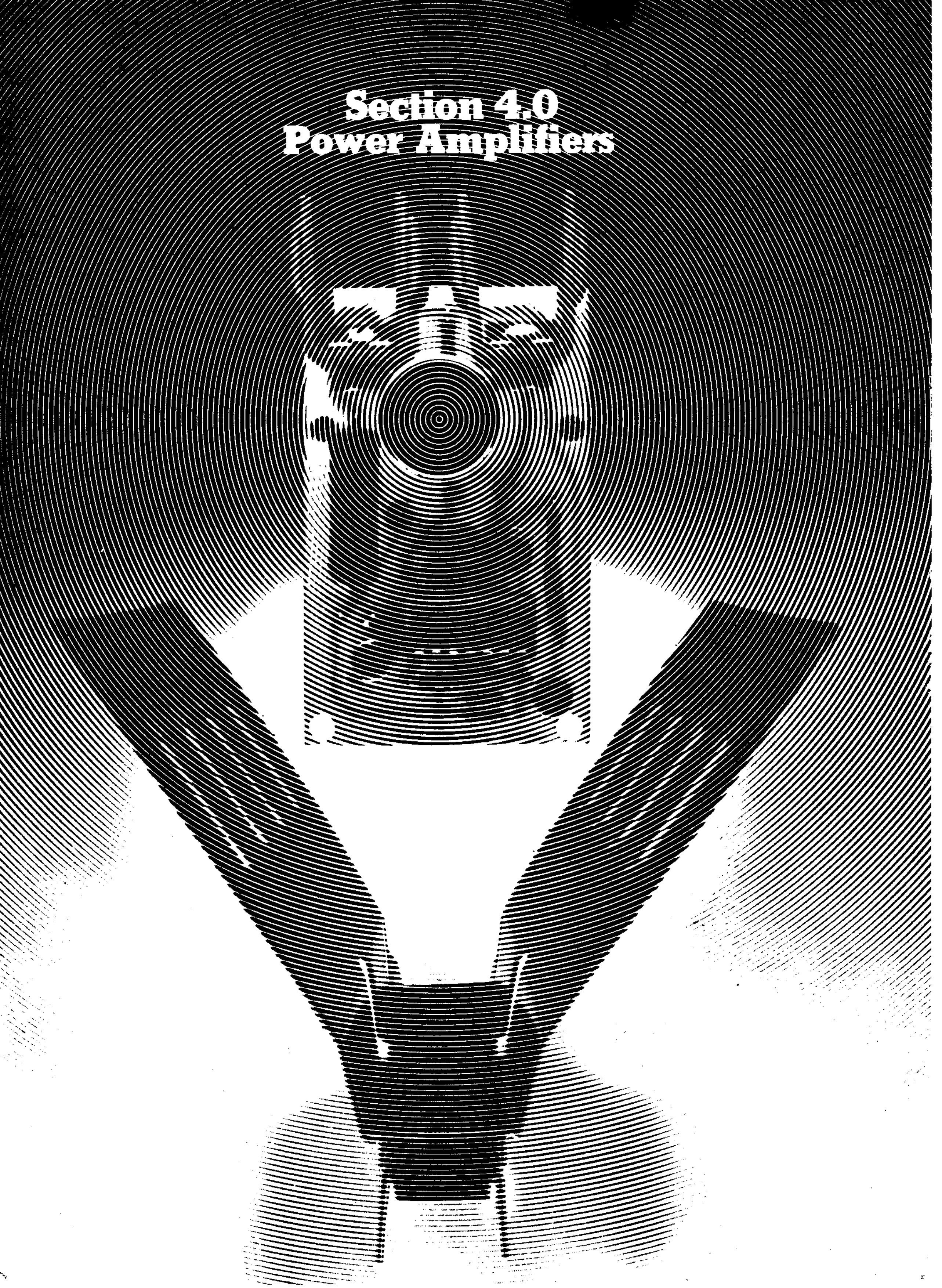
Section 4.0
Power Amplifiers



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:

$$v_o = v_1 \, gm \, X_c \tag{4.1.1}$$

or, rewriting in terms of gain:

$$A_V = \frac{v_o}{v_1} = gm \, X_c = \frac{gm}{sC} = \frac{gm}{j\omega C} \tag{4.1.2}$$

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

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

$$f_{UNITY} = \frac{gm}{2\pi fC} \tag{4.1.4}$$

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 the LM380 or LM377 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.

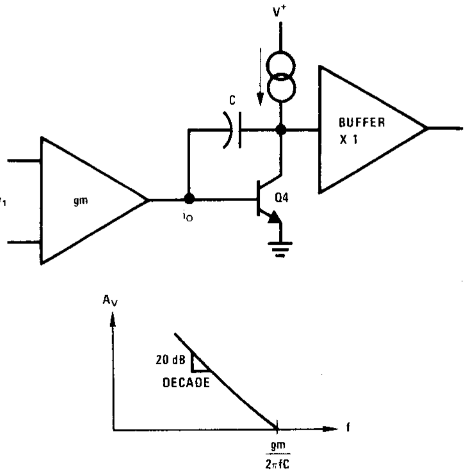


FIGURE 4.1.1 Audio Amp Small Signal Model

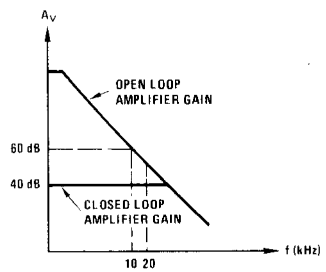


FIGURE 4.1.2 Feedback and "Loop Gain"

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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.

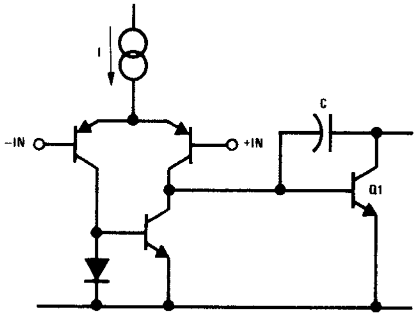

FIGURE 4.1.3 Typical gm Block

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 20 kHz sine wave swinging 40 $V_{p\text{-}p}$, 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 2.5 V/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 2-5 MHz 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 LM377, LM378, 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.

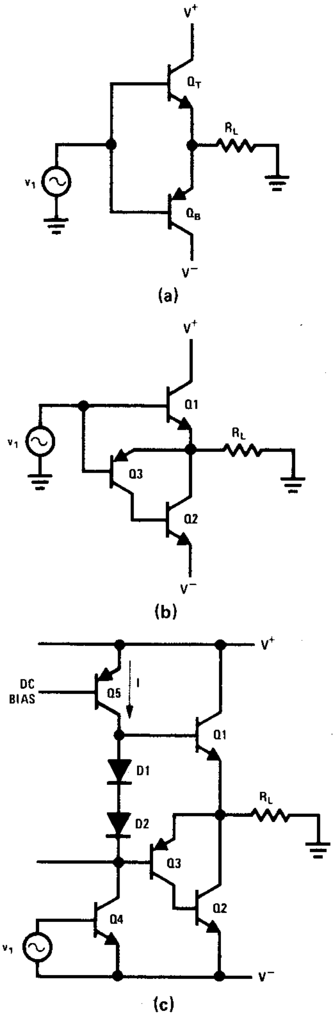

FIGURE 4.1.4 Basic Class B Output Drivers

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.

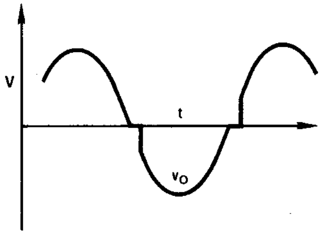

FIGURE 4.1.5 Crossover Distortion

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!

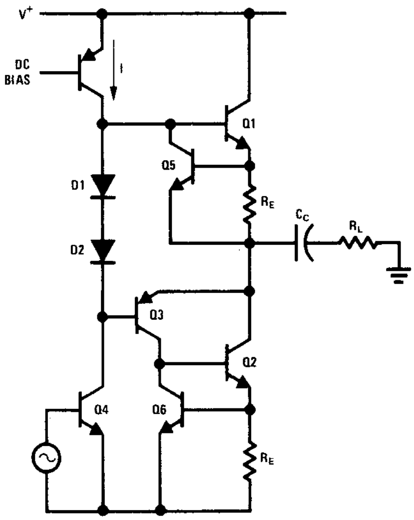

FIGURE 4.1.6 Simple Current Limit

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

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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.

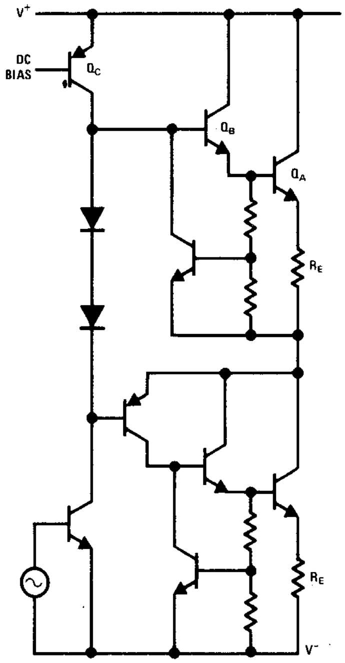

FIGURE 4.1.7 Improved Current Limit

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.

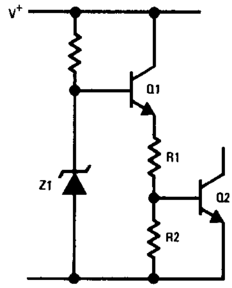

FIGURE 4.1.8 Typical Thermal Shutdown

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.

4.2 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 the LM377, LM378 and LM379 family, and the "Monos," represented by six products. Available power output ranges from miniscule 320mW battery operated devices to hefty 7W line operated systems. 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 six mono devices. Figures 4.3.1-4.3.3 provide graphical comparison of power output versus supply voltage for loads of 4, 8 and 16 ohms.

TABLE 4.3.1 Dual Power Amplifier Characteristics

PARAMETER	LM377N (14 Pin DIP)			LM378N (14 Pin DIP)			LM379[2]			UNITS
	MIN	TYP	MAX	MIN	TYP	MAX	MIN	TYP	MAX	
Supply Voltage	10	20	26	10	24	35	10	28	35	V
Quiescent Supply Current (P_{OUT} = 0W)		15	50		15	50		15	65	mA
Output Power[3]										
THD ≤ 5%	2	2.5		4	5			6		W
THD = 10%							6	7		W
Total Harmonic Distortion										
P_{OUT} = 1W/CH, f = 1kHz		0.07	1		0.07	1		0.07	1	%
P_{OUT} = 2W/CH, f = 1kHz		0.10			0.10					%
P_{OUT} = 4W/CH, f = 1kHz								0.20		%
Input Impedance	3			3			3			MΩ
Open Loop Gain ($R_s = 0\Omega$)	66	90		66	90		66	90		dBV
Channel Separation ($C_F = 250\mu F$, f = 1kHz)	50	70		50	70		50	70		dBV
Ripple Rejection (f = 120Hz, $C_F = 250\mu F$, input referred)	60	70		60	70			70		dBV
Slew Rate		1.4			1.4			1.4		V/μs
Equivalent Input Noise Voltage ($R_s = 600\Omega$, 100Hz-10kHz)		3			3			3		μV_{RMS}

1. Specifications apply for T_{TAB} = 25°C, $R_L = 8\Omega$, A_v = 50 (34dB), V_s = 20V (LM377), V_s = 24V (LM378), V_s = 28V (LM379), unless otherwise specified.
2. LM379S = 14 Pin "S" Type Power DIP.
3. For operation at ambient temperatures greater than 25°C the IC must be derated based on a maximum 150°C junction temperature using a thermal resistance obtained from device data sheet.
4. Output protection included on all devices.

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TABLE 4.3.2 Dual Audio Amplifier Typical P_o @ 10% THD

Supply	Device			Load Impedance 8Ω	Load Impedance 16Ω
12	LM377	LM378	LM379	1.6W	–
16				2.2	1.5W
18				3.0	1.8
20				3.8	2.4
22				4.6	2.8
24				5.4[1] 1.5[2]	3.6
26				6.0	4.2
28				7.0	5.0
30				–	5.5

1. LM379.
2. LM378 (thermal limit).

TABLE 4.3.3 Mono Power Amplifier Characteristics

Device	Supply Voltage (V)			Output Power (W) at 10% THD									Quiescent Current (mA)			Fixed Gain (dB)			Gain Control (Typ dB)	Output Protection
				$R_L = 4\Omega$			$R_L = 8\Omega$			$R_L = 16\Omega$										
	Min	Typ	Max	Min	Typ	Max	Min	Typ	Max	Min	Typ	Max	Min	Typ	Max	Min	Typ	Max		
LM386 (8 Pin DIP) [LM389[2]]	4		12[4]																	
		6			0.32		0.25	0.32			0.18			4, [6[2]]	8, [12[2]]		26		YES (46)	No
		9						0.5			0.5									
		12						0.3			0.9									
LM388 (14 Pin DIP)	4		12[4]																	
		6		0.45	0.6			0.4			0.2					23	26	30	YES (46)	No
		9			1.2			1.0			0.6									
		12			2.0		1.0	1.5			1.1			10	20					
LM390 (14 Pin DIP)	4		10																	
		6		0.8	1.0			0.6			0.34					23	26	30	YES (46)	No
		9			2.0			1.2			0.77			10	20					
LM380 (14 Pin DIP)	8		20																	
		12			2.5			1.5			0.5					32	34	36	No	YES
		14			3.3			2.2			1.0									
		18			4.2		2.5[3]	4.0			2.2			7	25					
LM384 (14 Pin DIP)	12		26																	
		18			4.2			4.0			2.2						34		No	YES
		22			3.5		5.0	5.7			3.5			8.5	25					

1. Specifications apply for $T_A = 25°C$. For operation at ambient temperatures greater than 25°C the IC must be derated based on a maximum 150°C junction temperature using a thermal resistance obtained from device data sheet.
2. LM389 identical to LM386, but includes three additional NPN transistors pinned out separately for customer use. 18 Pin DIP.
3. THD = 3%.
4. Parts selected for higher absolute maximum supply voltage available on special request.

