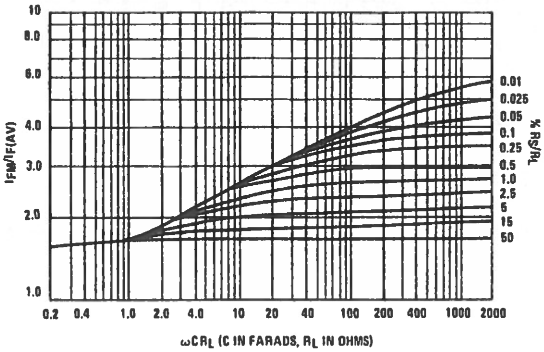


FIGURE A1.13 Ripple as a Function of Filter Constants for Full-Wave Voltage Doubler



RMS Rectifier Current as a Function of Filter Constants for Full-Wave Voltage Doubler

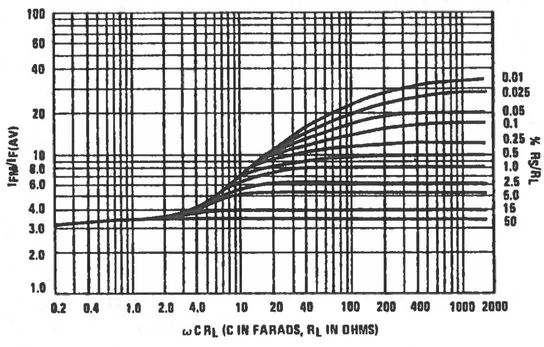


FIGURE A1.14 Relation of RMS to Peak and Average Diode Currents

A2.0 DECIBEL CONVERSION

A2.1 Definitions

The decibel (dB) is the unit for comparing relative levels of sound waves or of voltage or power signals in amplifiers.

The number of dB by which two power outputs P_1 and P_2 (in Watts) may differ is expressed by:

$$10 \log \frac{P_1}{P_2}$$

or, in terms of volts:

$$20 \log \frac{E_1}{E_2} \qquad \text{(Figure A2.1)}$$

or, in current:

$$20 \log \frac{I_1}{I_2}$$

While power ratios are independent of source and load impedance values, voltage and current ratios in these formulas hold true only when the source and load impedances Z_1 and Z_2 are equal. In circuits where these impedances differ, voltage and current ratios are expressed by:

$$\text{dB} = 20 \log \frac{E_1 \sqrt{Z_2}}{E_2 \sqrt{Z_1}} \quad \text{or} \quad 20 \log \frac{I_1 \sqrt{Z_1}}{I_2 \sqrt{Z_2}}$$

Specific reference levels, i.e., the 0 dB point, are denoted by a suffix letter following the abbreviation dB. Common suffixes and their definitions follow:

dBm – referenced to 1 mW of power

dBV – referenced to 1 V

dBW – referenced to 1 W

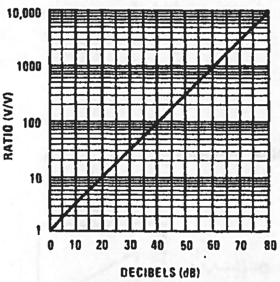


FIGURE A2.1 Gain Ratio to Decibel Conversion Graph (Note: For negative values of decibels, i.e., gain attenuation, simply invert the ratio number. For example, –20 dB = 1/10 V/V.)

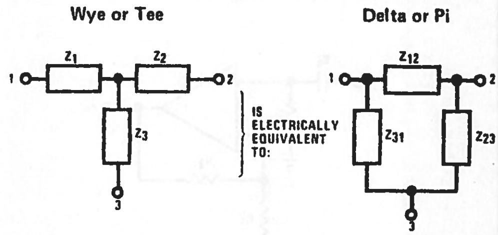
A2.2 Relationship Between dB/Octave and dB/Decade

dB/Octave	dB/Decade
3	10
6	20
9	30
10	33.3
12	40
15	50
18	60

A3.0 WYE-DELTA TRANSFORMATION

Wye-delta transformation techniques (and the converse, delta-wye) are very powerful analytical tools for use in understanding feedback networks. Known also as tee-pi and pi-tee transformations, their equivalencies are given below.

A3.1 Wye-Delta (Tee-Pi)



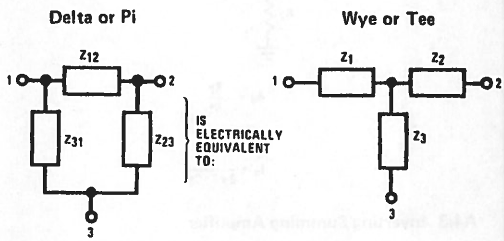
where:

$$Z_{12} = Z_1 + Z_2 + \frac{Z_1 Z_2}{Z_3} \qquad \text{(A3.1.1)}$$

$$Z_{23} = Z_2 + Z_3 + \frac{Z_2 Z_3}{Z_1} \qquad \text{(A3.1.2)}$$

$$Z_{31} = Z_3 + Z_1 + \frac{Z_3 Z_1}{Z_2} \qquad \text{(A3.1.3)}$$

A3.2 Delta-Wye (Pi-Tee)



where:

$$Z_1 = \frac{Z_{12} Z_{31}}{Z_{12} + Z_{23} + Z_{31}} \qquad \text{(A3.2.1)}$$

$$Z_2 = \frac{Z_{12} Z_{23}}{Z_{12} + Z_{23} + Z_{31}} \qquad \text{(A3.2.2)}$$

$$Z_3 = \frac{Z_{31} Z_{23}}{Z_{12} + Z_{23} + Z_{31}} \qquad \text{(A3.2.3)}$$

A4.0 STANDARD BUILDING BLOCK CIRCUITS

Definitions:

A_V = Closed Loop AC Gain

f_o = Low Frequency –3dB Corner

R_{in} = Input Impedance

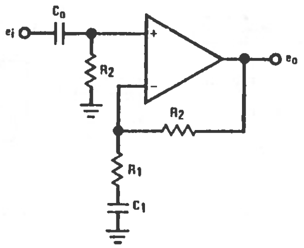
General Comments:

Power supply connections omitted for clarity.

Split supplies assumed.

Single supply biasing per A4.9 or A4.10.

A4.1 Non-Inverting AC Amplifier

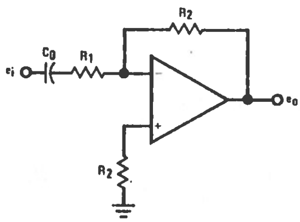


$$A_V = 1 + \frac{R_2}{R_1}$$

$$R_{in} = R_2$$

$$f_o = \frac{1.56}{2\pi T} \text{ where } T = R_2C_o = R_1C_1$$

A4.2 Inverting AC Amplifier

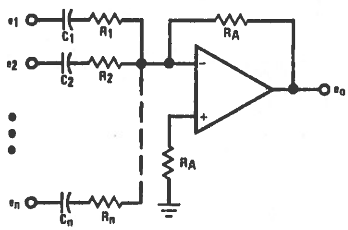


$$A_V = -\frac{R_2}{R_1}$$

$$R_{in} = R_1$$

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

A4.3 Inverting Summing Amplifier

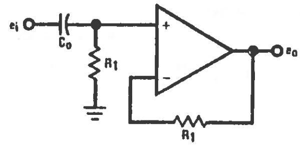


$$e_0 = -R_A\left(\frac{e_1}{R_1} + \frac{e_2}{R_2} + \ldots + \frac{e_n}{R_N}\right)$$

IF $R_1 = R_2 = \ldots = R_N$ THEN

$$e_0 = -\frac{R_A}{R_1}(e_1 + e_2 + \ldots + e_n)$$

A4.4 Non-Inverting Buffer

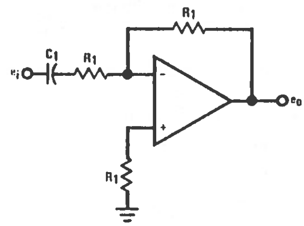


$$A_V = 1$$

$$R_{in} = R_1$$

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

A4.5 Inverting Buffer

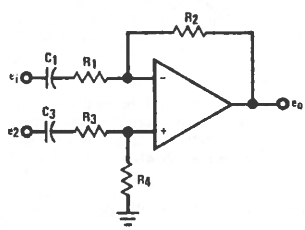


$$A_V = -1$$

$$R_{in} = R_1$$

$$f_o = \frac{1}{2\pi R_1 C_1}$$

A4.6 Difference Amplifier



$$e_o = \left(\frac{R_1 + R_2}{R_3 + R_4}\right)\frac{R_4}{R_1} e_2 - \frac{R_2}{R_1} e_1$$

