LM377, LM378 and LM379 Dual Two, Four and Six Watt Power Amplifiers

National Semiconductor
Application Note 125
Jim Sherwin
January 1975

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 shut-down protection, fast turn-on and turn-off without "pops" or pulses of active gain, an output which is self-entering at $V_{CC}/2$, and a 5 to 20 MHz 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.

CIRCUIT DESCRIPTION

The simplified schematic of *Figure 1* shows the important design features of the amplifier. The differential input stage made up of Q1–Q4 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 C1. (Unity gain bandwidth of an amplifier with pole-splitting compensation occurs where the first stage transconductance equals ω C1.) Further decrease of transconductance is provided by degeneration caused by resistors at Q2 and Q3 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 Q5 and Q6. 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 Q7 is actually preceeded by an emitter follower transistor not shown in the simplified circuit.

The second stage Q7 operates common-emitter with a current source load for high gain. Pole splitting compensation is provided by C1 to achieve unity gain bandwidth of about 10 MHz. Internal compensation is sufficient with closed-loop gain down to about $A_V = 25$.

The output stage is a complementary common-collector class AB composite. The upper, or current sourcing section, is a Darlington emitter follower Q12 and Q13. The lower, or current sinking, section is a composite PNP made up of Q14, Q15, and Q9. Normally, this type of PNP composite has low f_t and excessive delay caused by the lateral PNP transistor Q9. 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

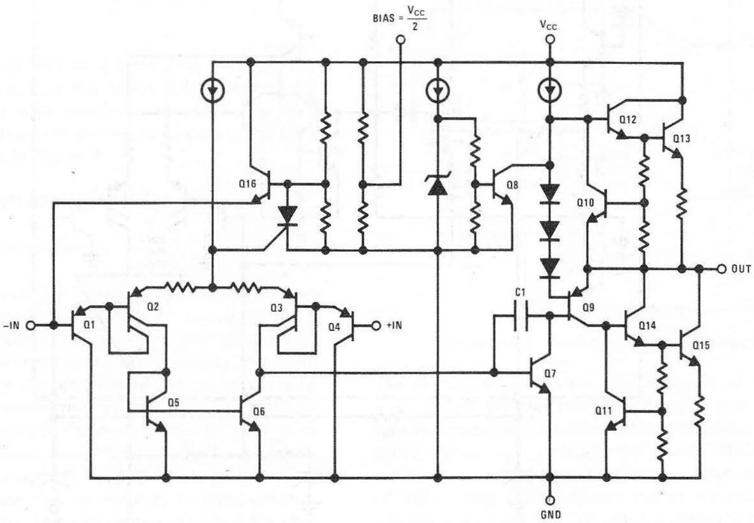


FIGURE 1. Simplified Schematic Diagram

RC network from output to ground to reduce loop gain of the composite PNP and so prevent the oscillation. In the LM377 series amplifiers, Q9 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. Q10 and Q11 provide output current limiting at 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, C1, in *Figure 2*.

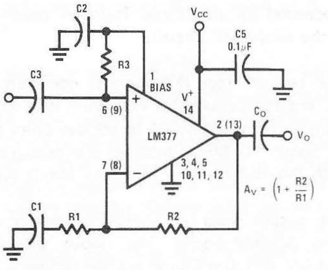


FIGURE 2. Non-Inverting Amplifier Connection

Establishing the initial charge on this capacitor results in a turn-on delay. An additional capacitor, C2, 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 3*). Where a supply rejection of 40 dB is required with 40 dB closed-loop gain, 80 dB ripple attenuation is required of R_S C2. The turn-on time can be calculated as follows:

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

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

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

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

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 1* as Q16 and an accompanying SCR; it is repeated and elaborated in *Figure 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 Q16, rapidly charging C1 and C2 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 Q3 and Q4 into balance. This, in turn, drives Q3 into conduction. Transistors Q2 and Q3 form an SCR latch which then triggers and clamps the base of Q16 to