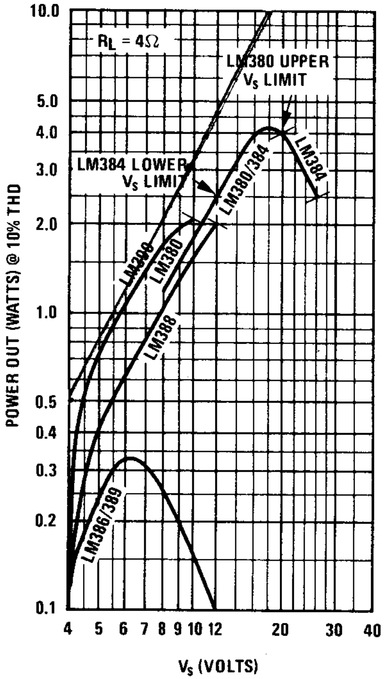


FIGURE 4.3.1 P_O vs. V_S for R_L = 4 Ohms

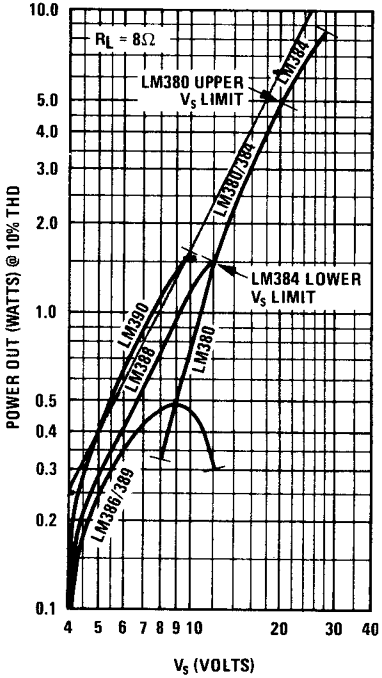


FIGURE 4.3.2 P_O vs. V_S for R_L = 8 Ohms

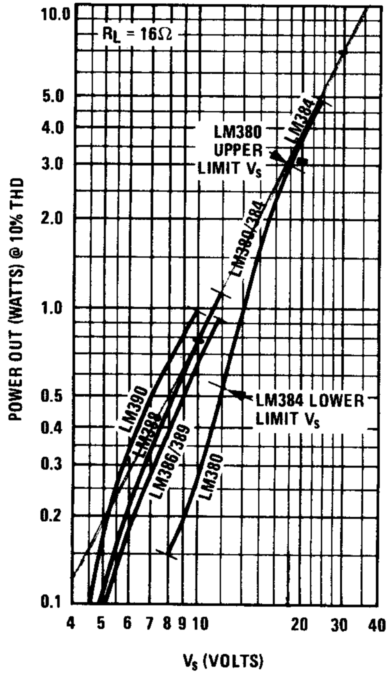


FIGURE 4.3.3 P_O vs. V_S for R_L = 16 Ohms

4.4 LM377, LM378, AND LM379 DUAL TWO, FOUR, AND SIX WATT POWER AMPLIFIERS

4.4.1 Introduction

The LM377, LM378 and LM379 are two-channel power amplifiers capable of delivering 2, 4, and 6 watts respectively 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 10 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

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 LM377 series amplifiers, Q_9 is made 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

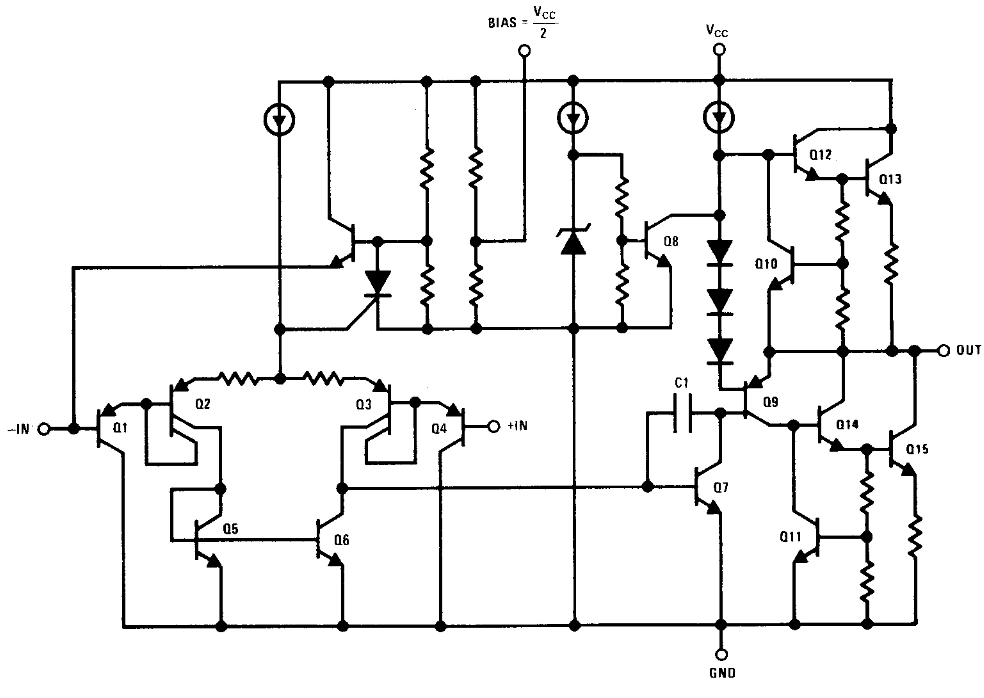


FIGURE 4.4.1 Simplified Schematic Diagram

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.

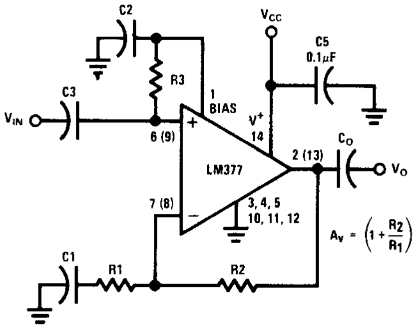


FIGURE 4.4.2 Non-Inverting Amplifier Connection

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:

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

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

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

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

The 3T delay might normally be considered excessive! The LM377 series 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

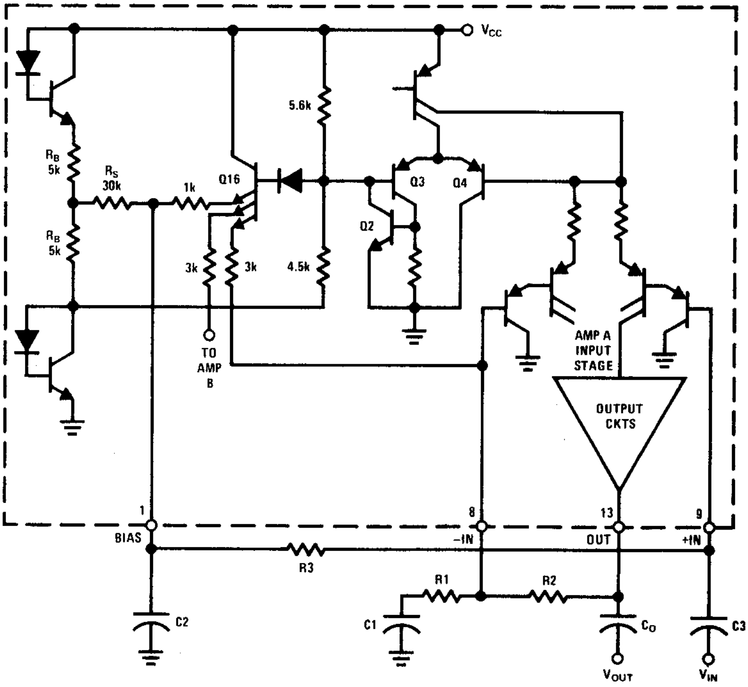


FIGURE 4.4.3 Internal Turn-On Circuitry

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.

4.3.3 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.

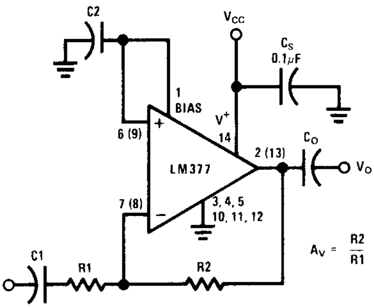
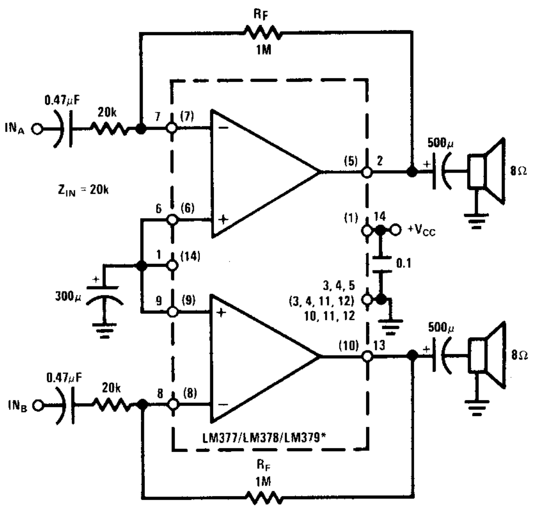


FIGURE 4.4.4 Inverting Amplifier Connection

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 2/4/6 Watt Stereo Amplifier Applications

The obvious and primary intended application is as an audio frequency power amplifier for stereo or quadraphonic music systems. The amplifier 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 4.4.5 shows the total parts count for such a stereo amplifier. The feedback resistor value of 1 meg in Figure 4.4.5 is about the largest practical value due to an input bias current max of approximately $1/2\mu A$ (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



*(LM379S pin nos. in parentheses)

		LM377	LM377/LM378	LM379
P_O	=	2W/CH	3W/CH	4W/CH
e_i	=	80mV MAX	98mV MAX	113mV MAX
A_V	=	50	50	50
VCC	=	18V	24V	28V

FIGURE 4.4.5 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 4.4.6, which has a superior input impedance. For applications utilizing high impedance tone and volume controls, the non-inverting connection will normally be used.

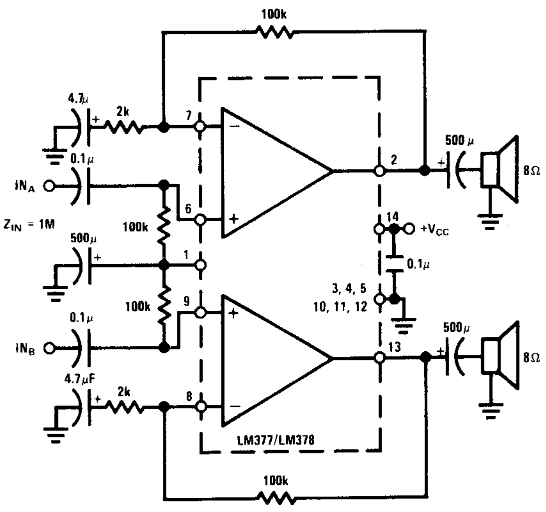
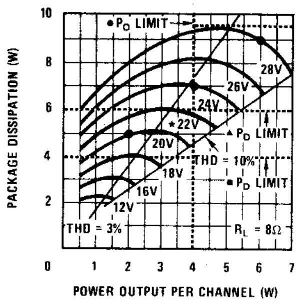
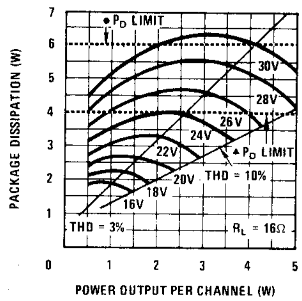


FIGURE 4.4.6 Non-Inverting Stereo Amplifier



- ● Approx. P_O limit acc't. 0.7A rms internal current limit at oper. die temp.
- ▲ P_D limit for LM377/LM378 on PC board w/Staver V7-1 heat sink.
- ■ P_D limit for LM377/LM378 on PC board (2.5 sq. in. Cu)
- ★ Safe limit for LM377.

FIGURE 4.4.7 Device Dissipation for 8Ω Load



- ● P_D limit LM377/LM378 on PC board w/Staver V7-1.
- ▲ P_D limit LM377/LM378 on PC board.

FIGURE 4.4.8 Device Dissipation for 16Ω Load

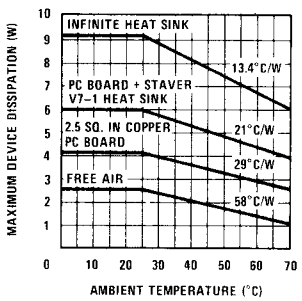


FIGURE 4.4.9 LM377/LM378 Power Derating

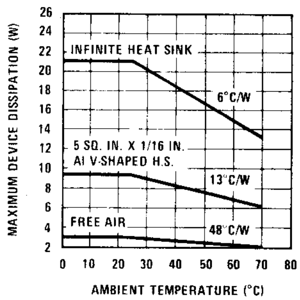


FIGURE 4.4.10 LM379 Power Derating

The prime limitations on output power of the LM377 and LM378 will be the type of heat sink employed, supply voltage, and load resistance. Reference to the data sheet curves will indicate the most efficient supply voltages to use for specific power output levels with 8 or 16Ω loads. The pertinent curves are reproduced in Figures 4.4.7 through 4.4.10. For other conditions $P_C = V_{CC}^2/20R_L$. At high power out, efficiency exceeds 50% and dissipation drops below output power. A dual 2W amplifier must then dissipate about 4.0W with an 18V supply or 4.9W with a 20V supply when $R_L = 8\Omega$. Normally, one would choose the 18V supply for lower dissipation; however, the 20V supply allows reduced distortion levels or considerably higher powers. A dual 4W amplifier will dissipate about 8W with a 26V supply. This is above the dissipation limit for an LM378 with normal heat sink. Accordingly, a fairly efficient heat sink must be employed in order to allow full continuous output from the LM378 (Figure 4.4.9). The recommended heat sinks are listed in Table 4.4.1 with measured power output levels at V_S = 18 to 29V for LM377 and LM378 (observe voltage limits on LM377) with 8 or 16Ω load.

4.4.5 Power Output per Channel (Both Channels Driven) Before Clipping

Power dissipation vs. power output/channel (both channels driven) is indicated in Figures 4.4.7 and 4.4.8 for load resistances of 8 and 16Ω.

Limiting points to keep in mind, noted on figures 4.4.7 and 4.4.8, are 4W package dissipation limit for LM377/LM378 when soldered to PC board with 2.5 sq. in. copper, 6W limit when a Staver V7-1 heat sink is added, and internal current limit at about 1.5A peak at 25°C die temperature, reducing to about 1A peak at operating die temperature.

TABLE 4.4.1 Continuous Power Out (Both Channels) Before Clipping

HEAT SINK	LM377 $R_L = 8\Omega$ $V_s = 18V$	$V_s = 20V$	$V_s = 22V$	LM378 $R_L = 16\Omega$ $V_s = 24V$	$V_s = 26V$	$V_s = 29V$
PC Board, 29°C/W	2.2W	0.8W	0.3W	2.2W	1W	0.3W
PC Board and Staver V7-1, 21°C/W	2.2W	2.5W	3.2W	2.2W	2.5W	3.3W

This results in an approximate P_O = 4W/channel limit for R_L = 8Ω. The onset of clipping occurs just to left of the THD = 3% line in Figures 4.4.7 and 4.4.8.

The overall result is that the LM377 and LM378 with practical heat sinks are limited to operation below package dissipation of 6W and below P_O = 4W/channel when R_L = 8Ω. Thus, maximum P_O = 3W/channel before clipping or 4W/channel at about 6% THD with either device at V_{CC} = 22V. With a 16Ω load the LM378 can deliver 4W/channel with 3-4% THD when V_{CC} = 29-30V. The LM379 is limited to P_O = 4-5W/channel before clipping at V_{CC} = 26-28V, R_L = 8Ω, or 4W/channel at V_{CC} = 30V, R_L = 16Ω. P_O = 6W occurs at 8-10% THD with V_{CC} = 28-30V and R_L = 8Ω. Note that the P_O = 6W rating on LM379 is at 10% THD where peak current is similar to that at P_O = 4W, V_{CC} = 26V, R_L = 8Ω.

What really exists then are power out before clipping of 2W/channel at V_{CC} = 18V with PC board mounting, 3W/channel at V_{CC} = 22V with maximum practical heat sinking on either LM377 or LM378, and 4W/channel at $V_{CC} \geqslant$ 26V for LM379. Clarification of how to interpret Figures 4.4.7-4.4.10 is presented as Section 4.4.6 for those who are not yet confused!

4.4.6 Interpretation of P_O vs. P_D Curves

The angled straight lines on the curves of Figures 4.4.7 and 4.4.8 indicate the loci of operating points where clipping occurs. When THD = 3%, the output waveform has noticeable clipping. The THD = 10% line is an operating area of severe clipping. Clipping begins just to the left of the THD = 3% line, so this discussion deals only with operation up to, but not quite at, the 3% line.

The three circles on Figure 4.4.7 are the data sheet spec limits for LM377/LM378/LM379; that is, 2, 4 and 6W/channel with 20, 24 and 28V supplies respectively.

Observe that the 2W point is well to the left of the THD = 3% line, or well under clipping. The 4 and 6W points march progressively further toward the THD = 10% line, or deeper into clipping. Also note the dissipation limits in Figure 4.4.9 for LM377/LM378 on PC board and on PC board with addition of Staver V7-1 heat sink are 4.1 and 6W respectively. These represent the limits for commonly available heat sinks for the DIP package. No doubt a special heat sink fabricated "just-so" could extend the 6W limit to 6-1/2 or 7W, but we'll stop at 6W. Data have been added to Figure 4.4.10 showing LM379 dissipation with a simple small heat sink. This heat sink is 5 square inches of 1/16" aluminum in a modified V shape which is clamped to the sink side of the LM379.

These practical limits are transferred to Figure 4.4.7 as horizontal dashed lines across the P_O vs. P_D curves at 4.1, 6 and 9.6W. We see that the reference points, 2W at 20V and 4W at 24V, are above the practical P_D limits for PC board alone and for PC board with Staver V7-1 heat sink. The third point, 6W at 28V, is o.k. so far. What may be most important, however, is performance short of clipping. For that reason, the remainder of this section will deal only with RMS power at levels below clipping.

Returning to Figure 4.4.7, it appears that the LM377 or LM378 with only PC board heat sinking will be able to deliver 2.2W/channel into 8Ω with an 18V supply. But, if the supply is raised to 20V, the P_D limit is exceeded at 1W. With PC board plus a Staver heat sink, the LM377/LM378 will deliver 3.2W/channel with 22V supply, yet raising the supply to 24V limits us to P_O = 1.9W/channel.