IF $R_1 = R_3$* AND $R_2 = R_4$* THEN

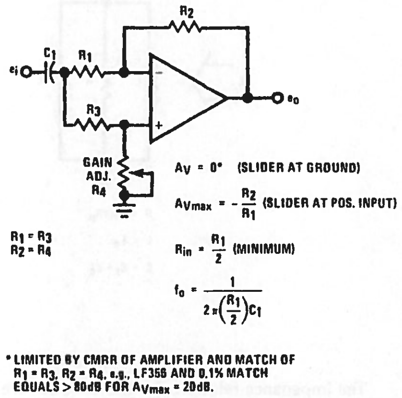
$$e_o = \frac{R_2}{R_1}(e_2 - e_1)$$

$$f_o = \frac{1.56}{2\pi T} \text{ where } T = R_1C_1 = (R_3 + R_4)C_3$$

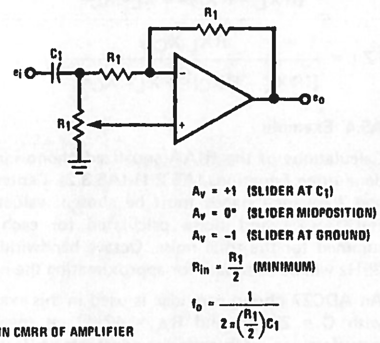
$R_2 = R_4$ FOR MINIMAL OFFSET ERROR

* – 0.1% MATCH FOR MAX CMRR

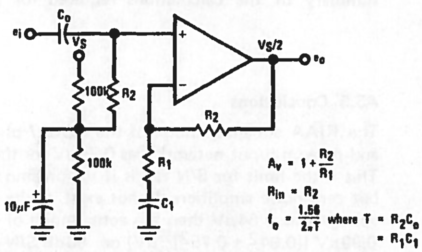
A4.7 Variable Gain AC Amplifier



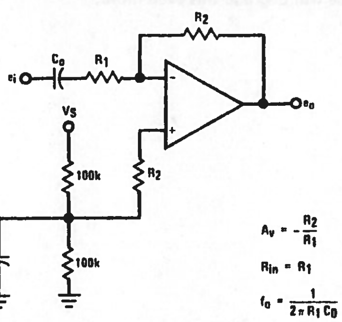
A4.8 Switch Hitter (Polarity Switcher, or 4-Quadrant Gain Control)



A4.9 Single Supply Biasing of Non-Inverting AC Amplifier



A4.10 Single Supply Biasing of Inverting AC Amplifier



A5.0 MAGNETIC PHONO CARTRIDGE NOISE ANALYSIS

A5.1 Introduction

Present methods of measuring signal-to-noise (S/N) ratios do not represent the true noise performance of phono preamps under real operating conditions. Noise measurements with the input shorted are only a measure of the preamp noise voltage, ignoring the two other noise sources: the preamp current noise and the noise of the phono cartridge.

Modern phono preamps have typical S/N ratios in the 70dB range (below 2mV @ 1kHz), which corresponds to an input noise voltage of 0.64μV, which looks impressive but is quite meaningless. The noise of the cartridge[1] and input network is typically *greater* than the preamp noise voltage, ultimately limiting S/N ratios. This must be considered when specifying preamplifier noise performance. A method of analyzing the noise of complex networks will be presented and then used in an example problem.

A5.2 Review of Noise Basics

The noise of a passive network is thermal, generated by the real part of the complex impedance, as given by Nyquist's Relation:

$$\overline{V_n^2} = 4\,k\,T\,\mathrm{Re}(Z)\,\Delta f \qquad \text{(A5.2.1)}$$

where: $\overline{V_n^2}$ = mean square noise voltage

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

T = absolute temperature (°K)

Re(Z) = real part of complex impedance (Ω)

Δf = noise bandwidth (Hz)

The total noise voltage over a frequency band can be readily calculated if it is white noise (i.e., Re(Z) is frequency independent). This is not the case with phono cartridges or most real world noise problems. Rapidly changing cartridge network impedance and the RIAA equalization of the preamplifier combine to complicate the issue. The total input noise in a non-ideal case can be calculated by breaking the noise spectrum into several small bands where the noise is nearly white and calculating the noise of each band. The total input noise is the RMS sum of the noise in each of the bands $N_1, N_2, \ldots, N_n$.

$$V_{noise} = (V_{N1}^2 + V_{N2}^2 + \ldots + V_{Nn}^2)^{1/2} \qquad \text{(A5.2.2)}$$

This expression does not take into account gain variations of the preamp, which will also change the character of the noise at the preamp output. By reflecting the RIAA equalization to the preamp input and normalizing the gain to 0dB at 1kHz, the equalized cartridge noise may then be calculated.

$$V_{EQ} = (|A_1|^2 V_{N1}^2 + |A_2|^2 + \ldots + |A_n|^2 V_{Nn}^2)^{1/2} \qquad \text{(A5.2.3)}$$

where: V_{EQ} = equalized preamp input noise

$|A_n|$ = magnitude of the equalized gain at the center of each noise band (V/V)

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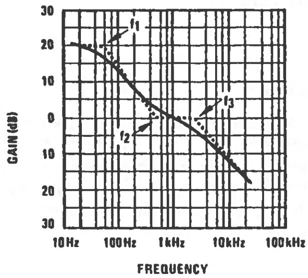
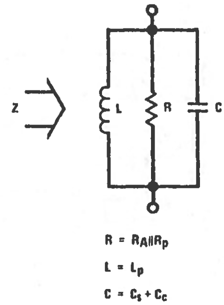


FIGURE A5.1 Normalized RIAA Gain



A5.3 Cartridge Impedance

The simplified lumped model of a phono cartridge consists of a series inductance and resistance shunted by a small capacitor. Each cartridge has a recommended load consisting of a specified shunt resistance and capacitor. A model for the cartridge and preamp input network is shown in Figure A5.2.

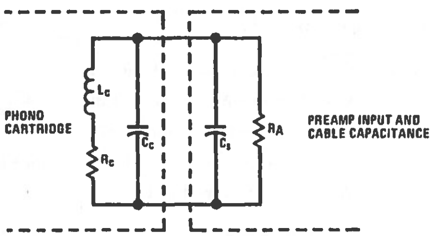
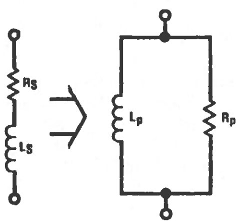


FIGURE A5.2 Phono Cartridge and Preamp Input Network

This seemingly simple circuit is quite formidable to analyze and needs further simplification. Through the use of Q equations,[2] a series L-R is transformed to a parallel L-R.



$$Q = \frac{\omega L_S}{R_S}$$

$$R_p = R_S(1 + Q^2) \qquad \text{(A5.3.1)}$$

$$L_p = L_S\left(\frac{1 + Q^2}{Q^2}\right)$$

Simplifying the input network,

The impedance relations for this network are:

$$\text{Re}(Z) = \frac{R\,X_L{}^2\,X_C{}^2}{(R\,X_L - R\,X_C)^2 + X_L{}^2\,X_C{}^2}$$

$$|Z| = \frac{R\,X_L\,X_C}{[(R\,X_L - R\,X_C)^2 + X_L{}^2\,X_C{}^2]^{1/2}} \qquad \text{(A5.3.2)}$$

A5.4 Example

Calculations of the RIAA equalized phono input noise are done using Equations (A5.2.1)-(A5.3.2). Center frequencies and frequency bands must be chosen: values of Rp, Lp, Re(Z), |Z| and noise calculated for each band, then summed for the total noise. Octave bandwidths starting at 25 Hz will be adequate for approximating the noise.

An ADC27 phono cartridge is used in this example, loaded with C = 250pF and R_A = 47kΩ, as specified by the manufacturer, with cartridge constants of Rs = 1.13kΩ and Ls = 0.75H. (C_C may be neglected.) Table A5.1 shows a summary of the calculations required for this example.