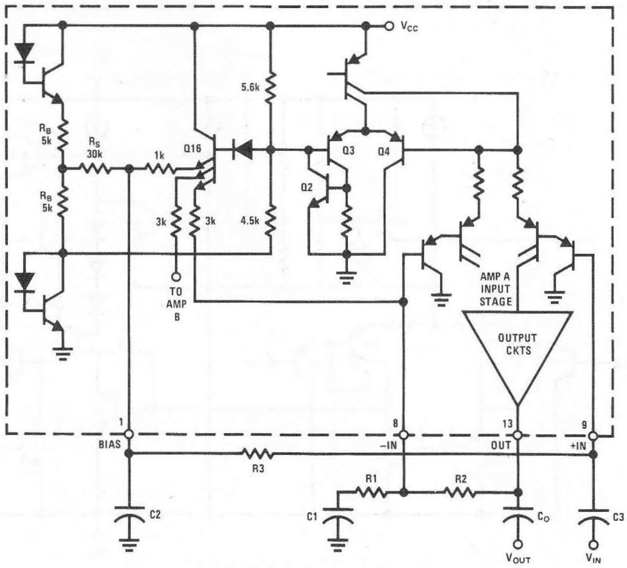


FIGURE 3. Internal Turn-On Circuitry

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 $C2 = 250\mu F$, the $t_{ON} = 3T \approx 0.3$ seconds and PSRR ≈ 75 dB at 120 Hz due to the 30k resistor R_S. Using $C2 = 1000\mu F$, PSRR would be 86 dB. 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 Q4 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.

EXTERNAL BIASING CONNECTION

The internal biasing is complete for the inverting gain connection of *Figure 4* except for the external C2 which provides power supply rejection. The bias terminal 1 may be connected directly to C2 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.

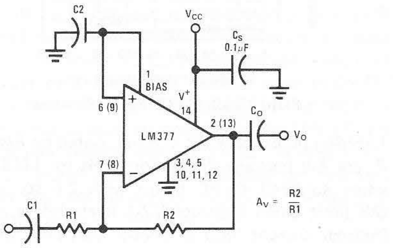


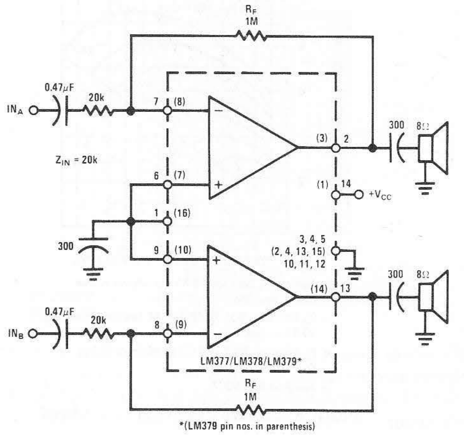
FIGURE 4. Inverting Amplifier Connection

The non-inverting circuit of *Figure 2* is only slightly more complex, requiring the input return resistor R3 from input to the bias terminal and additional input capacitor C3. C1 must remain in the circuit at the same or larger value than in *Figure 4*.

AUDIO AMPLIFIER APPLICATIONS

2/4/6 Watt Stereo Amplifier

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 2 and 4*. The inverting circuit has the lowest parts count so is most economical when driven by relatively low-impedance circuitry. *Figure 5* shows the total parts count for such a stereo amplifier. The feedback resistor value of 1 meg in *Figure 5* is about the largest practical value due to an input bias current max of approximately 1/2µA (100 nA typ). This will cause a −0.1 to 0.5V shift in dc output level, thus limiting peak negative signal



	LM377	LM377/LM378	LM379
P_O =	2W/CH	3W/CH	4W/CH
e_i =	80 mV max	98 mV max	113 mV max
A_V =	50	50	50
V_{CC} =	18V	24V	28V

FIGURE 5. Inverting Stereo Amplifier

swing. This output voltage shift can be corrected by the addition of series resistors (equal to the R_F in value) in the + input lines. However, when this is done, a potential exists for high frequency instability due to capacitive coupling of the 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 6* which has a superior input impedance. For applications utilizing high impedance tone and volume controls, the non-inverting connection will most surely be used.