So why use a LM378 if the supply limit is 22V? The reason is that few supplies are regulated in the consumer world. This means that if the supply is 22V under full load, the no-signal supply may rise 10% or more; and the variations in line voltage may add another 10% for a total supply maximum of at least 26.5V. Therefore, the LM377 is only recommended for full-load operating supplies of less than 20V. But remember, it can deliver over 2W/channel with an 18V supply on a PC board, or 2-1/2W/channel with 20V supply and Staver heat sink. The LM378 will provide 3.2W/channel with 22V supply and PC board plus Staver heat sinking. With poorly regulated supplies over 20V or with 16Ω load, the LM378 is the obvious choice, as higher supply voltages are required to obtain high powers with 16Ω loads, although no greater power is available than with 8Ω loads.

There is no reasonable P_D limit on the LM379, as we can dissipate nearly 20W with adequate practical heat sinking and 9.6W with minimal sink. Then V_{CC} is the limit, say 30V. That would put us off the graph on Figure 4.4.7 at about 5.5W/channel or at 3W/channel with 16Ω load. Even at 8Ω and 30V, package dissipation is only 11W, or 9.6W with 28V. The kicker is in the data sheet electrical characteristics under current limit: 1.5A typ when T_{TAB} = 25°C. The tab is above 25°C when package dissipation is 9-11W. Still, this is a realistic test for high speed machine testing. In actual use, the current limit moves down to maybe 1.25A or even less. What does this mean? Consider an 8Ω load in power equation, and that 1-1.25A pk is 0.7-0.88A RMS.

$$P = I^2R$$

$$= (0.7)^2 8 = (0.5)8 \qquad \text{or} = (0.88)^2 8 = (0.77)8$$

$$= 4W \qquad\qquad = 6.2W$$

Now we have the actual limits at $P_{O(MAX)}$ = 4-6W at 8Ω or 8W at 16Ω. Trouble is we are limited to 5W at 28V, 8Ω or 5.5W at 30V, 8Ω and 2W at 16Ω by a 30V operating limit. Current limits could run higher than data sheet

TABLE 4.4.2 Max P_O Before Clipping (8Ω Load)

Heat Sink:		PC BOARD (29°C/W)			PC BOARD + V7-1 (21°C/W)				13°C/W SINK		
V_{CC}	=	16	18	19	18	20	22	23	26	28	30
P_O/CH	=	1.5	2.2	1.4	2.2	2.5	3.2	1.9	4.3	5.0	5.5

LM377 (16–20V)

LM378 (16–23V)

LM379 (16–30V)

typicals; many do, in fact. Then we can get more than 4W/channel as a limit. Since this is a typical spec, there is no guarantee either way.

Note with interest that an LM377 with Staver V7-1 heat sink will deliver 3.2W/channel with 22V supply (but hold it close to 22V or use a LM378) and the LM379 will deliver 5W/channel with a 28V supply. The LM379 is the practical choice because it is easier and probably cheaper to heat sink, and there is more P_D headroom to allow for variations in supply voltage (very important). Also, the better the heat sink on the LM379, the lower the tab temperature, and the higher the operating current limit.

Beyond the limits discussed, the temperature or current limits operate, the peaks are clipped, the waveform remains at peak value for a longer portion of the input cycle, the RMS P_O increases, P_D decreases, and RMS power approaches peak power.

Table 4.4.2 summarizes performance the customer may encounter.

4.4.7 Heat Sinking

Device dissipation vs. ambient temperature with several heat sink types is indicated in Figures 4.4.9 and 4.4.10 for convenience of matching heat sink capacity to the circuit needs. In those cases where heat sink capacity is inadequate for device dissipation requirements, the internal thermal limit circuitry will automatically limit device dissipation on signal peaks. The result is similar to peak clipping in its effect and causes severe distortion. The device can provide momentary peak power output in excess of the maximum heat sink limited steady-state levels for a second or so, depending upon the margin between maximum steady-state level and the actual average power level prior to the peak demand. Once in thermal limiting, clipping occurs on each positive and/or negative half cycle of a steady waveform.

In the majority of audio amplifier applications, the heat sinking can be considerably smaller due to the approximately 30dB ratio between RMS and peak power levels in music and speech. If we assume willingness to accept clipping at peak levels 20dB above average level, then average power levels will be 0.2-0.3W/channel in LM377 and LM378. Heat sink requirements are thus significantly reduced, as these peak levels occur less than 10% of time periods of several seconds duration. Thus, the circuit does not go into thermal overload even though the heat sink is designed for 3W dissipation (LM377 operating at 0.3W/channel, V_{CC} = 18V).

4.4.8 Stabilization

The LM377 series amplifiers are internally stabilized so external compensation capacitors are not required. The high gain x BW provides a bandwidth greater than 50kHz, as seen in Figure 4.4.11. These amplifiers are, however, not intended for closed loop gain below 10. The typical Bode plot of Figure 4.4.12 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 x 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 LM377/378/379 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 4.4.13) requires 6 resistors and 9 capacitors (excluding power supply parts) and uses the typical power supply shown. The same circuit using split supplies (Figure 4.4.14) 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

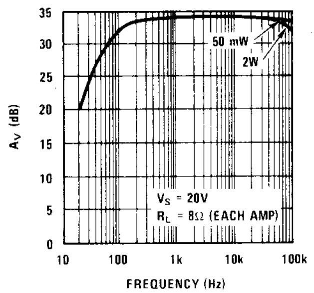


FIGURE 4.4.11 Frequency Response of the Stereo Amp of Figure 4.4.5

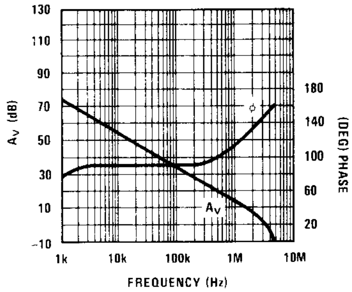


FIGURE 4.4.12 Open Loop Bode Plot (Approximately Worst Case)

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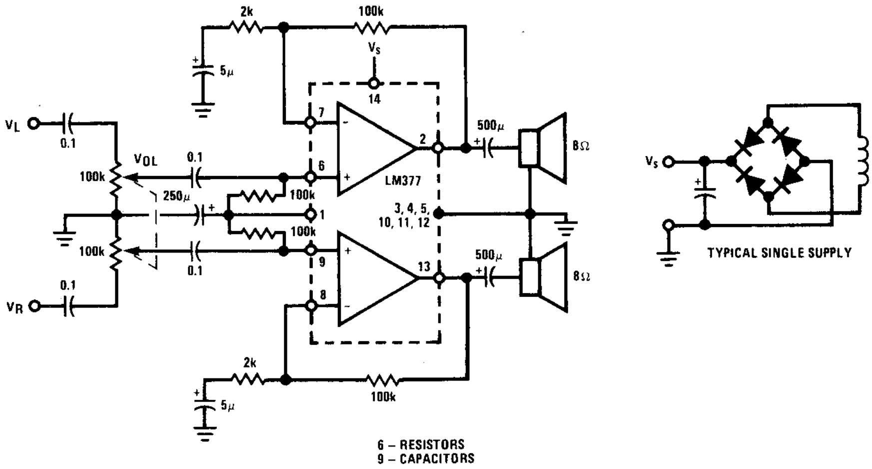


FIGURE 4.4.13 Non-Inverting Amplifier Using Single Supply

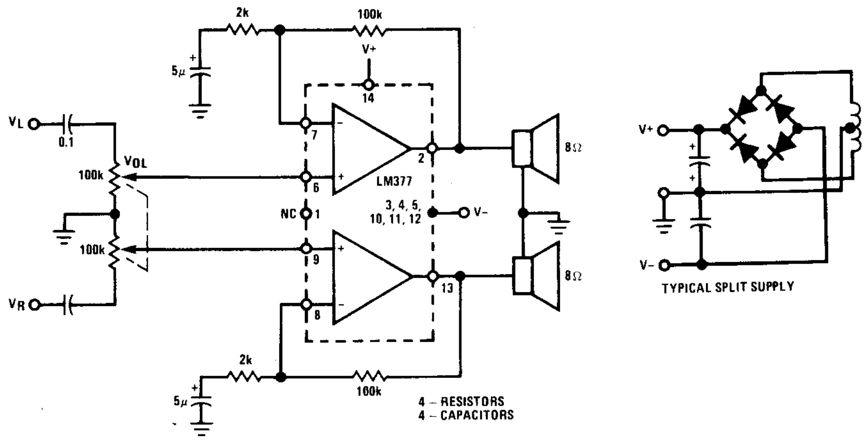


FIGURE 4.4.14 Non-Inverting Amplifier Using Split Supply

are omitted to allow bias currents from the positive inputs to flow directly through the volume pots to ground.

The complexity of the power supply has been increased only slightly, requiring a center-tapped secondary winding and one additional capacitor. Admittedly the added cap is a large electrolytic but its addition allows removal of three large electrolytics from the power amplifier stage, and since the supply is common to the whole system, the cost of the capacitor is shared with other subsystems benefiting from its use.

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 LM377/378/379 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 4.4.15) 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 2.8.7.

Non-inverting unity gain stability (Figure 4.4.16) 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:

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

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 (20kHz as shown) and f_P prior to hitting the open loop response (340kHz as shown) the frequency response of Figure 4.4.17 is obtained. This response satisfies the unity gain requirements, while allowing the gain to raise beyond audio to insure stable operation.

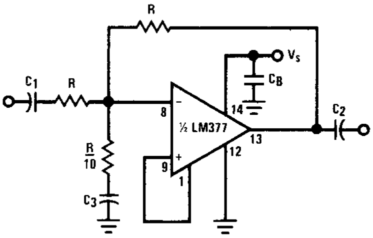


FIGURE 4.4.15 Inverting Unity Gain Amplifier

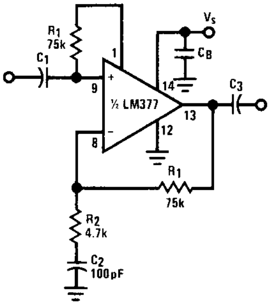


FIGURE 4.4.16 Non-Inverting Unity Gain Amplifier

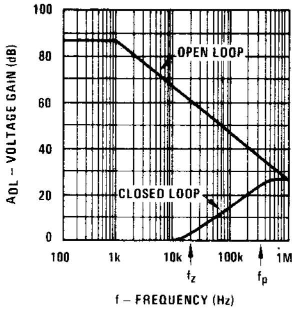


FIGURE 4.4.17 Frequency Response of Non-Inverting Unity Gain Amplifier

4.4.12 Bridge Amplifiers

The LM377 series 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 4.4.18. Response of this circuit is 20Hz to 160kHz as shown in Figure 4.4.19 and distortion is 0.1% midband at 4W, rising to 0.5% at 10kHz and 50mW output (Figure 4.4.20). 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 4.4.21 is similar except for higher input impedance. In Figure 4.4.21 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 20kHz (Figure 4.4.22). Frequency response is 27Hz to 60kHz as shown in Figure 4.4.23. The low frequency roll off is due to the double poles $C_3 R_3$ and $C_4 R_1$.

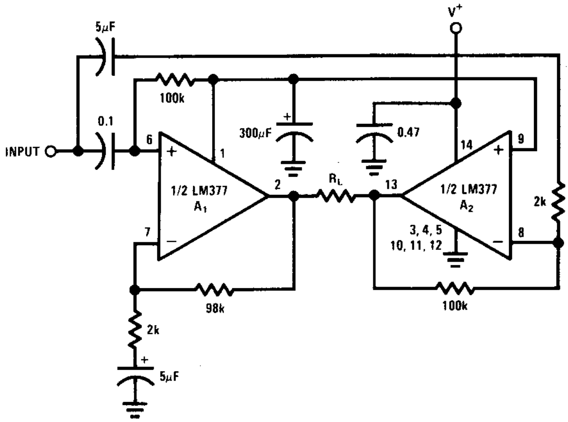


FIGURE 4.4.18 4-Watt Bridge Amplifier

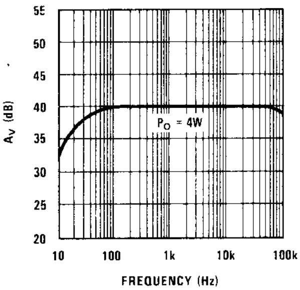


FIGURE 4.4.19 Frequency Response, Bridge Amp of Figure 4.4.18

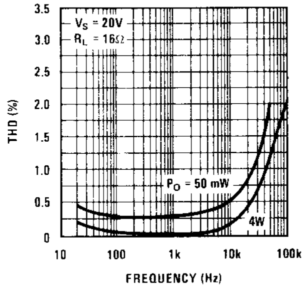


FIGURE 4.4.20 Distortion for Bridge Amp of Figure 4.4.18

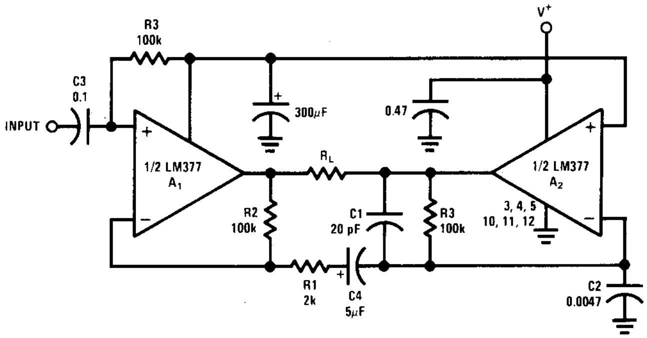


FIGURE 4.4.21 4-Watt Bridge Amplifier with High Input Impedance

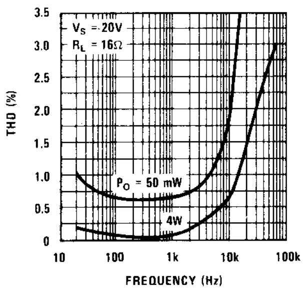


FIGURE 4.4.22 Distortion for Bridge Amp of Figure 4.4.21

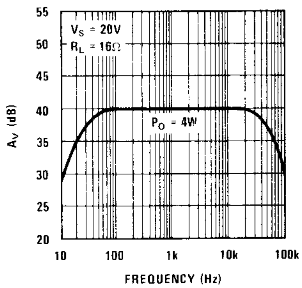


FIGURE 4.4.23 Frequency Response, Bridge Amp of Figure 4.4.21

4.4.13 Power Oscillator

One half of an LM377 may be connected as an oscillator to deliver up to 2W to a load. Figure 4.4.24 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 60Hz. 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 $0.1\mu F$ ceramic, or $0.47\mu F$ 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.