A5.5 Conclusions

The RIAA equalized noise of the ADC27 phono cartridge and preamp input network was 0.75μV for the audio band. This is the limit for S/N ratios if the preamp was noiseless, but zero noise amplifiers do not exist. If the preamp noise voltage was 0.64μV then the actual noise of the system is 0.99μV ($[0.64^2 + 0.75^2]^{1/2}$μV) or 66dB S/N ratio (re 2mV @ 1kHz input). This is a 4dB loss and the preamp current noise will degrade this even more.

TABLE A5.1 Summary of Calculations

f Range (Hz)	25 - 50	50 - 100	100 - 200	200 - 400	400 - 800	800 - 1.6k	1.6k - 3.2k	3.2k - 6.4k	6.4k - 12.8k	12.8k - 20k
f Center (Hz)	37.5	75	150	300	600	1200	2400	4800	9600	16.4k
f_{BW} (Hz)	25	50	100	200	400	800	1600	3200	6400	7.2k
$Q = \frac{\omega L_s}{R_s}$	0.156	0.313	0.625	1.25	2.5	5	10	20	40	68.4
Q^2	0.0244	0.098	0.391	1.56	6.25	25	100	400	1600	4678.6
$1 + Q^2$	1.0244	1.098	1.391	2.56	7.25	26	101	401	1601	4679.6
$\frac{1+Q^2}{Q^2}$	42	11.24	3.56	1.64	1.16	1.04	1.01	1.0	1.0	1.0
R_p (Ω)	1.16k	1.24k	1.57k	2.9k	8.2k	29.4k	114k	454k	1.8M	5.29M
L_p (H)	31.5	8.43	2.67	1.23	0.87	0.78	0.76	0.75	0.75	0.75
$R_p \| R_A$ (Ω)	1.13k	1.21k	1.52k	2.74k	7k	18.1k	32.9k	42.6k	45.8k	46.6k
X_L (Ω)	7.42k	3.97k	2.52k	2.32k	3.28k	5.88k	11.45k	22.6k	45.2k	77.2k
X_c (Ω)	17M	8.48M	4.24M	2.12M	1.06M	0.53M	0.265M	0.133M	66.3k	38.8k
$R_e(Z)$ (Ω)	1.11k	1.11k	1.11k	1.15k	1.26k	1.73k	3.86k	12.4k	41.5k	34k
$\|Z\|$ (Ω)	1.12k	1.15k	1.3k	1.77k	2.97k	5.59k	11.7k	24.4k	43.6k	40.1k
e_{nz} ($nV/\sqrt{Hz}$)	4.1	4.1	4.1	4.1	4.3	5.1	7.3	14	26	23
V_N (nV)	20.5	29	41	58	86	144.2	292	792	2080	1952
V_n^2 (nV^2)	420.3	840.5	1681	3362	7396	20.8k	85.3k	627.7k	4.33M	3.81M
A^2	63.04	31.6	10	3.17	1.59	0.89	0.45	0.159	0.05	0.025
$A^2 V_n^2$ (nV^2)	26.5k	26.6k	16.8k	10.7k	11.8k	18.5k	38.1k	99.7k	216.3k	95.2k

$(\Sigma V_n^2)^{1/2} = 2.98\mu V$ unequalized noise.

$(\Sigma |A_n|^2 V_n^2)^{1/2} = 0.75\mu V$ RIAA equalized noise.

Thus it is apparent that present phono preamp S/N ratio measurement methods are inadequate for defining actual system performance, and that a new method should be used – one that more accurately reflects true performance.

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2. Fristoe, H. T., "The Use of Q Equations to Solve Complex Electrical Networks," *Engineering Research Bulletin*, Oklahoma State University, 1964.
3. Korn, G. A. and Korn, T. M., *Basic Tables in Electrical Engineering*, McGraw-Hill, New York, 1965.
4. Maxwell, J., *The Low Noise JFET – The Noise Problem Solver*, Application Note AN-151, National Semiconductor, 1975.

A6.0 GENERAL PURPOSE OP AMPS USEFUL FOR AUDIO

National Semiconductor's line of integrated circuits designed specifically for audio applications consists of 4 dual preamplifiers, 3 dual power amplifiers, and 6 mono power amplifiers. All devices are discussed in detail through most of this handbook; there are, however, other devices also useful for general purpose audio design, a few of which appear in Table A6.1. Functionally, most of these parts find their usefulness between the preamplifier and power amplifier, where line level signal processing may be required. The actual selection of any one part will be dictated by its actual function.

TABLE A6.1 General Purpose Op Amps Useful for Audio

Device[1]	Single	Dual	Quad	Compensated	Decompensated[2]	Uncompensated	Slew Rate[3] (V/μs)	Supply Voltage Typical (Min. → Max.)	Supply Current Maximum (mA)	General Features of Audio Application Interest
LM301A	X					X	5[4]	±3 → ±18	3	Low THD.
LM310	X			X			30	±5 → ±18	5.5	Fast unity-gain buffer.
LM318	X			X			50	±5 → ±18	10	High slew rate.
LM324			X	X			0.3	3 → 30 (±1.5 → ±15)	2	Low supply current quad.
LM343	X			X			2.5	±4 → ±34	5	High supply voltage.
LM344	X					X	30	±4 → ±34	5	Fast LM343.
LM348			X	X			0.5	±5 → ±18	4.5	Quad LM741.
LM349			X		X		2	±5 → ±18	4.5	Fast LM348.
LF355	X			X			5	±5 → ±18	4	Low supply current LF356.
LF356[5]	X			X			12	±5 → ±18	10	Fast, JFET input, low noise.
LF357	X				X		50	±5 → ±18	10	Higher slew rate LF356.
LM358		X		X			0.3	3 → 30 (±1.5 → ±15)	1.2	Dual LM324.
LM394	–	–	–	–	–	–	–	–	–	Supermatch low noise transistor pair.
LM741	X			X			0.5	±3 → ±18	2.8	Workhorse of the industry.
LM747		X		X			0.5	±3 → ±18	5.6	Dual LM741 (14 pin).
LM1458		X		X			0.2	±3 → ±18	5.6	Dual LM741 (8 pin).
LM3900			X	X			0.5	4 → 30 (±2 → ±15)	10	Quad current differencing amp.
LM4250	X			X			0.03	±1 → ±18	0.1	Micropower.

1. Commercial devices shown (0°C-70°C); extended temperature ranges available.
2. Decompensated devices stable above a minimum gain of 5V/V.
3. A_V = 1V/V unless otherwise specified.
4. Compensation capacitor = 3pF; A_V = 10V/V minimum.
5. Highly recommended as general purpose audio building block.

A7.0 FEEDBACK RESISTORS AND AMPLIFIER NOISE

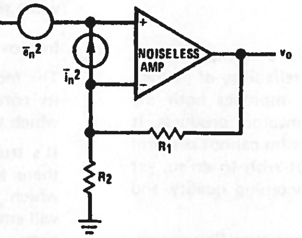

FIGURE A7.1 Practical Feedback Amplifier

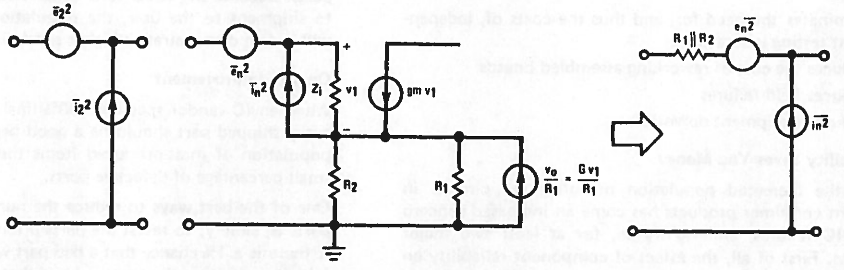

FIGURE A7.2 Model of First Stage of Amplifier

To see the effect of the feedback resistors on amplifier noise, model the amplifier of Figure A7.1 as shown in Figure A7.2, and neglect thermal noise.