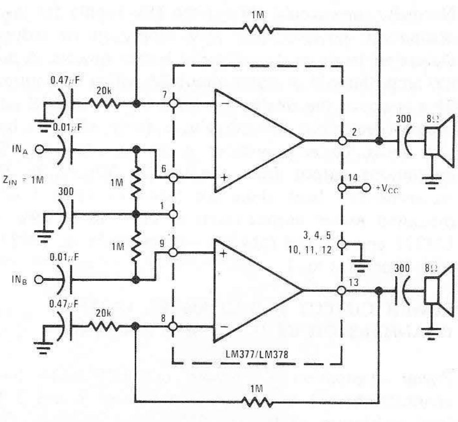
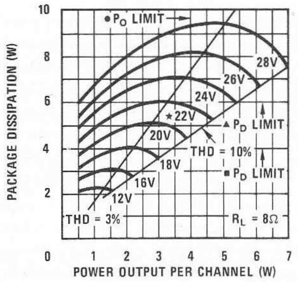


FIGURE 6. Non-Inverting Stereo Amplifier

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 7 through 10.* For other conditions $P_D = V_{CC}^2/20\ R_L$. At high power out, efficiency exceeds 50%



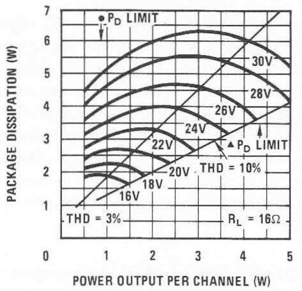
• Approx. P_O limit acc't. 0.7A rms internal current limit at oper. die temp.

▲ 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 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 8. Device Dissipation for 16Ω Load

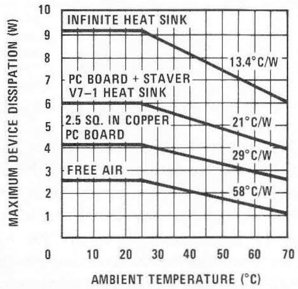


FIGURE 9. LM377/LM378 Power Derating

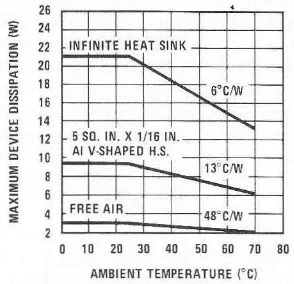


FIGURE 10. LM379 Power Derating

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 = 8Ω. 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 8W continuous output from the LM378 *(Figure 9)*. The recommended heat sinks are listed in Table I 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.

POWER OUTPUT PER CHANNEL (BOTH CHANNELS DRIVEN) BEFORE CLIPPING

Power dissipation vs power output/channel (both channels driven) is indicated in *Figures 7 and 8* for load resistances of 8 and 16Ω.

Limiting points to keep in mind, noted on *Figures 7 and 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. 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 7 and 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.

TABLE I. Continuous Power Out (Both Channels)

HEAT SINK	LM377 R_L = 8Ω			LM378 R_L = 16Ω		
	V_S = 18V	V_S = 20V	V_S = 22V	V_S = 24V	V_S = 26V	V_S = 29V
PC Board, 40°C/W	2.2W	0.8W	0.3W	2.2W	1W	0.3W
PC Board and Staver V7-1, 12°C/W	2.2W	2.7W	3.1W	2.2W	2.5W	3.3W

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 $V_{CC} \geq$ 26V for LM379.

Device dissipation vs ambient temperature with several heat sink types is indicated in *Figures 9 and 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 30 dB ratio between rms and peak power levels in music and speech. If we assume willingness to accept clipping at peak levels 20 dB 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).

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 50 kHz as seen in *Figure 11*. These amplifiers are,

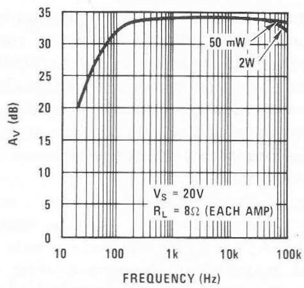


FIGURE 11. Frequency Response of the Stereo Amp of Figure 5

however, not intended for closed loop gain below 25. The typical Bode plot of *Figure 12* shows a phase margin of 70° for gain of 5.6 (15 dB) 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.