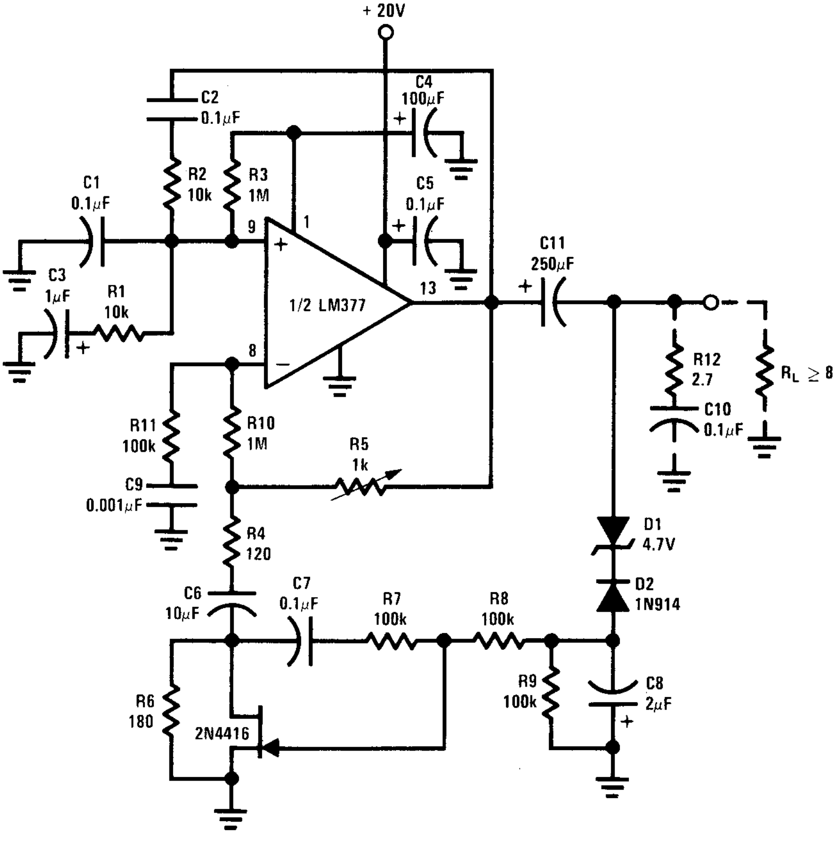


FIGURE 4.4.24 Wien Bridge Power Oscillator

4.4.14 Two-Phase Motor Drive

Figure 4.4.25 shows the use of the LM377 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

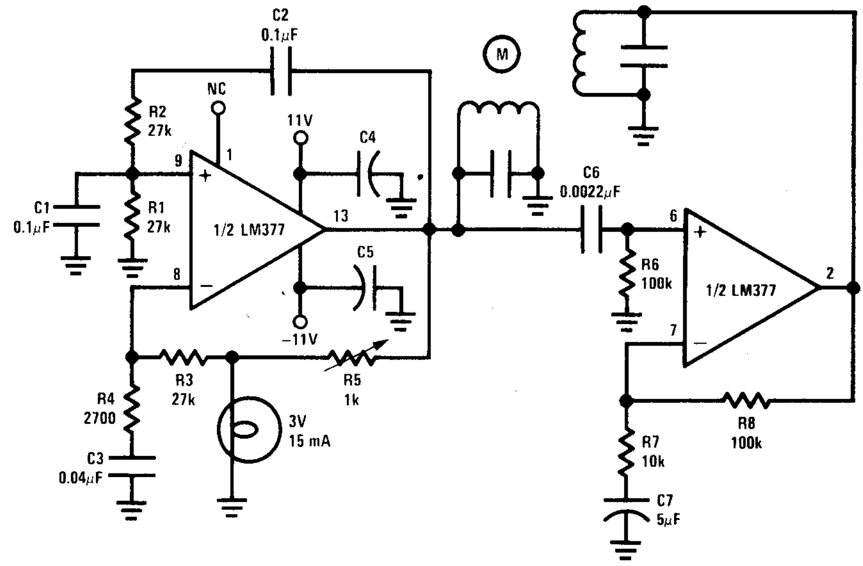


FIGURE 4.4.25 Two-Phase Motor Drive

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 milli-amps, 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 4.4.26) 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

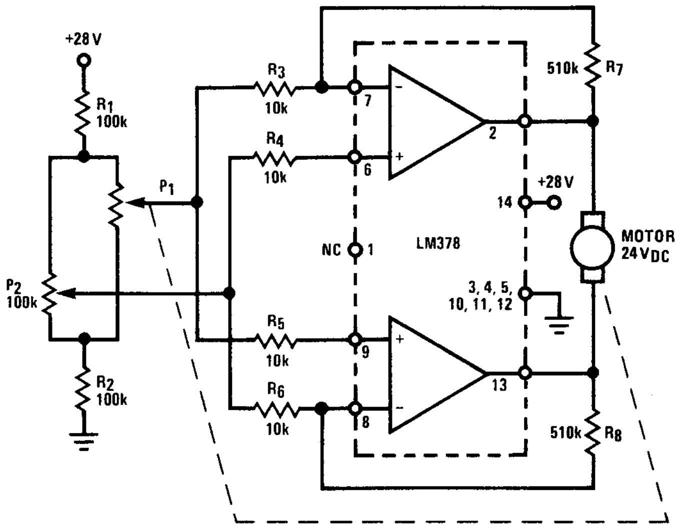


FIGURE 4.4.26 Proportional Speed Controller

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 LM377 to LM379 dual power amplifiers are useful in table or console radios, phonographs, tape players, intercoms, or any low to medium power music systems.

Figures 4.4.27 through 4.4.29 describe the complete electronic section of a 2-channel sound system with inputs for AM radio, stereo FM radio, phono, and tape playback. Figure 4.4.27 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 input 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 $Q_1 Q_2$ would be lost. It is believed that this circuit represents the lowest parts count for the complete system. Figure 4.4.28 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 4.4.29 shows the relationship between signal source impedance and gain or input impedance for the amplifier stage $Q_1 Q_2$. 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 4.4.28). 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 kHz or 21 to 24 dB for 4W.

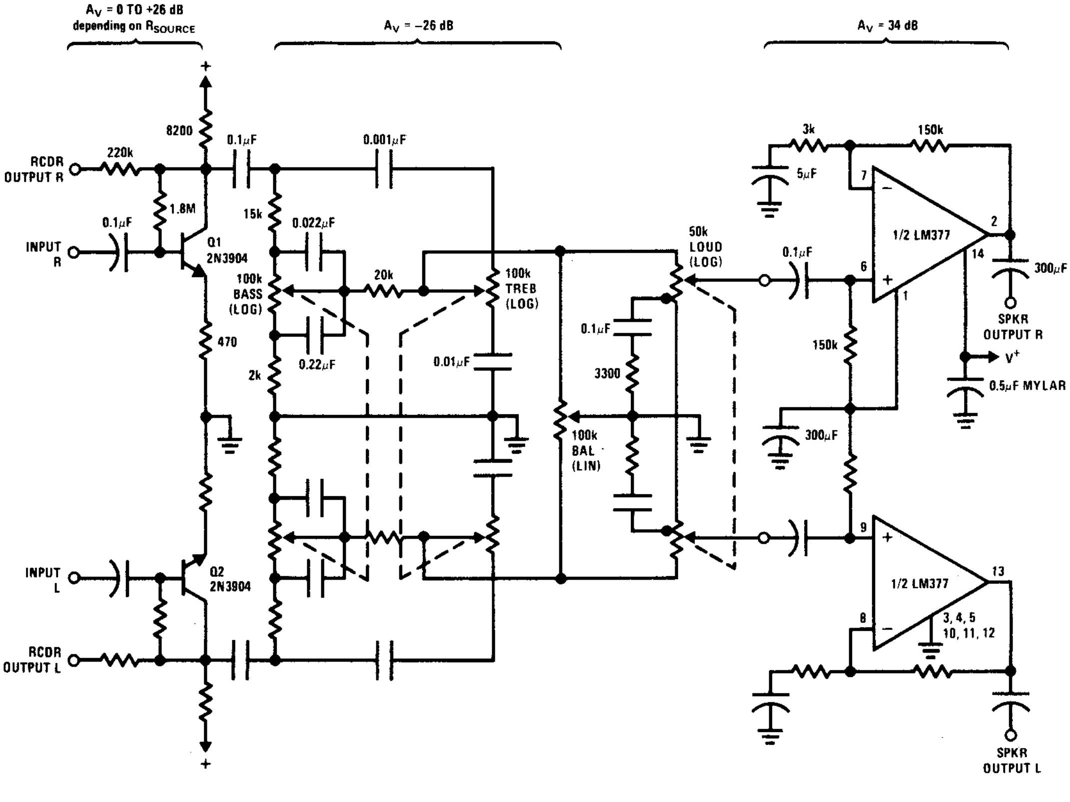


FIGURE 4.4.27 Two-Channel Power Amplifier and Control Circuits

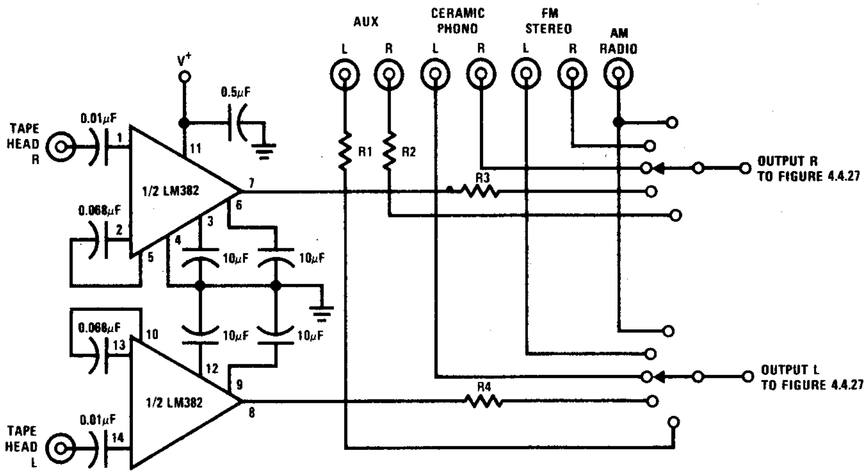


FIGURE 4.4.28 Two-Channel Tape-Playback Amplifier and Signal Switching

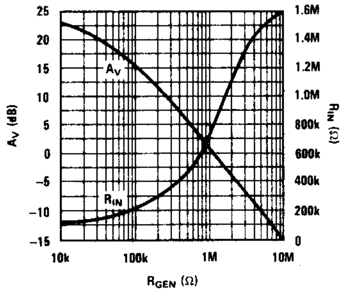


FIGURE 4.4.29 A_V and R_{IN} for Input Stage of Figure 4.4.26

4.4.17 Rear Channel Ambience Amplifier

The rear channel "ambience" circuit of Figure 4.4.30 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 LM377 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 4.4.30.

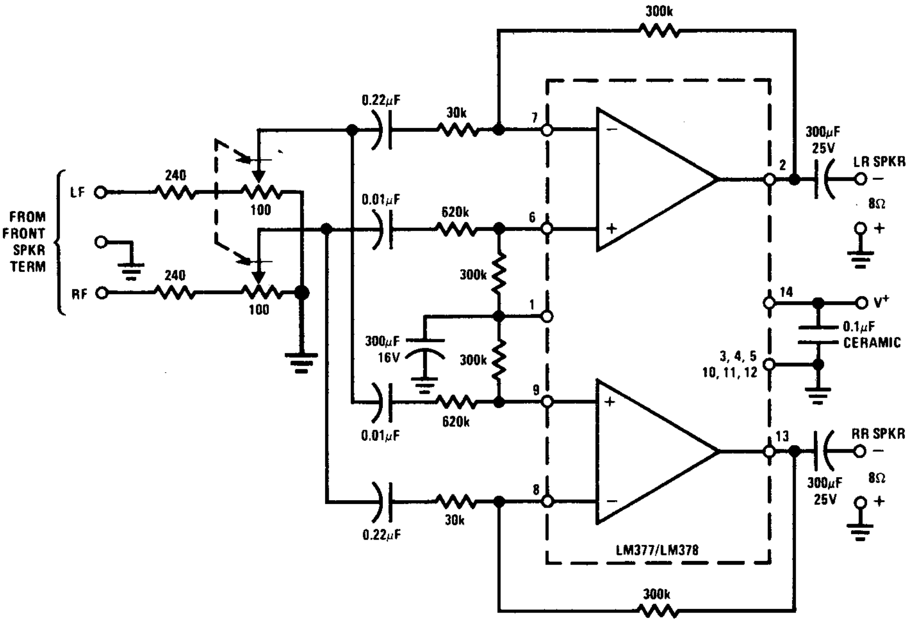


FIGURE 4.4.30 Rear Speaker Ambience (4-Channel) Amplifier

4.5 LM380 AUDIO POWER AMPLIFIER

4.5.1 Introduction

All of the mono power amplifiers listed in Table 4.3.2 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.

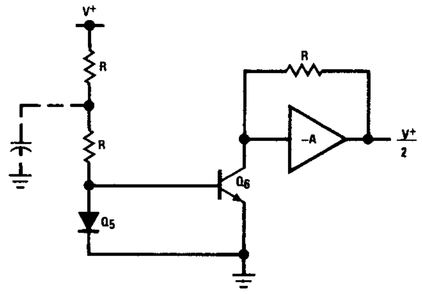


FIGURE 4.5.2 LM380 DC Equivalent Circuit

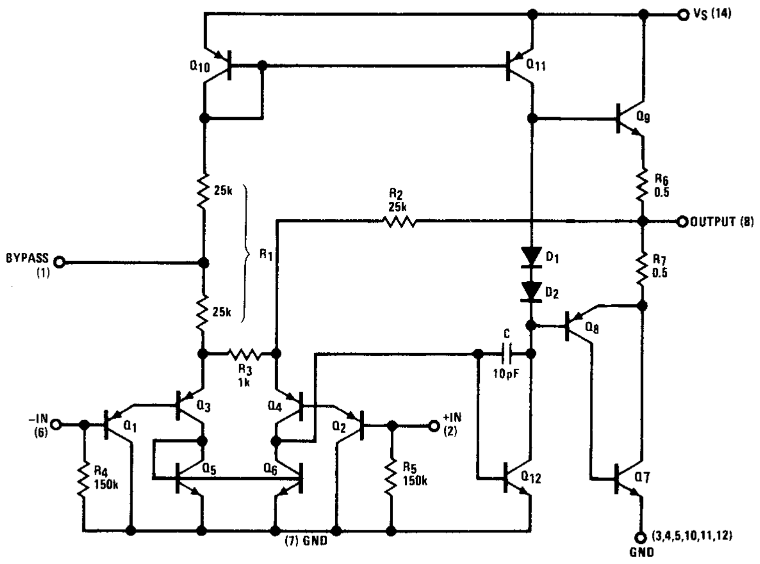


FIGURE 4.5.1 LM380 Simplified Schematic

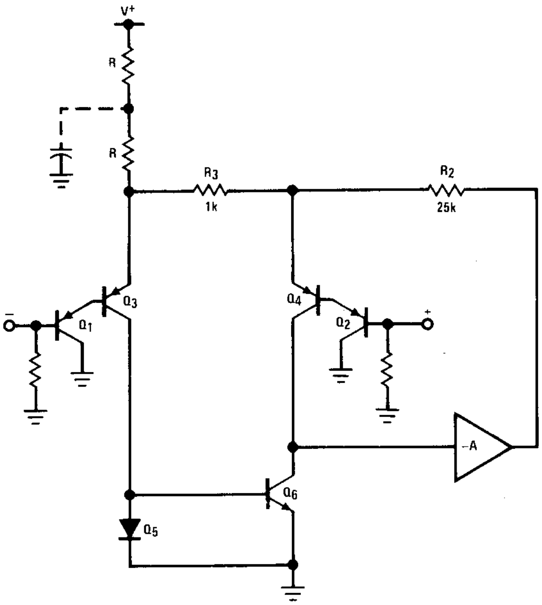


FIGURE 4.5.3 LM380 AC Equivalent Circuit

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.

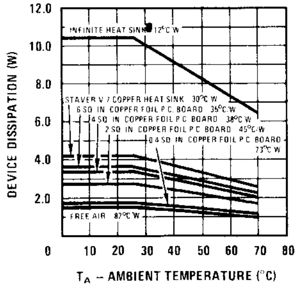


FIGURE 4.5.4a Device Dissipation vs. Maximum Ambient Temperature

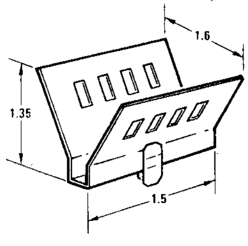
4.5.3 General Operating Characteristics

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

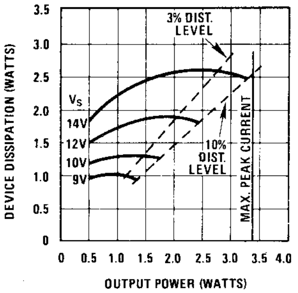


FIGURE 4.5.5a Device Dissipation vs. Output Power – 4Ω Load

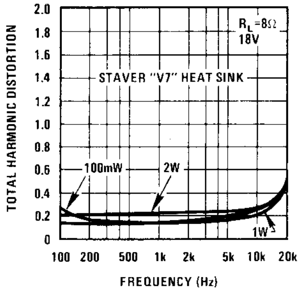


FIGURE 4.5.6 Total Harmonic Distortion vs. Frequency

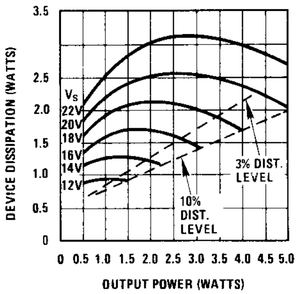


FIGURE 4.5.5b Device Dissipation vs. Output Power – 8Ω Load

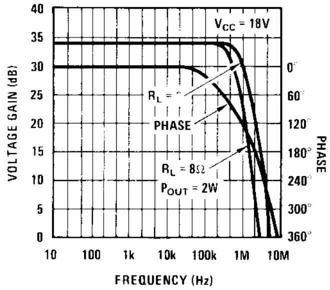


FIGURE 4.5.7 Output Voltage Gain vs. Frequency

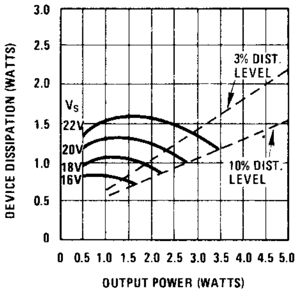


FIGURE 4.5.5c Device Dissipation vs. Output Power – 16Ω Load

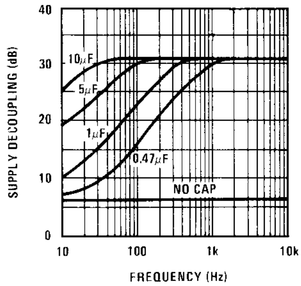


FIGURE 4.5.8 Supply Decoupling vs. Frequency

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 0.1μF disc ceramic capacitor at the V_S terminal of the IC.