We must now show that the intrinsic noise generators $\overline{e_n^2}$ and $\overline{i_n^2}$ are related to the noise generators outside the feedback loop, $\overline{e_2^2}$ and $\overline{i_2^2}$. In addition, the output noise at v_O can be related to v_1 by the open loop gain of the amplifier G, i.e.,

$$v_O = v_1 G$$

Thus v_1 is a direct measure of the noise behavior of the amplifier. Open circuit the amplifier and equate the effects of the two noise current generators. By superposition:

$$v_1 = i_2 Z_i$$

also $v_1 = i_n Z_i$

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

Short circuit the input of the amplifier to determine the effect of the noise voltage generators. To do this, short the amplifier at $\overline{e_2^2}$ and determine the value of v_1, then short circuit the input at $\overline{e_n^2}$ and find the value of v_1.

$$e_2 = v_1 + R_1\|R_2\left(g_m v_1 + \frac{G v_1}{R_1}\right)$$

$$v_1 = e_2 \frac{1}{1 + g_m R_1\|R_2 + G\dfrac{R_1\|R_2}{R_1}}, \qquad \text{(A7.1)}$$

Now short the input at $\overline{e_n^2}$; $\overline{e_n^2}$ and $\overline{i_n^2}$ both affect v_1.

$\overline{e_n^2}$ gives:

$$v_1 = e_n \frac{1}{1 + g_m R_1\|R_2 + G\dfrac{R_1\|R_2}{R_1}} \qquad \text{(A7.2)}$$

$\overline{i_n^2}$ gives:

$$-v_1 = Z_i\|R_1\|R_2\left(g_m v_1 + G\frac{v_1}{R_1} - i_n\right)$$

Assume $Z_i \gg R_1\|R_2$

$$v_1 = \frac{i_n R_1\|R_2}{1 + g_m R_1\|R_2 + G\dfrac{R_1\|R_2}{R_1}} \qquad \text{(A7.3)}$$

Add Equations (A7.2) and (A7.3) and equate to Equation (A7.1):

$$\frac{\overline{e_n^2} + \overline{i_n^2}(R_1\|R_2)^2}{\left(1 + g_m R_1\|R_2 + G\dfrac{R_1\|R_2}{R_1}\right)^2} = \frac{\overline{e_2^2}}{\left(1 + g_m R_1\|R_2 + G\dfrac{R_1\|R_2}{R_1}\right)^2}$$

$$\therefore \overline{e_2^2} = \overline{e_n^2} + \overline{i_n^2}(R_1\|R_2)^2.$$

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A8.0 RELIABILITY

A8.1 Consumer plus program

National's Consumer Plus Program is a comprehensive program that assures high quality and reliability of molded integrated circuits. The C+ Program improves both the quality and reliability of National's consumer products. It is intended for the manufacturing user who cannot perform 100% inspection of his ICs, or does not wish to do so, yet needs significantly-better-than-usual incoming quality and reliability levels for his ICs.

Integrated circuit users who specify Consumer Plus processed parts will find that the program:

- eliminates 100% the need for incoming electrical inspection
- eliminates the need for, and thus the costs of, independent testing laboratories
- reduces the cost of reworking assembled boards
- reduces field failures
- reduces equipment downtime

Reliability Saves You Money

With the increased population of integrated circuits in modern consumer products has come an increased concern with IC failures, and rightly so, for at least two major reasons. First of all, the effect of component reliability on system reliability can be quite dramatic. For example, suppose that you, as a color TV manufacturer, were to choose ICs that are 99% reliable. You would find that if your TV system used only seven such ICs, the overall reliability of IC portion would be only 50% for one out of each ten sets produced. In other words, only nine out of your ten systems would operate. The result? Very costly to produce and probably very difficult to sell. Secondly, whether the system is large or small, you cannot afford to be hounded by the spectre of unnecessary maintenance costs, not only because labor, repair or rework costs have risen -- and promise to continue to rise -- but also because field replacement may be prohibitively expensive.

Reliability vis-a-vis Quality

The words "reliability" and "quality" are often used interchangeably as though they connoted identical facets of a product's merit. But reliability and quality are different and IC users must understand the difference to evaluate various vendors' programs for product improvement that are generally available, and National's Consumer Plus Program in particular.

The concept of *quality* gives us information about the population of faulty IC devices among good devices, and generally relates to the number of faulty devices that arrive at a user's plant. But looked at in another way, quality can instead relate to the number of faulty ICs that escape detection at the IC vendor's plant.

It is the function of a vendor's Quality Control arm to monitor the degree of success of that vendor in reducing the number of faulty ICs that escape detection. QC does this by testing the outgoing parts on a sampled basis. The Acceptable Quality Level (AQL) in turn determines the stringency of the sampling. As the AQL decreases it becomes more difficult for bad parts to escape detection; thus the quality of the shipped parts increases.

The concept of *reliability*, on the other hand, refers to how well a part that is initially good will withstand its environment. Reliability is measured by the percentage of parts that fail in a given period of time.

Thus ICs of high quality may, in fact, be of low reliability, while those of low quality may be of high reliability.

Improving the Reliability of Shipped Parts

The most important factor that affects a part's reliability is its construction: the materials used and the method by which they are assembled.

It's true that reliability cannot be tested into a part, but there are tests and procedures that can be implemented which subject the IC to stresses in excess of those that it will endure in actual use. These will eliminate most marginal parts.

In any test for reliability the weaker parts will normally fail first. Stress tests will accelerate the failure of the weak parts. Because the stress tests cause weak parts to fail prior to shipment to the user, the population of shipped parts will in fact demonstrate a higher reliability in use.

Quality Improvement

When an IC vendor specifies 100% final testing of his parts, every shipped part should be a good part. However, in any population of mass-produced items there does exist some small percentage of defective parts.

One of the best ways to reduce the number of such faulty parts is, simply, to retest the parts prior to shipment. Thus, if there is a 1% chance that a bad part will escape detection initially, retesting the parts reduces that probability to only 0.01%. (A comparable tightening of the QC group's sampled-test plan ensures this.)

National's Consumer Plus Program Gets It All Together

We've stated that the C+ Program improves both the quality and reliability of National's molded integrated circuits, and pointed out the difference between these two concepts. Now, how do we bring them together? The answer is in the C+ Program processing, which is a continuum of stress and double testing. With the exception of the final QC inspection, which is sampled, all steps of the C+ processes are performed on 100% of the program parts. The following flow chart shows how we do it.

Epoxy B Processing for All Molded Parts

At National, all molded semiconductors, including ICs, have been built by this process for some time now. All processing steps, inspections and QC monitoring are designed to provide highly reliable products. (A reliability report is available that gives, in detail, the background of Epoxy B, the reason for its selection at National and reliability data that proves its success.)

Six Hour, 150°C Bake

This stress places the die bond and all wire bonds into a tensile and shear stress mode, and helps eliminate marginal bonds and connections.

Five Temperature Cycles (0°C to 100°C)

Exercising the circuits over a 100°C temperature range generally eliminates any marginal bonds missed during the bake.

High Temperature (100°C) Functional Electrical Test

A high-temperature test such as this with voltages applied places the die under the most severe stress possible. The test is actually performed at 100°C – 30°C higher than the commercial ambient limit. All devices are thoroughly exercised at the 100°C ambient. (Even though Epoxy B has virtually eliminated thermal intermittents, we perform this test to insure against even the remote possibility of such a problem. Remember, the emphasis in the C+ Program is on the elimination of those marginally performing devices that would otherwise lower field reliability of the parts.)

DC Functional and Parametric Tests

These room-temperature functional and parametric tests are the normal, final tests through which all National products pass.

Tighter-Than-Normal QC Inspection Plans

Most vendors sample inspect outgoing parts to a 0.65% AQL. Some use even a looser 1% AQL. However, not only do we sample your parts to a 0.28% AQL for all data-sheet DC parameters, but they receive a 0.14% AQL for functionality as well. Functional failures – not parameter shifts beyond spec – cause most system failures. Thus, the five-times to seven-times tightening of the sampling procedure (from 0.65%-1% to 0.14% AQL) gives a substantially higher quality to your C+ parts. And you can rely on the integrity of your received ICs without incoming tests.

Ship Parts

Here are the QC sampling plans used in our Consumer Plus test program:

Test	Temperature	AQL
Electrical Functionality	25°C	0.14%
Parametric, DC	25°C	0.28%
Parametric, DC	100°C	1%
Parametric, AC	25°C	1%
Major Mechanical	–	0.25%
Minor Mechanical	–	1%

A8.2 Operating Junction Temperatures

For steady state operation within the operating junction temperature range of the part, most failure modes are due to die surface related effects such as zener voltage drift due to field effect changes caused by movement of ions in the oxide. After extensive life testing, National Semiconductor has developed some average "acceleration factors" relating increased surface related failure rates to increased junction temperature. For example: an IC device operating steady state at T_J = 125°C for 500 hours will experience approximately the same failure rate as if operated at T_J = 70°C for 72,500 hours. The acceleration factor from 70°C to 125°C (T_J) would be 145. From 125°C to 150°C (T_J) the acceleration factor is 6.3. This indicates the greatly increased part lifetime the user can realize by maintaining the part at a low operating junction temperature.