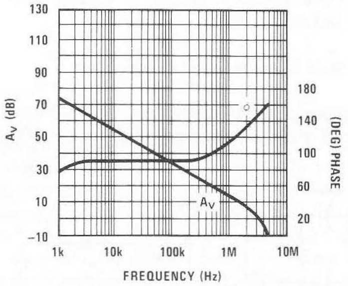


FIGURE 12. Open Loop Bode Plot (Approximately Worst Case)

As with any amplifier of high Gain x BW, careful circuit layout is important to insure unconditional stability. Specifically, coupling to the non-inverting inputs from output and feedback elements should be minimized. The supply should also be bypassed with an 0.05–0.1μF ceramic or 0.47μF mylar capacitor within 2 inches of the IC terminals. One layout which accomplishes this is illustrated in *Figure 13.* The signal circuits are symmetrically arranged on either side of the IC package.

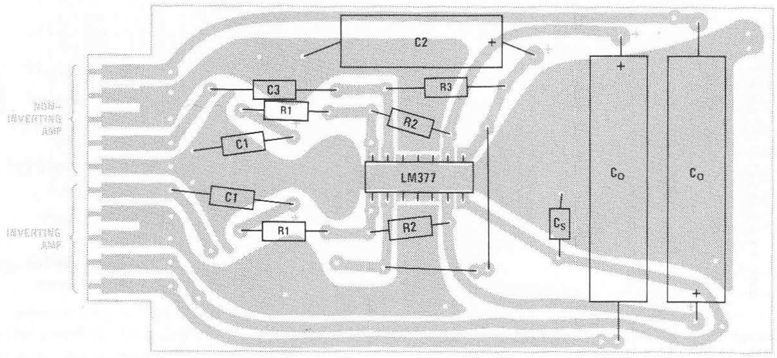


FIGURE 13. Parts Layout for LM377/LM378 Dual Amplifier

Operation may be either inverting or non-inverting, therefore, one side is shown connected each way. Normally both sides would be connected in the same sense. The PC board art-work is shown in *Figure 14.* The edge connector could, of course, be omitted and wire connections made to appropriate contacts at either end. This layout is satisfactory for gain as low as 10. Below A_V = 10, instability may occur.

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). Note that the layout in *Figure 14* has a connection at center of the card edge connector for signal ground and two separate, flanking, terminals for the load and power grounds. 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.

10–12 Watt Channel Boosted Amplifier

Where more power output is desired, the simple booster circuit of *Figure 15* allows power output of 10W/channel

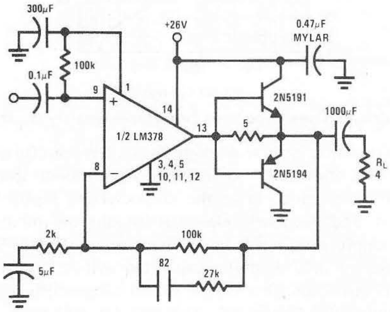


FIGURE 15. 10 Watt Power Amplifier

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 20 mW, the LM378 supplies the load directly through the 5Ω resistor to about 100 mA peak current. Above this level, the booster transistors are biased ON by the load current through the same 5Ω resistor.