The R_C and C_C shown as dotted line components on figure 4.5.9 and throughout this section suppresses 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.

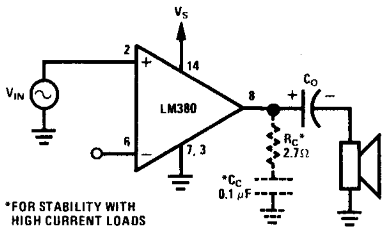


FIGURE 4.5.9 Oscillation Suppression Components

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.

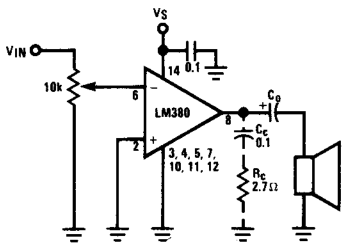


FIGURE 4.5.10 Minimum Component Configuration

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.

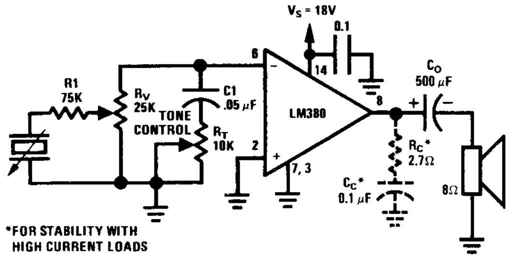


FIGURE 4.5.11 Ceramic Phono Amp

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.

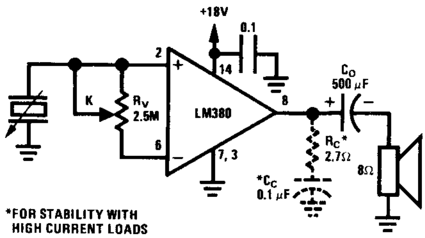


FIGURE 4.5.12 "Common Mode" Volume Control

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.

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

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

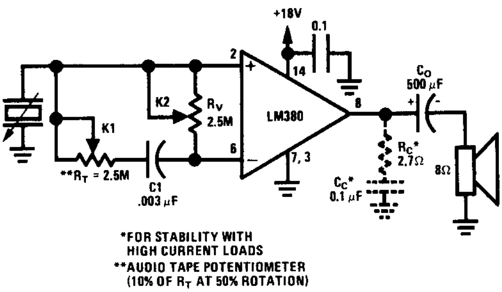


FIGURE 4.5.13 "Common Mode" Volume and Tone Control

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:

$$\frac{V_{OUT}}{V_{IN}} = 50\left(1 - \frac{150k}{150k + \dfrac{k_1 R_T k_2 R_V + \dfrac{k_2 R_V}{j 2 \pi f C_1}}{k_1 R_T + k_2 R_V + \dfrac{1}{j 2 \pi f C_1}}}\right)$$

$$0 \leqslant K_1 \leqslant 1$$
$$0 \leqslant K_2 \leqslant 1 \qquad (4.5.2)$$

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

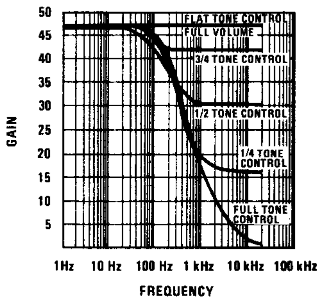
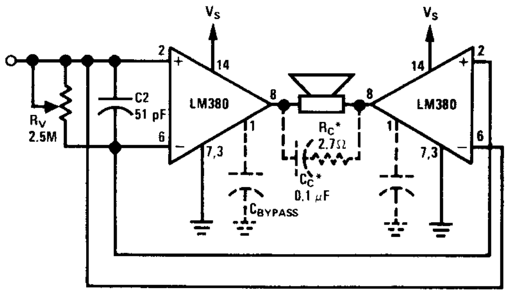


FIGURE 4.5.14 Tone Control Response

4.5.10 Bridge Amplifier

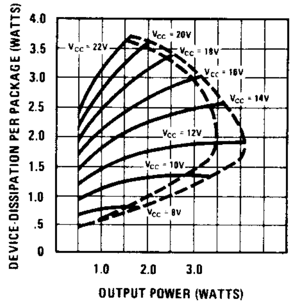
Where more power is desired than can be provided with one amplifier, two amps may be used in the bridge configuration shown in Figure 4.5.15.



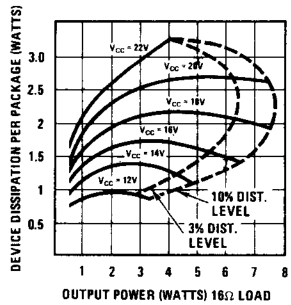
*FOR STABILITY WITH
HIGH CURRENT LOADS

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

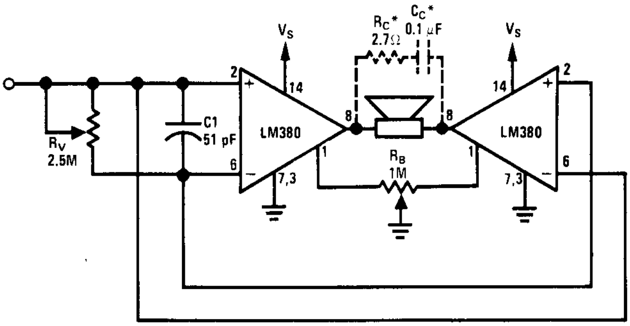
FIGURE 4.5.16

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.

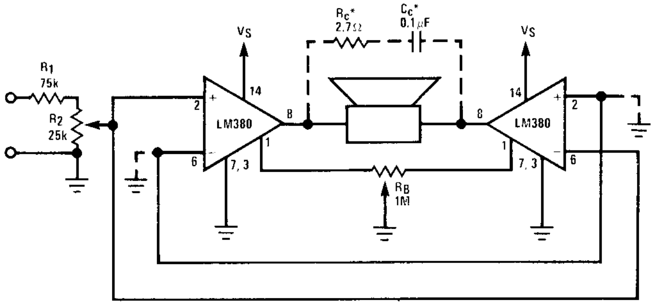
The circuits of Figures 4.5.15 and 4.5.17 employ the "common mode" volume control as shown before. However, any of the various input connection schemes discussed previously may be used. Figure 4.5.18 shows the bridge configuration with the voltage divider input. As discussed in the "Biasing" section 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.

4



*FOR STABILITY WITH HIGH CURRENT LOADS

FIGURE 4.5.17 Quiescent Balance Control



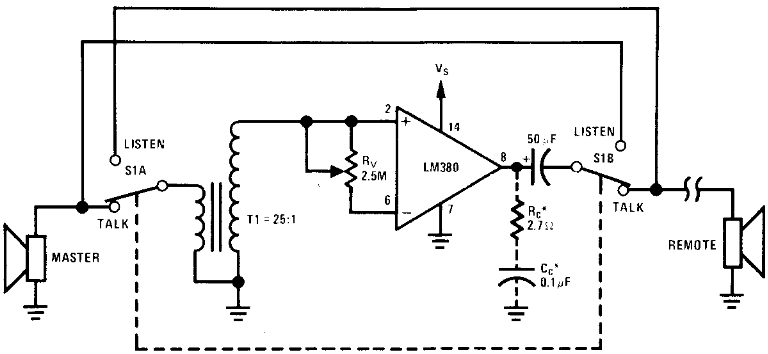
*FOR STABILITY WITH HIGH CURRENT LOADS

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.



*FOR STABILITY WITH HIGH CURRENT LOADS

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.

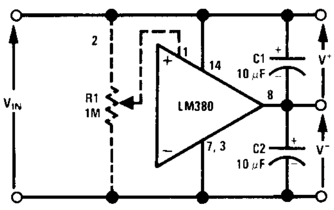


FIGURE 4.5.20 Dual Supply

With a 20 V input voltage (±10 V output) the circuit exhibits a change in output voltage of approximately 2% per 100 mA 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 1.3 A 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 22 MΩ for low frequency input signals. The gate to drain capacitance (2 pF maximum for the KE4221 shown) of the FET limits the input impedance as frequency increases.

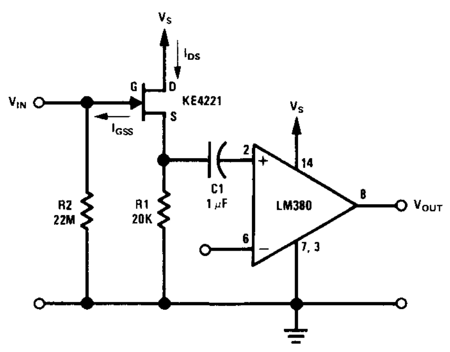


FIGURE 4.5.21 High Input Impedance

At 20 kHz the reactance of this capacitor is approximately −j4 MΩ, giving a net input impedance magnitude of 3.9 MΩ. 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 KE4221.

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 1 A over variable loads using the circuit shown in Figure 4.5.22.

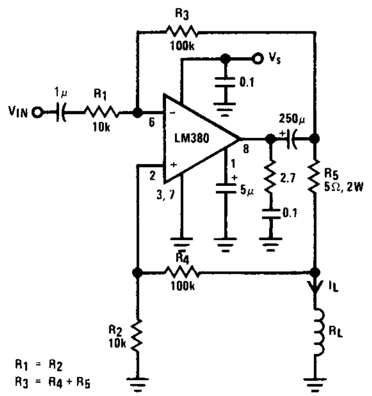


FIGURE 4.5.22 Power Voltage-to-Current Converter

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 250 mV_{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.

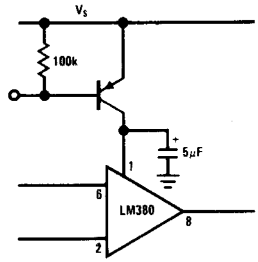


FIGURE 4.5.23 Muting the LM380

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.

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.

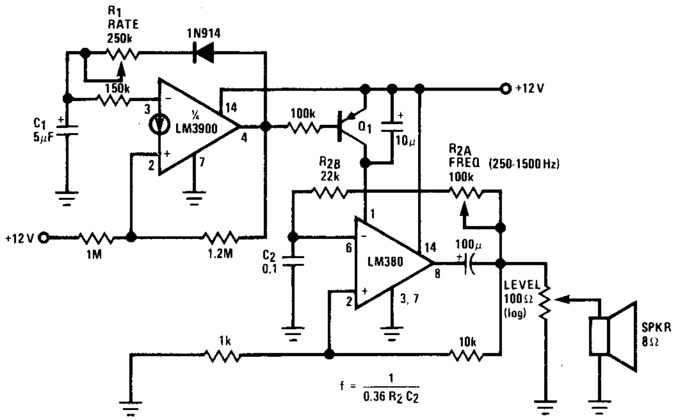


FIGURE 4.5.24 Siren with Programmable Frequency and Rate Adjustment

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.

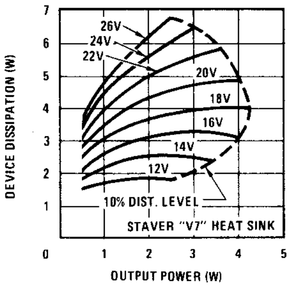


FIGURE 4.6.1 Device Dissipation vs. Output Power — 4Ω Load

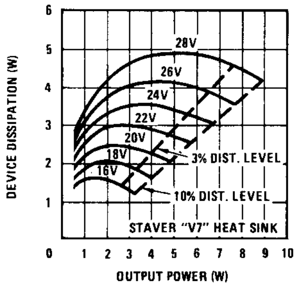


FIGURE 4.6.2 Device Dissipation vs. Output Power — 8Ω Load

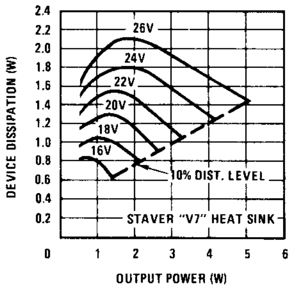


FIGURE 4.6.3 Device Dissipation vs. Output Power — 16Ω Load

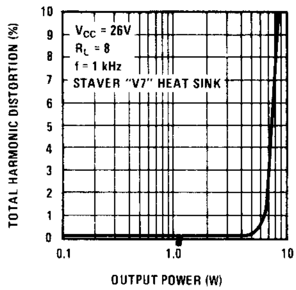


FIGURE 4.6.4 Total Harmonic Distortion vs. Output Power

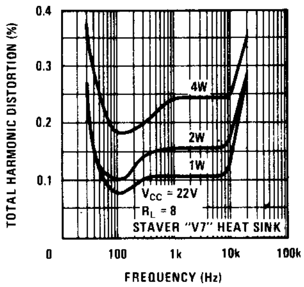


FIGURE 4.6.5 Total Harmonic Distortion vs. Frequency

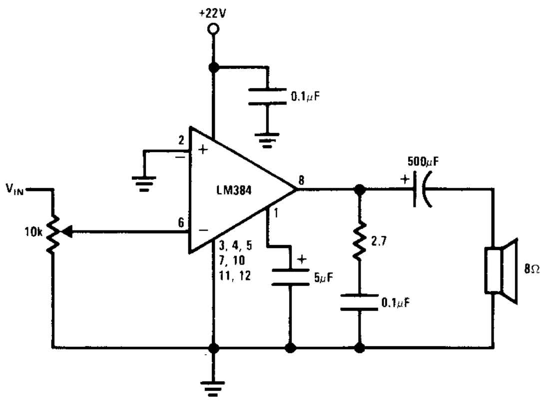


FIGURE 4.6.6 Typical 5W Amplifier

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.

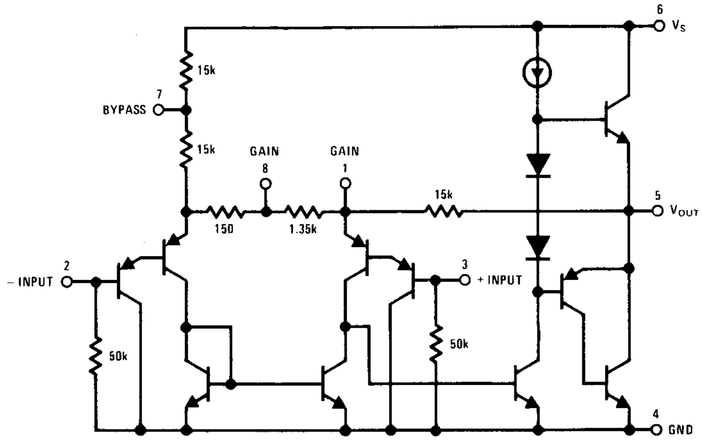


FIGURE 4.7.1 LM386 Simplified Schematic

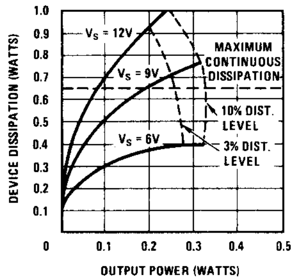


FIGURE 4.7.2 Device Dissipation vs. Output Power — 4Ω Load

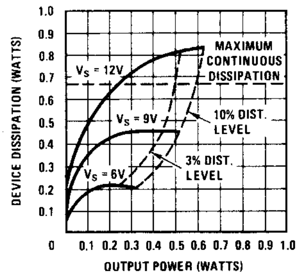


FIGURE 4.7.3 Device Dissipation vs. Output Power — 8Ω Load

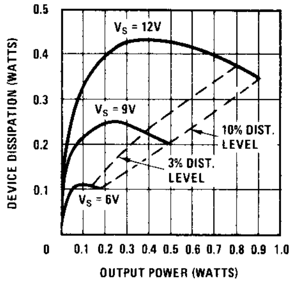


FIGURE 4.7.4 Device Dissipation vs. Output Power — 16Ω Load

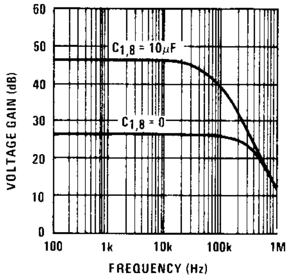


FIGURE 4.7.5 Voltage Gain vs. Frequency

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.