A9.0 AUDIO RADIO GLOSSARY

"A" Line Filter

The L-C filter used in the power supply lead of an automotive radio to suppress transient voltage spikes. A series inductance from 0.2mHz to 2mHz with a 1000µF to 2000µF capacitor to ground is typical.

AB-Bias

A technique used in class B audio amplifiers to prevent crossover distortion. The complementary output devices of the amplifier are biased "on" so that a small amount of current runs in them in the absence of a signal allowing a smooth transition from a positive signal swing to a negative signal swing and vice-versa. This AB-Bias current is typically from 1 – 30mA and is the major component of the amplifier quiescent current and stand-by power dissipation.

Adjacent Channel Rejection

A measure of AM receiver selectivity – the ratio of the detected signal level of a desired r.f. carrier to the detected signal level of an undesired carrier of similar strength located ±10kHz from the desired carrier. Usually > 20dB. (see also selectivity)

Ambience

The indirect sounds heard in a concert hall or other large listening area that contribute to the overall auditory effect obtained when listening to live performances.

Amplifier

Class A

A class A transistor audio amplifier refers to an amplifier with a single output device that has a collector flowing for the full 360° of the input cycle.

Class B

The most common type of audio amplifier that basically consists of two output devices each of which conducts for 180° of the input cycle (see AB-Bias however).

Class C

In a class C amplifier the collector current flows for less than 180°. Although highly efficient, high distortion results and the load is frequently tuned to minimize this distortion (primarily used in R.F. power amplifiers).

Class D

A switching or sampling amplifier with extremely high efficiency (approaching 100%). The output devices are used as switches, voltage appearing across them only while they are off, and current flowing only when they are saturated.

AM Rejection
(AM Suppression)

The ratio of the recovered audio output produced by a desired FM signal of specified level and deviation to the recovered audio output produced by an unwanted AM signal of specified amplitude and modulation index. Usually the AMR of a system is measured over a range of input signal levels with 100% FM and 30% AM, 1kHz modulating frequency.

High quality tuner receiver:	AMR > 50dB
Mid quality/multi-band/TV sound:	AMR > 40dB

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Anechoic Chamber

A derived term for a room or enclosure that is designed to be echo-free over a specified frequency range. Any sound reflections within this frequency range must be less than 10% of the source sound pressure.

A.F.C.

Automatic Frequency Control — controlling the frequency of the local oscillator of a superheterodyne receiver at the value required to produce the desired intermediate frequency. An AFC system will correct for mistuning and oscillator frequency drifts caused by temperature/supply voltage variations and ageing. In higher quality receivers the AFC circuit output can be used to drive a display meter to facilitate tuning.

A.G.C.

Automatic Gain Control — an AGC system operates to maintain the output of an amplifier approximately constant despite input signal level variations, by changing the amplifier gain as the input signal changes. This allows tuning a radio from strong to weak signals without resetting the manual volume control.

AGC Figure of Merit — the widest possible range of input signal level required to make the output signal drop by a specified amount from the specified maximum output level. Typical F.O.M. numbers are from 40dB to 50dB, for domestic radios and about 60dB for automotive radios (for −10dB output level change).

A.L.C.

Automatic Level Control — a compressor circuit usually located at the microphone input of a tape recorder that operates to keep the recorded sound level within predetermined limits regardless of input sound level changes. A figure of merit can be measured similar to that for an A.G.C. system.

Average Power

The signal produced by amplifier into a given load with a given input signal. For sine-wave inputs the average power (also termed continuous or RMS power) is a measure of the amplifier capability to deliver peak outputs while delivering significant power at all levels below the peak. For a peak-to-peak sine-wave output signal Eo into a load R_L the power is given by

$$P_O = \frac{Eo^2}{8R_L}$$

Azimuth

The angle of a tape head's pole-piece slot relative to the direction of tape travel. Misalignment (Azimuth error) will cause a loss of high frequencies. For a track of width W and a recorded wavelength λ, and angle of misalignment θ(Azimuth error) will give a level loss of

$$20\mathrm{LOG}_{10}\left\{\frac{\mathrm{Sin}\,\dfrac{\pi W\theta}{\lambda}}{\dfrac{\pi W\theta}{\lambda}}\right\}\mathrm{dB}$$

For example, at 5kHz a −3dB loss will be incurred at a tape speed of 1 7/8 I.P.S. by an error of 0.48 degrees.

Bandwidth

AM

The width of the band of frequencies over which the detector output amplitude does not drop to less than one half the center tuned response with a constant input signal strength. Because of the effect of the normal AM radio AGC circuit, the bandwidth is measured both before and after the onset of AGC action. Below the AGC threshold typical bandwidths are from 4kHz to 10kHz. Above the AGC threshold (signal input +40dB above level required for rated detector output) the I_F bandwidth is from 25kHz to 80kHz.

FM

The range of frequencies at the detector limited by the −3dB amplitude points. The measurement is made with an input signal that produces a detector output −3dB below the detector level obtained with a large r.f. input signal. For monophonic receivers the typical BW is 180kHz, and for stereo 225kHz.

Noise

Noise bandwidth is a term used in the design of phase locked loops (PLL) to describe the response of the loop to signals on either side of the desired locking frequency. It is not measured directly but is the equivalent bandwidth of the PLL derived by plotting the square of the loop amplitude/frequency response and deriving a rectangular pass band characteristic having the same peak value and enclosing the same area. For a loop filter with a single RC roll-off, the noise bandwidth of the filter is 1.57 f-3dB.

Power

The power bandwidth of an audio amplifier is the frequency range over which the amplifier voltage gain does not fall below 0.707 of the flat band voltage gain specified for a given load and output power.

Power bandwidth also can be measured by the frequencies at which a specified level of distortion is obtained while the amplifer delivers a power output 6dB below the rated output. For example, an amplifier rated at 60 watts with $\leqslant$ 0.25% THD, would make its power bandwidth measured as the difference between the upper and lower frequencies at which 0.25% distortion was obtained while the amplifier was delivering 30 watts.

Biamplification

The technique of splitting the audio frequency spectrum into two sections and using individual power amplifiers to drive a separate woofer and tweeter. Cross-over frequencies for the amplifiers usually vary between 500Hz and 1600Hz. "Biamping" has the advantages of allowing smaller power amps to produce a given sound pressure level and reducing distortion effects produced by overdrive in one part of the frequency spectrum affecting the other part.

Bias (Tape)

The magnetic coating on audio tapes exhibits non-linear regions in the magnetization characteristic at zero magnetization and at saturation levels. If a steady state magnetic field is applied to the tape during the recording process, the signal or audio information is restricted to the linear portion of the magnetization characteristic. This is called "Biasing" (analogous to the dc bias for solid state devices used to ensure operation in the linear region).

DC Bias

The simplest method of biasing a tape is to apply a steady state dc current to the recording head so that the tape is magnetized to a linear part of the characteristic.

AC Bias

An ultrasonic (50kHz — 110kHz) alternating current applied to the recording head so that ideal or "anhysteric" magnetization of the tape takes place. In the presence of the bias waveform, the signal (audio) magnetization characteristic is linearized enabling larger flux levels to be recorded, and improving the S/N ratio compared to dc bias.

Peak Bias

AC biasing also increases the tape sensitivity (larger recorded flux for given recording current). However, beyond a certain bias level, called the peak bias, the recorded flux level starts to decrease and distortion levels begin to increase. The required peak bias depends on the tape formulation used and is typically lower for ferric oxide tape than for chromium dioxide tapes.

Capstan

A motor driven spindle that feeds tape at a constant speed past the tape heads. The tape is held against the capstan by an idler or pinch wheel.

Capture Range

The capture range of a PLL is the frequency range, centered about the V.C.O. free running frequency, over which the loop can acquire lock with the input signal.

Capture Ratio

A measure of the ability of an FM tuner to select the stronger of two r.f. signals at or near the same frequency. It is the ratio of the signal strength of the carriers required for 30dB suppression of the audio from the weaker signal at the tuner output.

The rated capture ratio is measured by increasing the signal strength of an unmodulated carrier until there is a drop in the tuner audio output being obtained from a 100% modulated carrier at a 1mV signal level. The ratio in dB of the unmodulated carrier levels required to cause 1dB and 30dB drops in the audio output, divided by two, is the capture ratio.