The response of the 10W boosted amplifier is indicated in *Figure 16* for power levels below clipping. Distortion

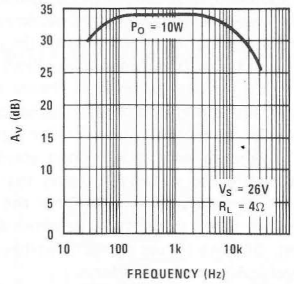


FIGURE 16. 10 Watt Bosted Amplifier, Frequency Response

is below 2% from about 50 Hz to 30 Hz. Fifteen watts 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 17* 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 18 and 19* for low and high power levels. Note that the input coupling capacitor is still required, even though the input may be ground

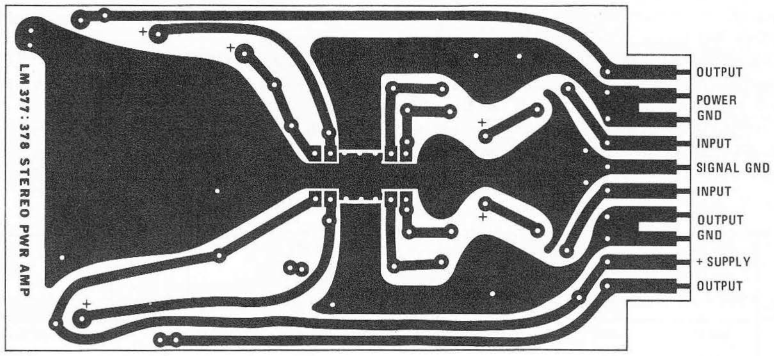


FIGURE 14. PC Board for LM377/LM378 Dual Amplifier

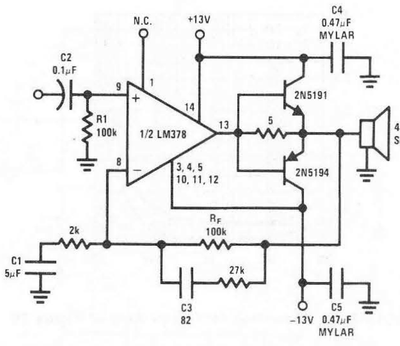


FIGURE 17. 12 Watt Low-Distortion Power Amplifier

referenced, in order to isolate and balance the dc input offset due to input bias current. The feedback coupling capacitor, C1, maintains dc loop gain at unity to insure zero dc output voltage and zero dc load current. Capacitors C1 and C2 both contribute to decreasing gain at low frequencies. Either or both may be increased for better low frequency bandwidth. C3 and the 27k resistor provide increased high frequency feedback for improved high frequency distortion characteristics. C4 and C5 are low inductance mylar capacitors connected within 2 inches of the IC terminals to ensure high frequency stability. R1 and R_f are made equal to maintain $V_{OUT\ DC} = 0$. The output should be within 10 to 20 mV of zero volts dc. The internal bias is unused; pin 1

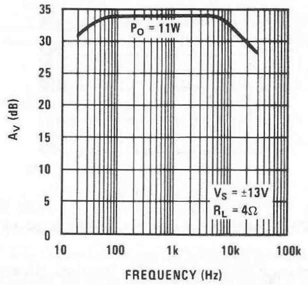


FIGURE 18. Response for Amplifier of Figure 17

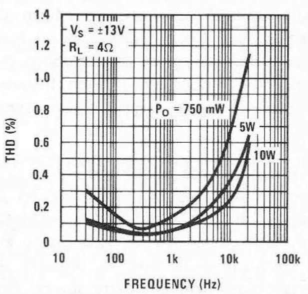


FIGURE 19. Distortion for Amplifier of Figure 17

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.

Bridge Amplifier

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 20*. Response of this circuit is 20 Hz