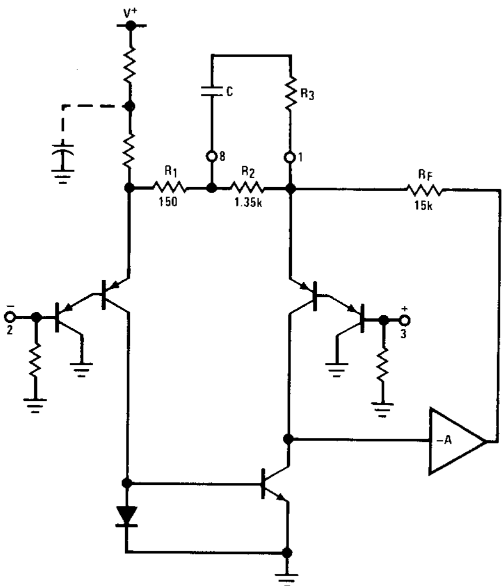


FIGURE 4.7.6 LM386 AC Equivalent Circuit

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 Typical Applications

Three possible variations of the LM386 as a standard audio power amplifier appear as Figures 4.7.7-4.7.9. Possible gains of 20, 50 and 200V/V are shown as examples of various gain control methods. The addition of the optional 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).

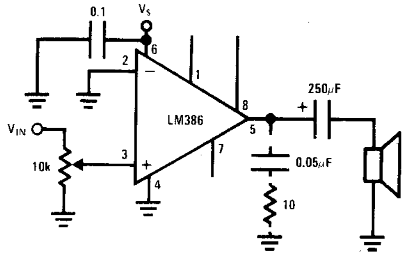


FIGURE 4.7.7 Amplifier with Gain = 20V/V (26dB) Minimum Parts

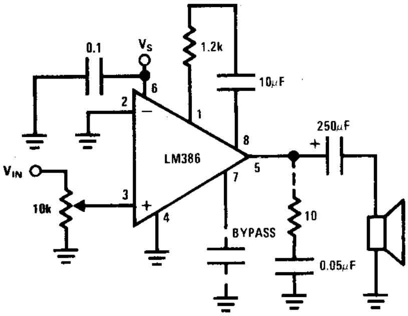


FIGURE 4.7.8 Amplifier with Gain = 50V/V (34dB)

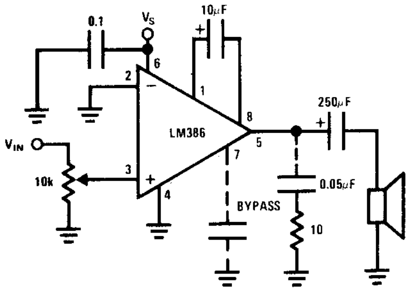


FIGURE 4.7.9 Amplifier with Gain = 200V/V (46dB)

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 4.7.11. 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 between pins 1 and 8, thus saving one capacitor.

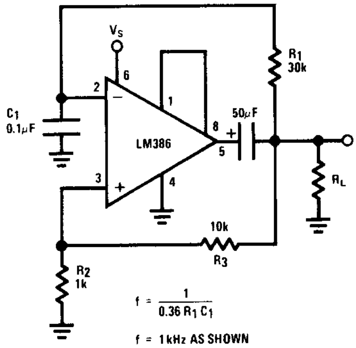


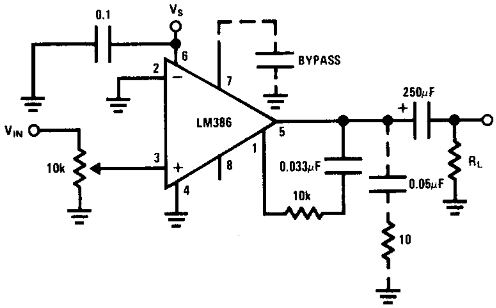
FIGURE 4.7.11 Square Wave Oscillator

4.7.7 Bass Boost Circuit

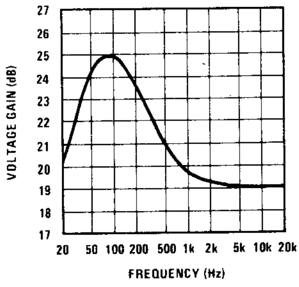
Additional external components can be placed in parallel with the internal feedback resistors (Figure 4.7.10) to tailor the gain and frequency response for individual applications. For example, we can compensate poor speaker bass response by frequency shaping the feedback path. This is done with a series 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 only compensated for closed-loop gains greater than 9.

4.7.9 Power Wien Bridge Oscillator

The LM386 makes a low cost, low distortion audio frequency oscillator when wired into a Wien bridge configuration (Figure 4.7.12). 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.



(a) Amplifier with Bass Boost



(b) Frequency Response with Bass Boost

FIGURE 4.7.10 LM386 with Bass Boost

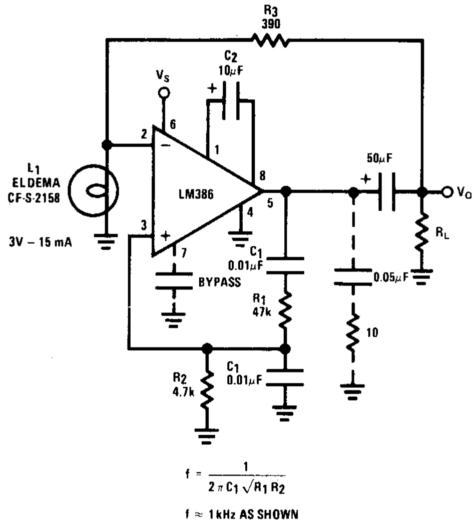


FIGURE 4.7.12 Low Distortion Power Wien Bridge Oscillator

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

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 0.1μF 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 1μA to 5mA. These transistors make good LED driver devices; V_{SAT} is only 150mV when sinking 10mA.

4

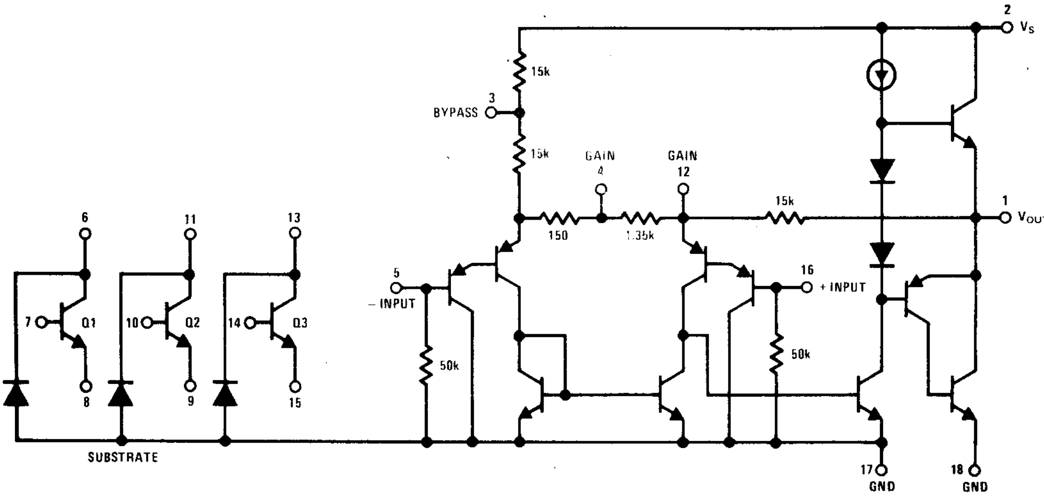


FIGURE 4.8.1 LM389 Simplified Schematic

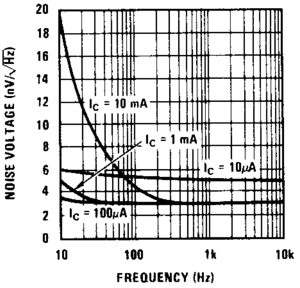


FIGURE 4.8.2 Noise Voltage vs. Frequency

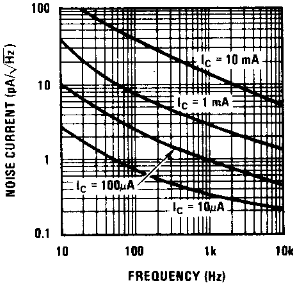


FIGURE 4.8.3 Noise Current vs. Frequency

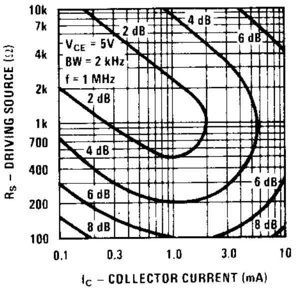


FIGURE 4.8.4 Contours of Constant Noise Figure

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.

Several applications of the LM389 follow as examples of practical circuits and also as idea joggers.

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.

4.8.7 Ceramic Phono Amplifier with Tone Controls

For proper frequency response (particularly at the low end), ceramic cartridges require a high termination impedance. Figure 4.8.7 shows a low-cost single IC phono amplifier where one of the LM389 transistors is used as a high input impedance emitter follower to provide the required cartridge load. The remaining transistors form a high-gain Darlington pair, used as the active element in a low distortion Baxandall tone control circuit (see Section 2.14.7).

4.8.8 Siren

The siren circuit of Figure 4.8.8 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}.

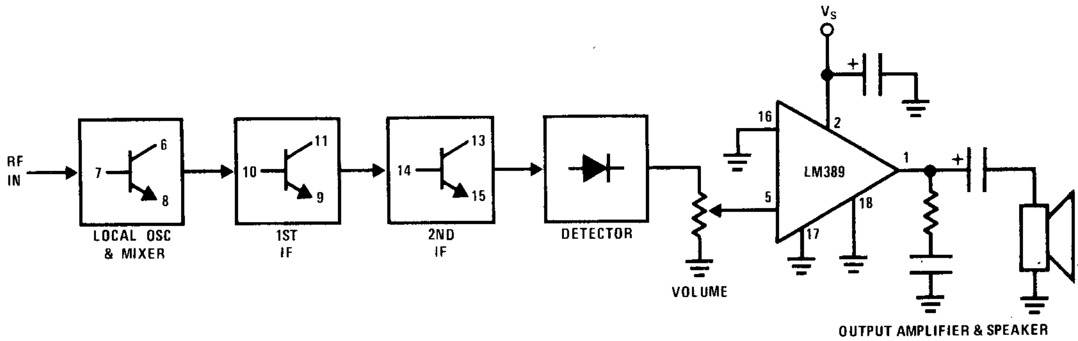


FIGURE 4.8.5 AM Radio

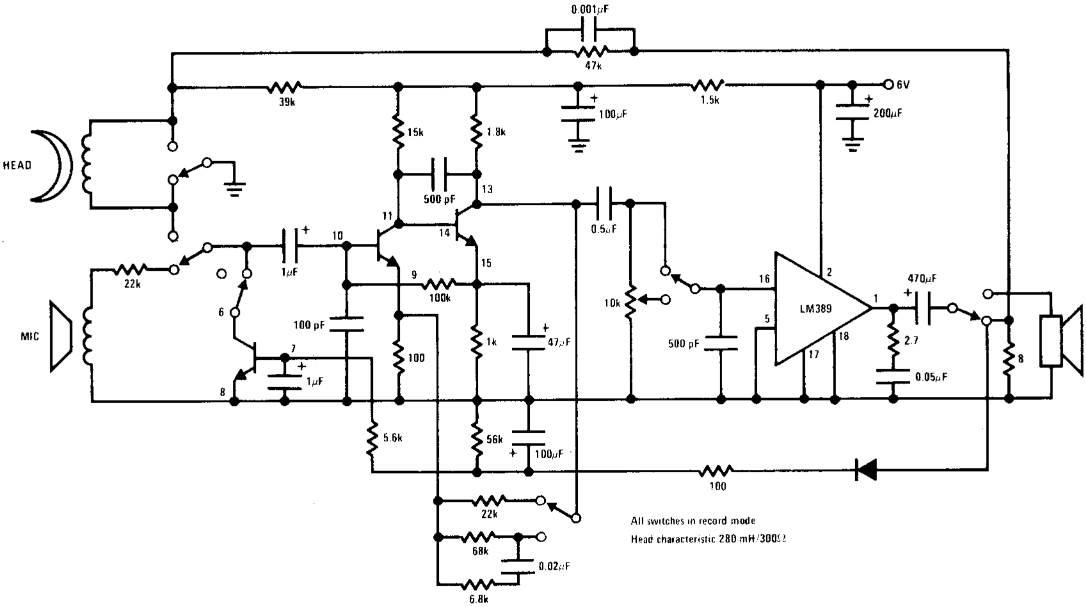


FIGURE 4.8.6 Tape Recorder

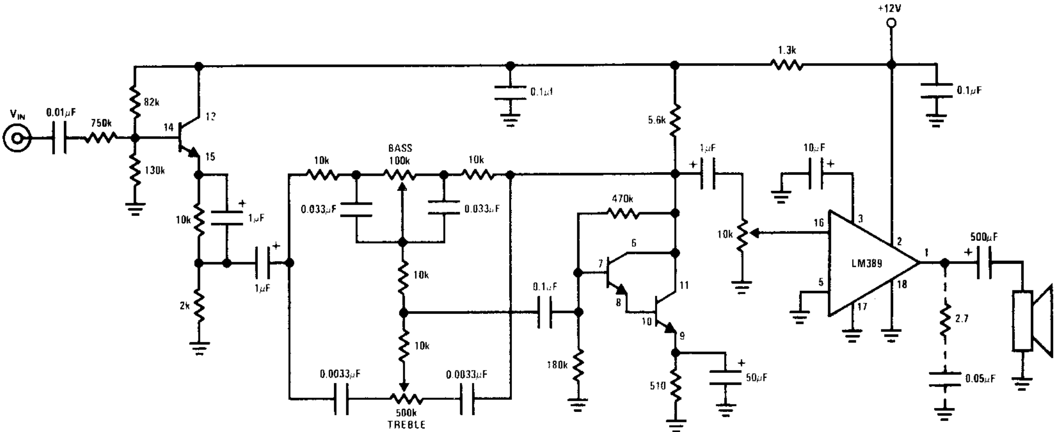


FIGURE 4.8.7 Ceramic Phono Amplifier with Tone Controls

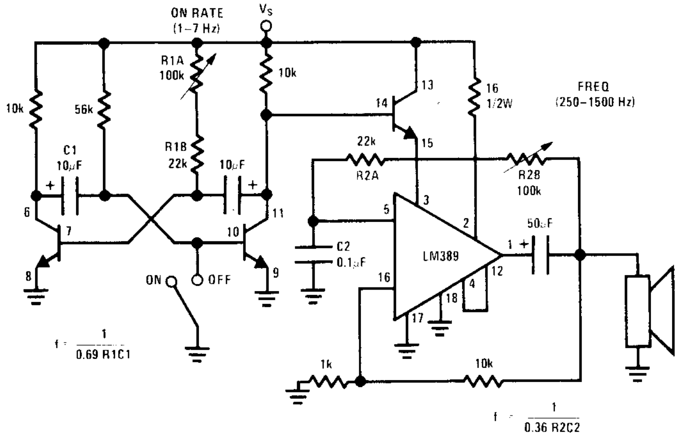


FIGURE 4.8.8 Siren

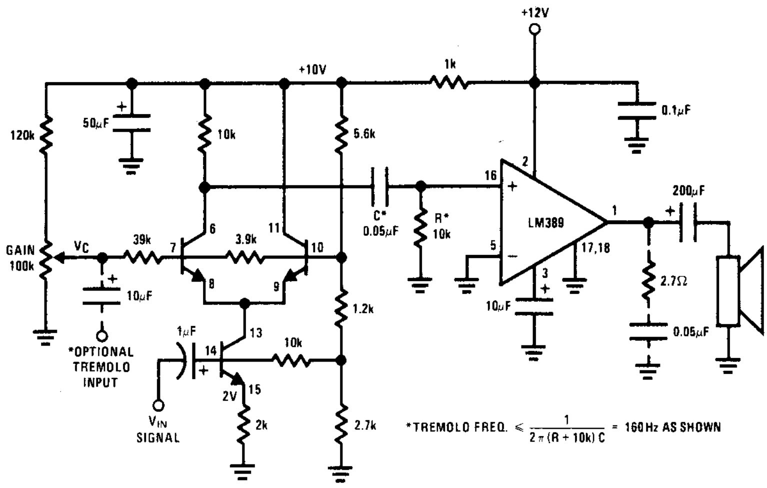


FIGURE 4.8.9 Voltage-Controlled Amplifier or Tremolo Circuit

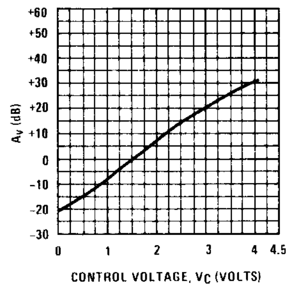


FIGURE 4.8.10 VCA Gain vs. Control Voltage

4.8.9 Voltage-Controlled Amplifier or Tremolo Circuit

A voltage-controlled amplifier constructed from the LM389 appears as Figure 4.8.9. 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 1mA 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 4.8.10). 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 4.8.11. 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.