Cartridge

A phonograph pick-up and stylus combination

Constant Amplitude Pick-Ups:

Known as ceramic or crystal cartridges, piezo electric pick-ups depend on the piezo-electric effect — i.e. when crystals (rochelle salt) or ceramics (barium titanate) are mechanically flexed, an EMF is developed directly proportional to the degree of flexure. Very popular in low fidelity sound systems, these cartridges have very high output levels from 100mV to 2V.

Constant Velocity Pick-ups:

Moving coil or moving magnet cartridges develop an output proportional to the velocity of the stylus motion and are used in high fidelity systems. Substantially lower output levels around 3mV to 5mV are obtained compared to crystal cartridges.

Channel Separation

The degree to which the signal in one amplifier is kept separate from an adjacent undriven amplifer. Channel separation for FM stereo decoders is typically >40dB whereas phono cartridge channel separation is typically between 20dB and 30dB.

C.C.I.R./A.R.M.

Literally: International Radio Consultative Committee/Average Responding Meter

This refers to a weighted noise measurement for a Dolby B type noise reduction system. A filter characteristic is used that gives a closer correlation of the measurement with the subjective annoyance of noise to the ear. Measurements made with this filter cannot necessarily be related to unweighted noise measurements by some fixed conversion factor since the answers obtained will depend on the spectrum of the noise source.

Coercivity

A measure of the magnetic field strength required to erase a tape to a state of zero magnetism. High coercivity tapes are harder to erase but suffer less from high frequency losses caused by self-erasure (see self-erasure).

Compandor

A complementary compression and expansion system used for audio noise reduction. Before recording or being transmitted the entire signal is compressed according to some fixed law and afterwards is expanded to its original dynamic range for replay.

Composite Signal

The stereo FM broadcast modulation signal consisting of a 19Hz pilot tone, (L + R) information and (L − R) information modulated on a suppressed 38kHz carrier and (if any) a 67kHz FM carrier with ±6kHz deviation S.C.A. channel.

Crest Factor

The ratio of the peak value of a waveform to its RMS value. For example, the crest factor for audio amplifier noise is from 3:1 to 5:1.

Crossmodulation

Crossmodulation is the name given to the phenomenon whereby information from an AM carrier is transferred to another carrier. For a reciever, non-linearities in the R.F. or mixer stages can cause the modulation from an adjacent undesired signal to modulate the desired carrier signal to which the receiver is tuned. For measurements typically 30% modulated carriers are used and the level of crossmodulation specified to be ⩽1%.

Crossover Distortion

Distortion caused in the output stage of a class B amplifier. It can result from inadequate bias current (see AB Bias) allowing a dead zone where the output does not respond to the input as the input cycle goes through its zero crossing point. Also for I/Cs an inadequate frequency response of the output PNP device can cause a turn-on delay giving crossover distortion for negative going transition through zero at the higher audio frequencies.

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Crossover Frequency

A frequency at which other frequencies above and below it are separated. A crossover network will separate the high and low frequencies in a tweeter/woofer speaker system for example, with a single crossover frequency between 1kHz and 3kHz.

dbx

An audio noise reduction system operating on the wide band companding principle. A true RMS detector controls the gain of an amplifier before recording with operator variable compression factor from 1.0 to 3.0. On playback the dynamic range is restored by a similar expansion process.

De-Emphasis

To reduce the effect of broadband noise in an FM broadcast the signal has pre-emphasis of the higher frequencies defined by a 75μS time constant (50μS in Europe). At the receiver, a 75μS de-emphasis restores the frequency/amplitude relationships while reducing the higher frequency noise added during the broadcast signal transmission.

Detector

The point in a receiver at which the modulating information is recovered from the carrier waveform.

Differential Peak Detector

An FM detector that operates by comparing the peak voltages detected on either side of a single-tuned circuit.

Quadrature Detector

Compares the relative phases of the I.F. signal on either side of a circuit tuned to give 90° phase shift at the intermediate frequency.

Synchronous Detector

A P.L.L. detector where the modulated signal is compared in a phase detector to a local oscillator signal.

Power Detector

An AM peak detector where the diode is the base-emitter junction of a transistor.

D.I.N.

Literally Deutsche Industrie Norm.

Designates a European performance standard or test procedure. It also describes a unitized audio connector plug and socket.

Deviation

The instantaneous frequency difference of an FM signal from the unmodulated carrier frequency.

Distortion

The effects produced by an electronic circuit when the signal output from the circuit does not exactly duplicate the input signal in all respects except magnitude (see intermodulation, THD, etc.)

Dolby B

Dolby B is a simplified version of the Dolby A professional quality noise reduction system. The amplitude of low level signals over a selected frequency range is increased prior to recording to enhance them above tape noise. On playback the original levels are restored causing a corresponding reduction in the audible tape noise. The major difference with Dolby A which used four frequency bands, is the use of a single variable frequency band with a cut-off frequency that increases in the presence of high level high frequency signals.

Dolby Level

Because of the complementary nature of the Dolby B noise reduction system, the audio channel between the encoder and the decoder must have a fixed gain such that the decoding signal level is within 2dB of the encoding signal level. Also if recordings are interchangeable the signals in the noise reduction system must be related to the levels in the audio channel. Dolby level provides this reference and corresponds to a specified tape flux density when recorded with a 400Hz tone. For reel to reel and eight track cartridge tapes this is 185nWb/m, and for cassettes Dolby level is 200nWb/m.

Equalization

The adjustment of the frequency or phase characteristics of a signal or audio device. Examples are the R.I.A.A. recording characteristic for phonograph discs and the N.A.B. recording playback characteristic for tape. Equalizers are audio equipment devices inserted into playback systems to compensate for signal variations, room acoustics and loudspeaker responses.

Erasure

The exposure of magnetic tape to a strong alternating magnetic field in order to leave the tape in a neutral state.

Self Erasure

The tendency for strongly magnetized sections of an audio tape to erase adjacent sections of opposite polarity magnetization. This is a significant cause of loss of high frequencies at lower tape speeds.

Excess Noise

A fudge factor to account for the extra noise components exhibited by passive electronic components that are not described by thermal noise effects.

Flutter

Rapidly repeating fluctuations in tape or turntable speed that give rise to warbling variations in the pitch of the reproduced sound. Flutter can be considered a higher frequency version of wow and typically measurements are made of wow and flutter combined.

Flux (Magnetic)

The magnetic force existing in the neighborhood of a magnetic pole (on an audio tape) can be represented by lines of force known as the magnetic flux. The recorded flux level of a tape is specified by the number of lines of force per unit track width of the tape and has units of nWb/m (nanowebers per meter). A typical reference flux level for tapes is 185nWb/m with less than 1% distortion being obtained at this level. In Europe the DIN reference level of 320nWb is used to calibrate equipment using peak reading program meters. (see also Dolby level).

Gap

The narrow slot between the pole pieces of the record or playback head of a tape machine.

Gap Length

The dimension of the gap in the direction of tape travel. When the gap length becomes comparable to the recorded signal wavelength there will be no output from the head. This is the most significant high frequency limitation of a tape recorder.

Gap Smear

Continual abrasion of the head by the tape causes head material to cold flow into the gap causing magnetic shorts.

Harmonic Distortion

A form of distortion characterized by the presence of spurious harmonics in the signal. It is usually measured by comparing the percentage amplitude of the spurious harmonics to the amplitude of the signal fundamental tone.

Head

A magnetic transducer used to record and/or playback tape signal.

Hyperbolic Head

A head, the pole pieces of which are shaped to follow a hyperbolic function. This helps to maintain good tape to gap contact — for separation d the loss is given by $55d/\lambda$

Note: The wavelength (λ) is the recorded wavelength. If d is measured in inches, the $\lambda = \dfrac{\text{tape speed (I.P.S.)}}{\text{Hz.}}$

Headroom

The margin betwen an actual signal operating level and the level that would cause substantial distortion. For a tape recorder this would be the level above zero VU that gives a (specified) distortion

Image Rejection

A superheterodyne receiver can usually respond to two frequencies whose difference from the local oscillator frequency is equal to the intermediate frequency. One is the desired frequency, the other is the image frequency. The receiver is tuned to the desired frequency and the RF level adjusted for a specified output (usually maximum sensitivity). The input signal is then switched to the image frequency and the RF level increased to obtain the same output. This change in levels is a measure (in dB) of the image rejection.

Input Sensitivity

A measure of a device's input signal requirement to produce a desired output. "High" sensitivity indicates a low input signal level whereas "Low" sensitivity implies a higher input signal requirement. Typical specifications on amplifier systems are: phono and mic, 2mV; auxiliary (radio) and tape, 200mV.

I.H.F.M.

Institute of High Fidelity Manufacturers

Intermodulation Distortion

Distortion characterized by the presence of the sum and the difference frequencies of the fundamentals and harmonics of two or more simultaneous tones being passed through a system.

The CCIR measurement procedure is to use a 1:1 ratio of test tones of nearly equal frequency with the distortion given by the amplitude of the beat note.

The SMPTE method used a 4:1 tone amplitude ratio.

Intermediate Frequency (I.F.)

In a superheterodyne receiver, the frequency to which the RF signal carrier frequency is converted by action of the local oscillator. Popular intermediate frequencies are: AM radio, 455kHz; AM automotive radio, 262.5kHz; FM radio 10.7mHz.

I.F. Rejection

A measure of the ability of a tuner to reject an RF signal at the intermediate frequency.

Limiter

An amplifier whose output signal has a constant amplitude when the input signal is above a certain specified level. Limiting amplifiers are used in FM IF strips to help eliminate spurious amplitude modulation of the signal. Limiters can also be found in audio equipment used to suppress short duration peak transients.

Limiting Threshold

The input signal level required for the amplifier output to be limited in amplitude. For an FM IF amplifier this is measured as the input signal level for which the output falls to 0.707 of the amplitude obtained with a strong signal input.

M.R.L.

Maximum recorded level. The signal level required at a given frequency to give a 3% distortion level on a tape (at mid-range frequencies — at high frequencies self-erasure determines the MRL). It is the performance ceiling for the particular tape with a given bias and equalization.

Medium Wave

A term applied by the CCIR to a frequency band between 300kHz and 3mHz.

In Europe this is popularly taken to mean the AM broadcast band encompassing carrier wavelengths between 580m and 190m. There is also a band designated long wave (LW) which includes wave lengths from 2000m to 1150m. (U.S. AM broadcast transmissions are in the MW band only).

Mixer

A circuit in which two separate frequencies, usually called the carrier and local oscillator signals, can be mixed or converted to the difference frequency between them (the intermediate frequency).

Microphone/Line Mixer

A device for adding two or more input signals in a linear fashion while exercising individual control over the amplitude of each.

Modulation Index (Modulation Percentage)

A measure of the degree of modulation of a carrier signal on a carrier waveform.

AM Modulation

For a sinusoidal modulation of a carrier waveform, if E_{MAX} is the peak waveform amplitude while the null amplitude is E_{MIN}, the modulation index m is given by the ratio

$$\frac{E_{MAX}/E_{MIN}-1}{E_{MAX}/E_{MIN}+1}$$

The modulation percentage is given by $m \times 100\%$.

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FM Modulation

The modulation index for a frequency modulated wave (mf) is given by the ratio of the peak frequency deviation of the carrier to the modulating frequency, i.e.

$$mf = \frac{\Delta f}{fm}$$

For Broadcast FM signals, modulation percentage has quite a different meaning and 100% modulation refers to the peak deviation permitted by the particular broadcast standard. For example, 100% modulation can be:

Radio Broadcast (U.S.): ±75kHz
NTSC Television (U.S.): ±25kHz
PAL Television: ±50kHz

Motorboating

Audible spurious low frequency oscillations in a system usually caused by inadequate de-coupling of power supply leads. Signals in an output stage couple back through the common internal impedance of the power supply to the input stages.

Multi-Path

Multi-path describes a signal condition whereby the antenna of a receiving system receives not only the directly radiated signal (line of sight) but also delayed signals reflected from large buildings or hills.For an FM receiver in an automobile the delay time caused by the longer reflected signal path changes as the car moves causing a "fluttering" of the recovered audio at the rate of change of phase between the direct and reflected signals. This is sometimes called "Picket Fencing."

Music Power

A measurement of the peak output power capability of an amplifer with either a signal duration sufficiently short that the amplifier power supply does not sag during the measurement, or when high quality external power supplies are used. This measurement (an IHF standard) assumes that with normal music program material the amplifier power supplies will sag insignificantly.

Mute

To suppress the audio output of an amplifier in response to a command signal even though an input may be present. Used in stereo receivers to prevent the off-channel spurious response produced while tuning from reaching the speakers.

N.A.B.

National Association of Broadcasters — usually associated with various tape standards.

Noise

A term for unwanted electrical distrubances, other than crosstalk or distortion components, that occur at the output of a reproducing amplifier.

Noise Bandwidth (see Bandwidth)

Noise, Excess (see Excess Noise)

Noise Figure

The logarithmic ratio of the input signal to noise and the output signal to noise ratios.

Noise Current

The equivalent open circuit RMS noise current which occurs at the input of a noiseless amplifier due to current flowing at that input. It is measured by shunting an impedance across the input terminals and comparing this output noise obtained with the output noise obtained when the input is shorted (see below).

Noise Voltage

The equivalent short circuit input RMS noise voltage which occurs at the input of a noiseless amplifier if the input teminals are shorted. It is measured at the output, divided by the amplifier gain and the square root of the bandwidth over which the measurement is made to yield units of $nV/\sqrt{Hz}$.

Flicker Noise

1/f or flicker noise has a random amplitude similar to shot and thermal noise but with a 1/f spectral power sensity. This means that the noise increases at low frequencies and is associated with the level of direct current in the device.

Popcorn Noise

So named for the audible characteristic, popcorn noise is randomly occurring, random amplitude noise, lasting from a few microseconds to several seconds.

Shot Noise

The noise generated by a charge crossing a potential barrier. For medium and high frequencies it is the dominant noise mechanism in bipolar devices. Shot noise has a constant spectral density.

Thermal Noise

Also called Johnson Noise, this mechanism of noise voltage generation occurs spontaneously in all resistive devices and involves the random thermal agitation of electrons. It has a constant spectral density and the noise voltage is given by Nyquist's formula

$$V_N = \sqrt{4KTBR}$$

Modulation Noise

The noise produced on playback of a tape that is a function of the instantaneous amplitude of the signal. It is caused by poor particle dispersion and surface irregularities.

Pink Noise

Noise that has a constant mean squared voltage (or power) per octave, i.e. the mean squared noise voltage per unit bandwidth increases at 3dB per octave (10dB per decade) with falling frequency. Noise sources with this characteristic are popular in audio work since it allows correlation between successive octave equalizer stages by ensuring that the same voltage amplitude is available as a reference standard.

White Noise

Noise with a constant spectral density — the mean square noise voltage per unit bandwidth is independent of frequency. Resistor thermal noise has this characteristic.

Pan-Pot

A potentiometer used to adjust the stereo balance of a monophonic signal allowing it to be positioned anywhere across the stereo stage.

Phantom

A signal derived from two sources in such a way as to appear located from a third source. Stereo signals "appearing" between speakers are said to be "phantomed".

Print-Through

The transfer of signal through adjacent layers of tape on a reel. It causes faint echoes preceding or following loud passages.

Quad

Generally taken to mean quadrasonic or quadraphonic sound systems designed to give the impression of a field of sounds coming from an apparent 360° around the listener.

Quieting

A measure of the usable sensitivity of an FM tuner and is expressed as the least RF signal level (100% modulated with a 400Hz tone) that reduces the receiver internal noise and distortion to 30dB below the output level obtained with the modulated tone present (S + N/N = 30dB, a null filter tuned to 400Hz is used to remove the tone).

For an AM receiver the carrier is modulated by 30% and the field strength (μV/m) is measured that is necessary to provide a 20dB S + N/N ratio.

R.I.A.A.

Record Industry Association of America

Usually referred to in connection with a phonograph disc recording equalization that helps limit the frequency and amplitude swings of a record cutting head over the audio frequency range. The reproducing amplifier has the inverse characteristic.

Retentivity

When a tape has a signal recorded on it the resulting magnetic field strength per unit coating cross section (width X thickness) is known as the retentivity of the tape.

Remanence

The magnetic field strength retained by a ¼" wide tape.