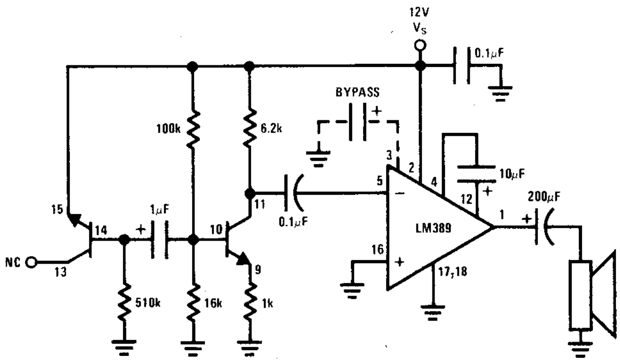


FIGURE 4.8.11 Noise Generator Using Zener Diode

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 4.8.12-4.8.14 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

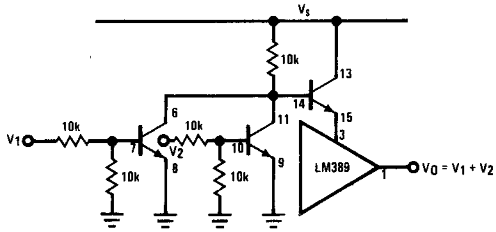


FIGURE 4.8.13 OR Muting

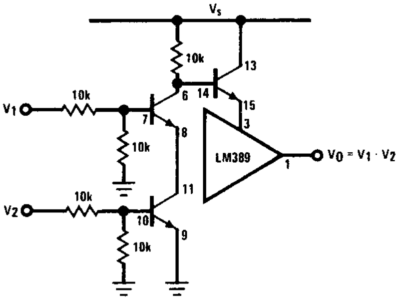


FIGURE 4.8.12 AND Muting

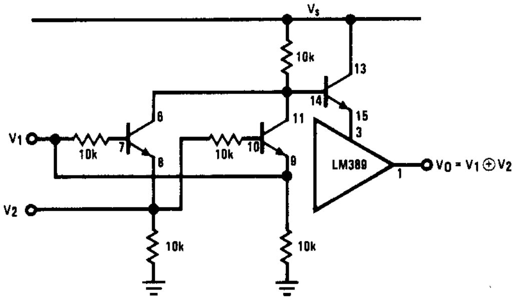


FIGURE 4.8.14 Exclusive-OR Muting

4.9 LM388 BOOTSTRAPPED AUDIO POWER AMPLIFIER

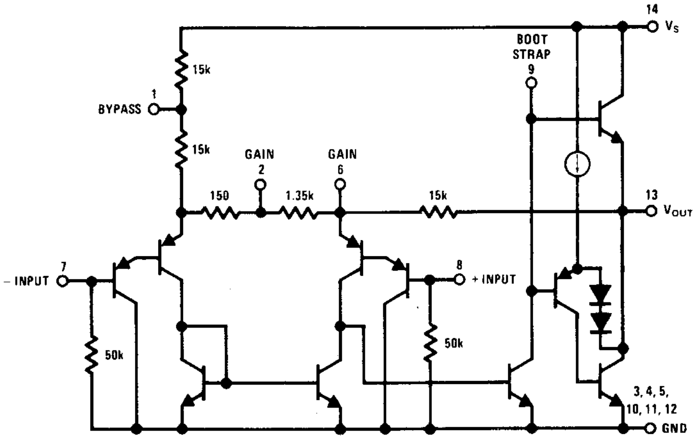


FIGURE 4.9.1 LM388 Simplified Schematic

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 boot-strapped 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.

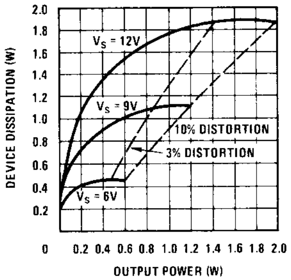


FIGURE 4.9.2 Device Dissipation vs. Output Power — 4 Ω Load

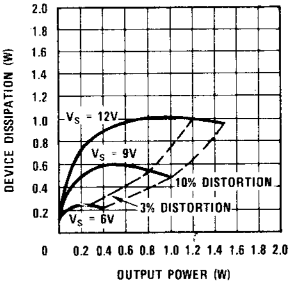


FIGURE 4.9.3 Device Dissipation vs. Output Power — 8 Ω Load

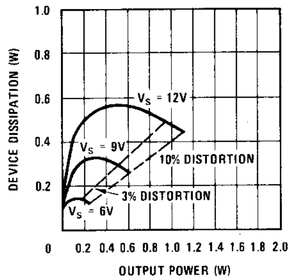


FIGURE 4.9.4 Device Dissipation vs. Output Power — 16 Ω Load

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.

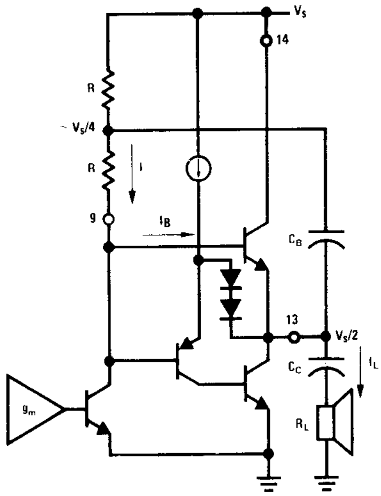


FIGURE 4.9.5 LM388 Output Stage

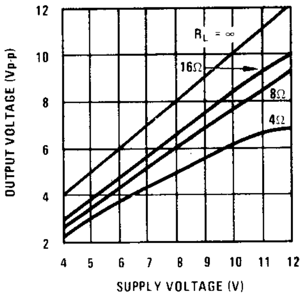


FIGURE 4.9.6 Peak-to-Peak Output Voltage Swing vs. Supply Voltage

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:

$$\frac{R}{2} C_B = R_L C_C \qquad (4.9.2)$$

Substituting Equation (4.9.1) into (4.9.2) yields:

$$C_B = \frac{4\,C_C}{\beta} \qquad (4.9.3)$$

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

$$R = 50\,R_L \qquad (4.9.4)$$

$$C_B = \frac{C_C}{25} \qquad (4.9.5)$$

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.

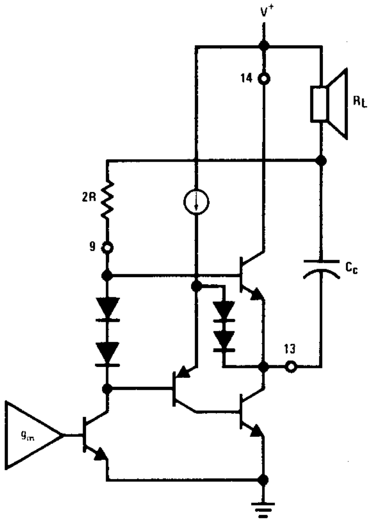


FIGURE 4.9.7 Bootstrapping with Load to Supply

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.

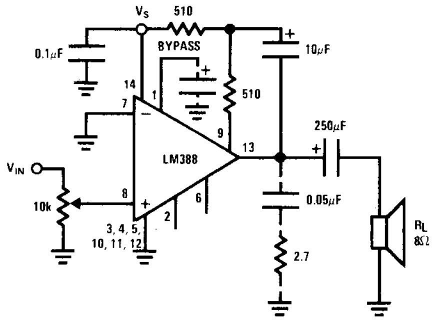


FIGURE 4.9.8 Load Returned to Ground (Amplifier with Gain = 20)

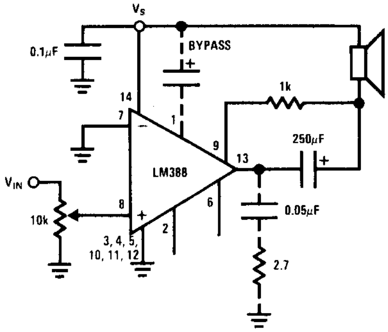


FIGURE 4.9.9 Load Returned to V_S (Amplifier with Gain = 20)

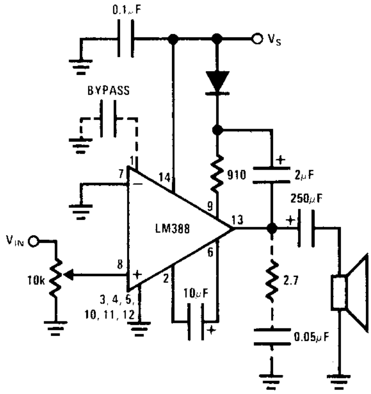
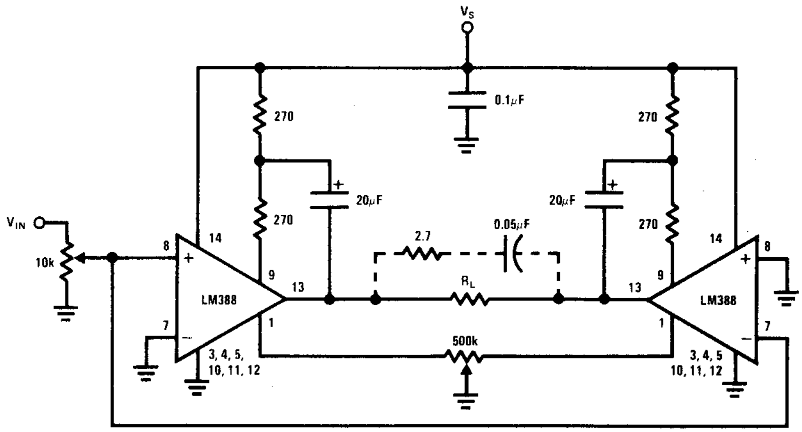


FIGURE 4.9.10 Amplifier with Gain = 200 and Minimum C_B

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.



$V_S = 6V$ $R_L = 4\Omega$ $P_O = 1.0W$
$V_S = 12V$ $R_L = 8\Omega$ $P_O = 3.5W$

FIGURE 4.9.11 Bridge Amp

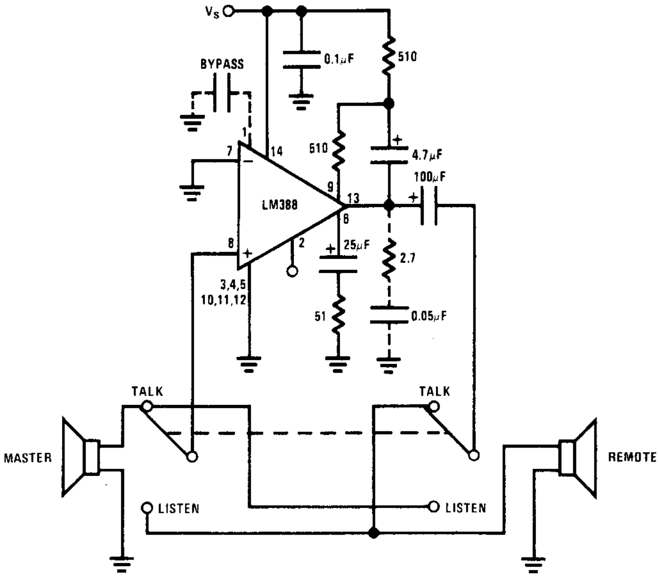


FIGURE 4.9.12 Intercom

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 optional 2.7Ω-$0.05\mu F$ 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.

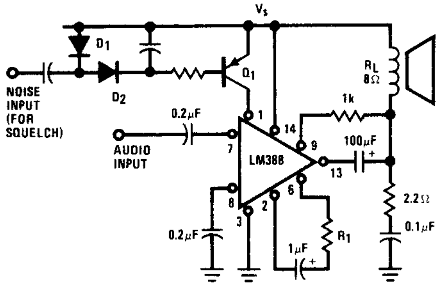


FIGURE 4.9.13 LM388 Squelch Circuit for FM Scanners and Walkie Talkies

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 = 8Ω, THD = 5%)
- P_O = 0.19W (V_S = 4.5V, R_L = 8Ω, 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.

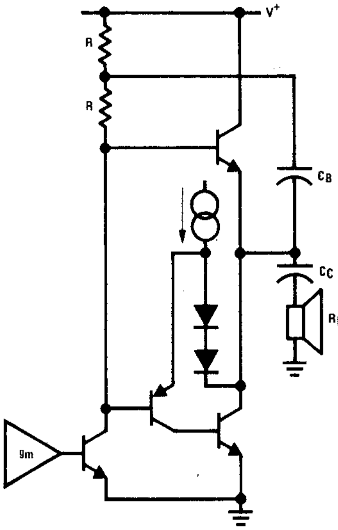


FIGURE 4.10.2 LM390 Output Stage

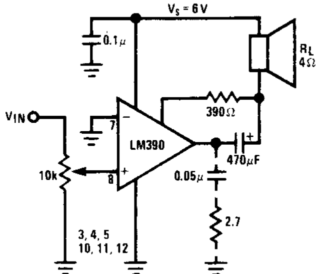


FIGURE 4.10.3 1 Watt Power Amplifier for 6 Volt Systems

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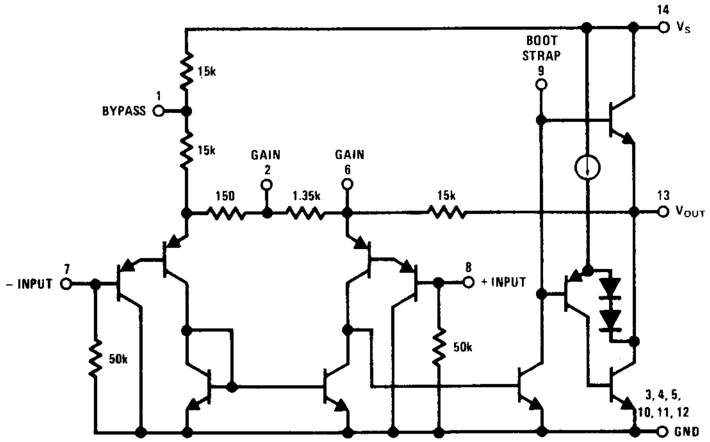


FIGURE 4.10.1 LM390 Simplified Schematic

4.11 BOOSTED POWER AMPLIFIERS

4.11.1 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.

4.11.2 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.

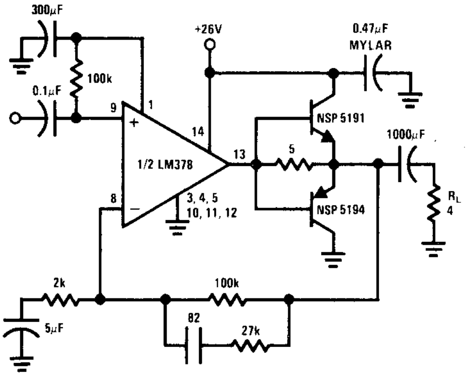


FIGURE 4.11.1 10 Watt Power Amplifier

The response of the 10W boosted amplifier is indicated in Figure 4.11.2 for power levels below clipping. Distortion is below 2% from about 50 Hz to 30 kHz. 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 4.11.3 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 4.11.4 and 4.11.5 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

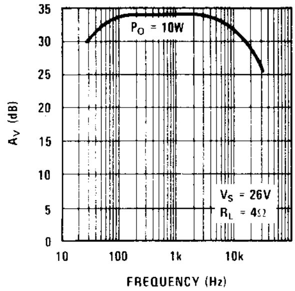


FIGURE 4.11.2 10 Watt Boosted Amplifier, Frequency Response

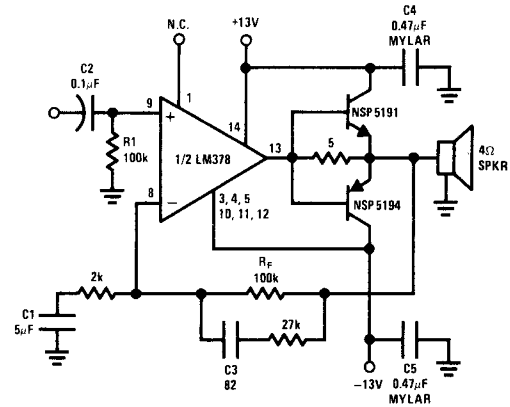


FIGURE 4.11.3 12 Watt Low-Distortion Power Amplifier

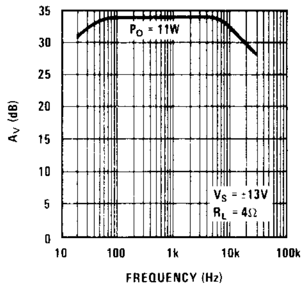


FIGURE 4.11.4 Response for Amplifier of Figure 4.11.3

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

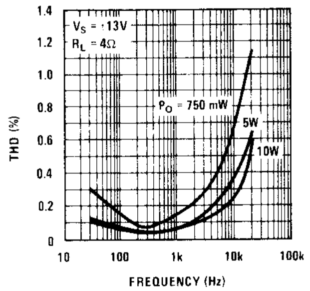


FIGURE 4.11.5 Distortion for Amplifier of Figure 4.11.3

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.