Reverberation

The persistence of sound in an enclosure after the original sound has ceased. Reverberation can be regarded as a series of multiple echoes closely spaced so as to appear continuous but gradually decaying in intensity. Electromechanical (or solid state) devices can be used to simulate reverberation with delay times from a few milliseconds and decay times up to 2 seconds using a frequency range from 100Hz to 5kHz.

Rumble

The name given to low frequency noise (below 100Hz) caused by turntable and tape transport mechanisms.

S.O.A.

Safe operating area — of a solid state device. The curves displaying the collector current and collector voltage limits of the transistor that must be observed for reliable operation. Curves indicating instantaneous power dissipations are often shown as well as dc limits.

Saturation

The condition of a tape coating that has accepted its maximum degree of magnetization. It can also refer to an amplifier output that is at the point of clipping.

Selectivity

A tuner's ability to select a station in the presence of strong adjacent or alternate channel signals.

For FM, either 30% or 100% modulated signals can be used and involves the measurement of the ratio of the desired carrier at the usable sensitivity level (30dB quieting) and the level of undesired carrier (0.2mHz away for adjacent channel and 0.4mHz away for alternate channel) needed to cause a 30dB reduction in the tuner recovered audio level.

For AM receivers a 20dB reduction is required with the undesired carrier located 10kHz away (adjacent channel) or 20kHz away (alternate channel).

Sensitivity

See Input Sensitivity

FM Sensitivity:

The radio frequency input signal (μV) required to produce 30dB quieting of the recovered audio (also called usable sensitivity). The carrier is modulated to ±75kHz deviation with a 400Hz tone.

AM Sensitivity:

The radio frequency input field strength required to produce 20dB quieting of the recovered audio. A 30% modulated carrier is used (IMF standard).

A popular technique with O.E.M.'s is to measure the field strength required to produce a given level at the speaker — for table radios this is 50mW, for automotive radios it is 1 watt.

Microphone Sensitivity

The output voltage in dB referenced to 1 volt for an S.P.L. of 1 microbar (74dB SPL).

A microphone with a sensitivity of −85dBV will have an output of 5.6V for an S.P.L. of 174dB (output = 174dB − 74dB − 85dB = +15dB above IV).

S/N

The ratio of a system's output signal level and the noise level obtained in the absence of signal. The reference signal level is either specified or measured as that which related to a specified distortion level.

Sinad

A measurement of the signal to noise ratio of a receiver system where the signal level measurement includes the system noise and distortion (S + N + D)/N.

Skating

The tendency of a pivoted tone arm to be pulled to the center spindle. It is caused by friction between the stylus and the record surface.

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S.P.L.

Sound pressure level — usually measured with a microphone/meter combination calibrated to a pressure level of 0.0002μBars (approximately the threshold hearing level).

$$S.P.L. = 20 \, Log_{10} \, P/0.0002 dB$$

where P is the R.M.S. sound pressure in microbars. (1 Bar = 1 atmosphere = 14.5 lb/in^2 = 194dB S.P.L.).

Squelch

An audio squelch is one that cuts off (or mutes) the output of the audio section of a receiver when there is no input signal. It is used to prevent listener fatigue on communications channels caused by noise in the absence of carrier signals.

S.C.A.

Literally subsidiary communications authorization. This applies to an additional modulation on the standard FM carrier (see composite signal) intended to provide commercial-free background music for stores, etc.

Thermal Resistance (R_{TH})

An analogy for heat transfer where the ability of a heat conductive system to transfer heat is described in similar terms to those used in an electrical system for power dissipated in a resistor with a given applied voltage. The thermal resistance is given by the temperature differential established when a given amount of power is being dissipated ($\theta = T1 - T2/P_D$) with units of °C/watt.

Tweeter

A loudspeaker designed for high frequencies (see cross-over frequency).

Ultrasonic Rejection

The level of rejection of the 19kHz pilot tone and 38kHz V.C.O. frequency in a stereo FM receiver. The intrinsic rejection of a stereo decoder is the logarithmic ratio of the level of 19kHz and 38kHz to a 1kHz reference tone with only the standard de-emphasis filter at the decoder outputs.

VU

The abbreviation for Volume Unit, a form of decibel referenced to a standard value of 1mW in a 600Ω load.

VU Meter

A recording level meter with a needle motion damped according to internationally recognized standards, which will respond to 99% of the input signal within 0.3 seconds and have less than 1.5% overshoot. The frequency response also has to be better than ±0.5dB from 25Hz to 16kHz.

Woofer

Speaker designed to reproduce relatively low frequencies.

Wow

A slow variation in the pitch of a reproduced signal caused by tape or turntable speed variations (see flutter).

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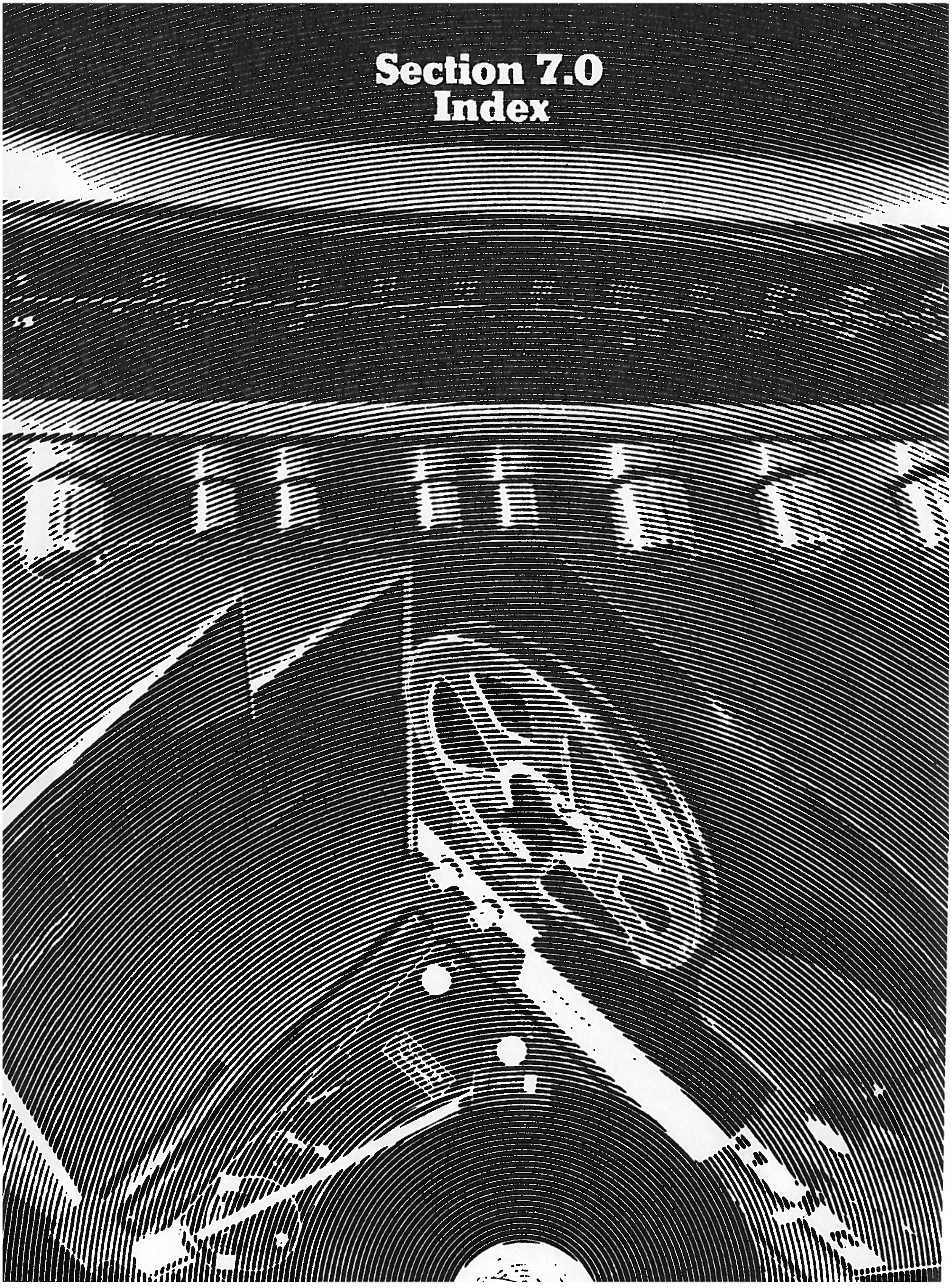

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