4.11.3 LM391 Power Driver

Coming in late 1976 will be National's LM391 power driver IC designed to provide complementary output drive for external transistors. Power amplifiers up to 50W will be possible with complete SOA protection provided on-chip, allowing for simple, low parts-count designs. User gain control, set externally, offers maximum flexibility, while special internal techniques allow for the high supply voltages required by high power amplifiers, thus eliminating the expense and inconvenience of two power supplies. Optimized for the top-of-the-line medium power amplifiers, the LM391 promises to simplify and cut costs of these designs while retaining true high quality performance.

For high power, battery operated audio products, work is being finalized on a new low voltage driver IC designed to complement the LM391 in operation and performance, but optimized for 6-12V, 2Ω designs. Scheduled for introduction in early 1977, this IC will greatly reduce the cost and difficulties of obtaining the high output swing and large currents demanded.

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:

$$P_O = \frac{V_O^2}{R_L} = I_O^2 R_L \tag{4.12.1}$$

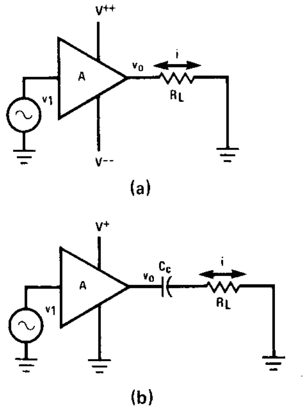


FIGURE 4.12.1 Simple Audio Circuits

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:

$$P_O = \frac{V_{OPP}^2}{8 R_L} = \frac{R_L \, I_{OPP}^2}{8} \tag{4.12.2}$$

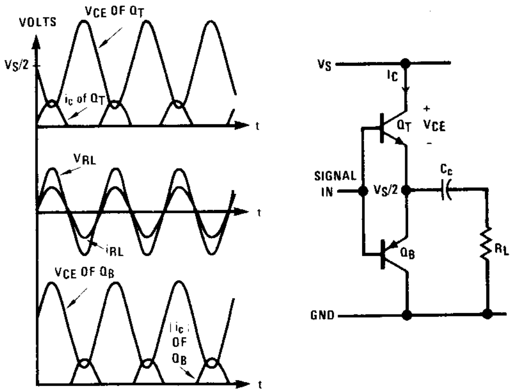


FIGURE 4.12.2 Class B Waveforms

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:

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

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

$$P_D = \frac{1}{2\pi}\,(2)\int_0^{\pi} p_d\, d(\omega t)$$

two transistors operated Class B (since both transistors are in the same IC package)

where: P_D = average power

p_d = instantaneous power

then

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

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

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

$$= \frac{V_s V_L}{\pi R_L} - \frac{V_L^2}{2 R_L} \tag{4.12.4}$$

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:

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

$$\therefore\ V_{Lp} = \frac{V_s}{\pi} \tag{4.12.5}$$

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:

$$V_{Lp\text{-}p} = \frac{2 V_s}{\pi} = 0.637\, V_s \tag{4.12.6}$$

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

$$\max P_D = \frac{V_s^2}{2\pi^2 R_L} \approx \frac{V_s^2}{20 R_L} \tag{4.12.7}$$

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

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

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.

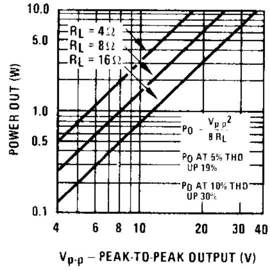


FIGURE 4.12.3 Power Out

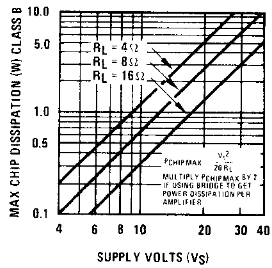
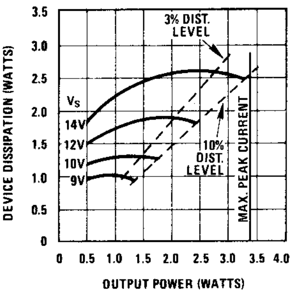


FIGURE 4.12.4 Max Chip Dissipation

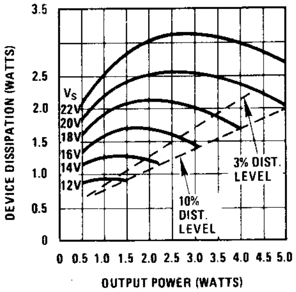
The nomographs of Figures 4.12.3 and 4.12.4 make it easy to determine package power dissipation as well as output VI characteristics for popular conditions. Since part of the audio amplifier specmanship game involves juggling output power ratings given at differing distortion levels, it is useful to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

P_O increases by 19% at 5% THD

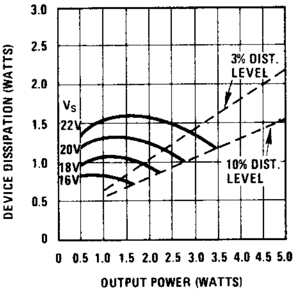
P_O increases by 30% at 10% THD



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

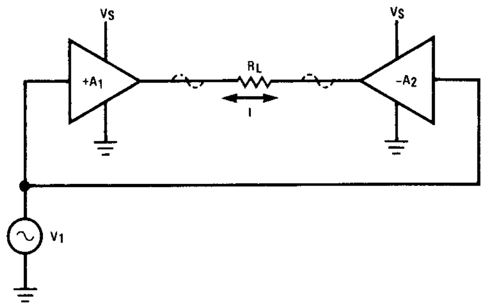


FIGURE 4.12.6 Bridge Audio

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 $0.637 V_S$ to the load, and then is driven harder, say to $0.8 V_S$, 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 single. 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:

$$P_{A1 \text{ or } A2} = \frac{V_S^2}{\pi^2 R_L} = \frac{V_S^2}{10 R_L} \qquad (4.12.8)$$

4.13 EFFECT OF SPEAKER LOADS

The power dissipation results found in the previous section assumed a purely resistive load; however, real-world speakers are anything *but* resistive. Figure 4.13.1 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 degrees. This means that the voltage through the speaker leads the current by 60 degrees.

Abandoning mathematical rigor for a more 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 degrees phase angle. This becomes clear by considering the resistive case of zero phase angle depicted in Figure 4.13.2a, 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 dissipation 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 degree phase

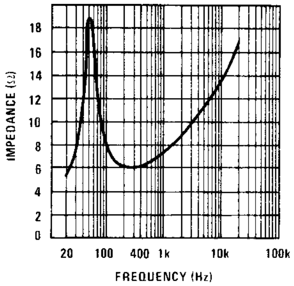


FIGURE 4.13.1 Impedance Curve for a Typical Dynamic Loudspeaker

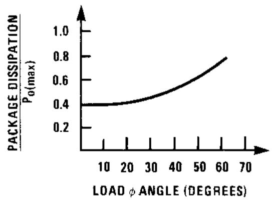


FIGURE 4.13.3 Class B Package Dissipation for Reactive Loads

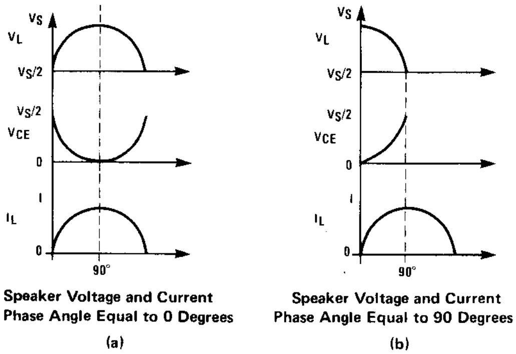


FIGURE 4.13.2 Phase Angle Relationship Between Voltage and Current

difference Figure 4.13.2b results, where there exists zero volts across the load, maximum voltage across the package, and maximum current flowing through both, producing maximum package dissipation.

Returning to mathematics for a moment to derive a new expression containing phase angle and plotting the results produces the curve shown in Figure 4.13.3. The importance of Figure 4.13.3 is seen by comparing the power ratio at zero degrees (0.405) with that at 60 degrees (0.812) – double! This means that *the maximum package dissipation can be twice as much for a loudspeaker load as for a resistive load*. What softens this hard piece of reality is the relative rarity and short duration of amplifiers running at (or near) maximum power output; also, most heat sinks have adequate thermal capacity to ride through these peaks. In any event, phase angle is real and it does increase power dissipation and needs to be considered in heat sink design.

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.

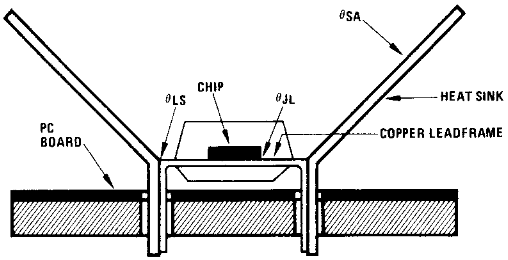
If testing a breadboarded power IC results in premature waveform clipping, or a "truncated shape," or a "melting down" of the positive peaks, the IC is probably in thermal shutdown and requires more heatsinking. The following information is provided to make proper heat sink selection easier and help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

4.14.1 Heat Flow

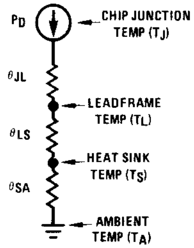
Heat can be transferred from the IC package by three methods, as described and characterized in Table 4.14.1.

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_{JC} and θ_{CS}. 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_{CA}. 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:

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

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.

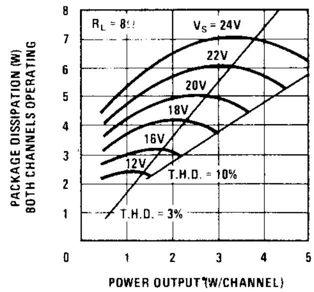


FIGURE 4.14.2 Power Dissipation vs. Power Output

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 (or four times Equation (4.14.2).*

θ_{LS}

Thermal resistance between lead frame and heatsink is a function of how close the bond can be made. The method recommended is use of 60/40 solder. When soldered, θ_{LS} may be neglected or a value of θ_{LS} = 0.25°C/W may be used.

$T_{J(max)}$

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.

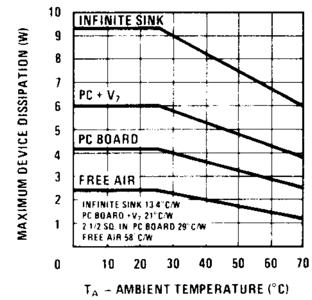


FIGURE 4.14.3 Maximum Dissipation vs. Ambient Temperature

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., $\theta_{SA} = \theta_{JA} - \theta_{JL} - \theta_{LS}$.

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:

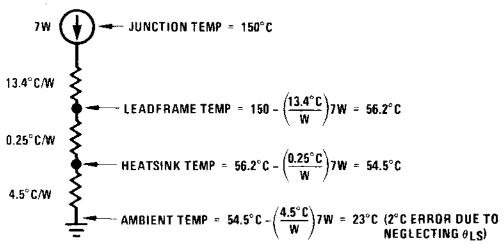


FIGURE 4.14.4 Heat Flow Model for LM378 Example

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 regulator 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.
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:

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

where: H = height of vertical plate in inches

η = fin effectiveness factor

h_C = convection heat transfer coefficient (4.14.6)

h_r = radiation heat transfer coefficient (4.14.7)

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

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

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

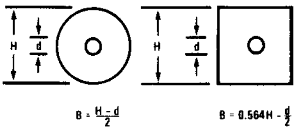
T_A = ambient temperature in °C

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



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 black anodized 1/16" thick aluminum to have a thermal resistance of 4°C/W. LM379 operating conditions are:

1. T_J = 150°C, T_A = 60°C, P_D = 9.5W, θ_{JL} = 6°C/W, neglect θ_{LS}.
2. T_S = 150°C - 6°C/W (9.5W) = 93°C.
3. 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).

4. $$h_C = 2.21 \times 10^{-3}\left(\frac{93 - 60}{3.5}\right)^{1/4}$$

$$= 3.86 \times 10^{-3}\,\text{W/}^\circ\text{C in}^2$$

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

4

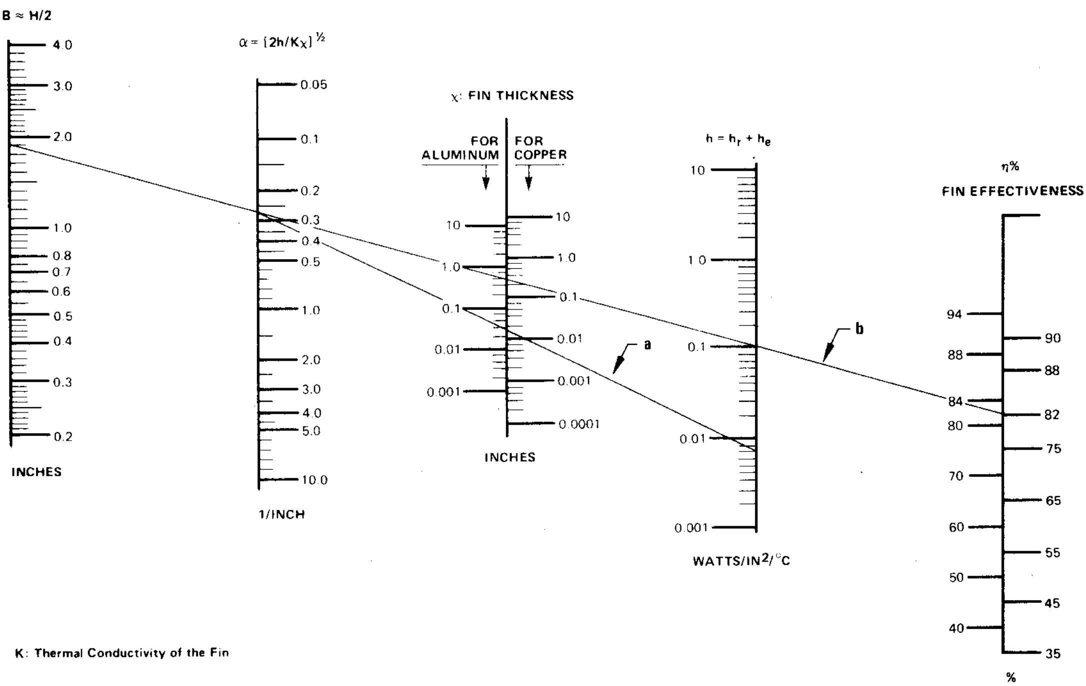


FIGURE 4.14.6 Fin Effectiveness Nomogram for Symmetrical, Flat, Uniformly-Thick, Vertically Mounted Fins

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

$h = h_c + h_r = 9.46 \times 10^{-3} W/^{\circ}C\ in^2$

5. $\eta = 0.84$ from figure 4.14.6.

6. $$\theta_{SA} = \frac{10^3}{2 \times 12.3 \times 0.84 \times 9.46} = 5.1^{\circ}C/W,$$

which is too large.

7. A larger fin is required, probably by about 40% in area. Accordingly, using a fin of 4.25" square, a new calculation is made.

4'. $$h_c = 2.21 \times 10^{-3}\left(\frac{33}{4.2}\right)^{1/4} = 3.7 \times 10^{-3}$$

$h_r = 5.6 \times 10^{-3}$ as before

$h = 9.3 \times 10^{-3}$

5'. $\eta = 0.75$ from Figure 4.14.6.

6. $$\theta_{SA} = \frac{10^3}{2 \times 18 \times 0.75 \times 9.3} = 3.98^{\circ}C/W,$$

which is satisfactory.

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.

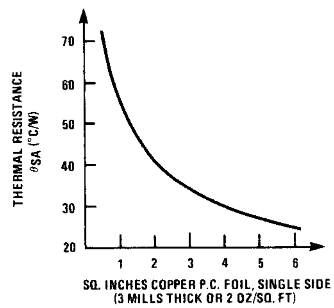


FIGURE 4.14.7 Thermal Resistance vs. Square Inches of Copper Foil