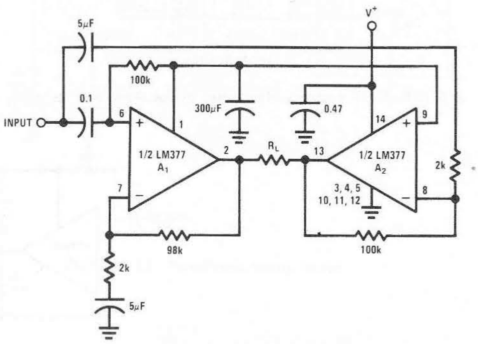


FIGURE 20. 4-Watt Bridge Amplifier

to 160 kHz as shown in *Figure 21* and distortion is 0.1% midband at 4W rising to 0.5% at 10 kHz and 50 mW output *Figure 22*. The higher distortion at low power is due to a small amount of crossover notch distortion which becomes more apparent at low powers and high frequencies. The circuit of *Figure 23* is similar except for higher input impedance. In *Figure 23*, the signal drive for the inverting amplifier is derived from the feedback voltage of the non-inverting amplifier. Resistors R1 and R3 are the input and feedback resistors for A_2, whereas R1 and R2 are the feedback network for A_1. So far as A_1 is concerned, R2 sees a virtual ground at the (−) input to A_2; therefore, the gain of A_1 is (1 + R2/R1). 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 R2 is 100k. A_1 can be considered a unity gain amplifier with internal R = R2 = 100k and R_L = R1 = 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 C1 and C2 may be required to prevent oscillation. With the values shown, the amplifier will drive a 16Ω load to 4W with less than 0.2% distortion midband, rising to 1% at 20 kHz *(Figure 24)*. Frequency response is 27 Hz to 60 kHz as shown in *Figure 25*. The low frequency roll off is due to the double poles C3 R3 and C4 R1.

Power Oscillator

One half of an LM377 may be connected as an oscillator to deliver up to 2W to a load. *Figure 26* shows a Wein bridge type of oscillator with FET amplitude stabilization in the negative feedback path. The circuit employs

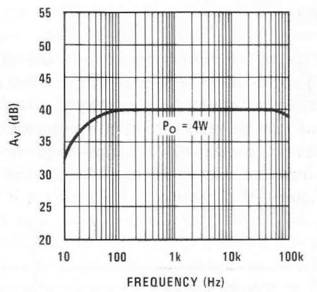


FIGURE 21. Frequency Response, Bridge Amp of Figure 20

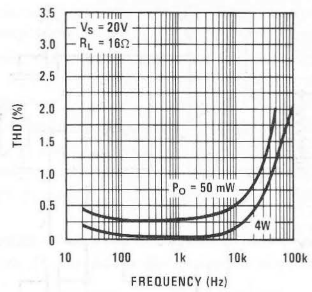


FIGURE 22. Distortion for Bridge Amp of Figure 20

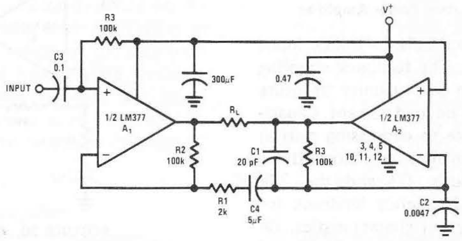


FIGURE 23. 4-Watt Bridge Amplifier with High Input Impedance

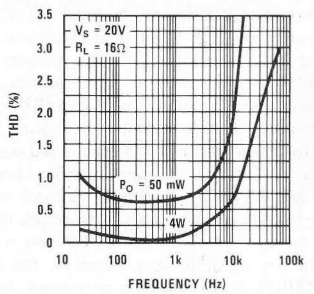


FIGURE 24. Distortion for Bridge Amp of Figure 23

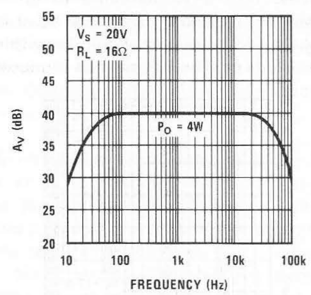


FIGURE 25. Frequency Response, Bridge Amp of Figure 23

internal biasing and operates from a single supply. C3 and C6 allow unity gain dc feedback and isolate the bias from ground. Total harmonic distortion is under 1% to 10 kHz, and could possibly be improved with careful adjustment of R5. The FET acts as the variable element in the feedback attenuator R4 to R6. Minimum negative feedback gain is set by the resistors R4 to R6, while the FET shunts R6 to increase gain in the absence of adequate output signal. The peak detector D2 and C8 senses output level to apply control bias to the FET. Zener diode D1 sets the output level although adjustment could be made if R9 were a potentiometer with R8 connected to the slider. Maximum output level with the values shown is 5.3V rms at 60 Hz. C7 and the attenuator R7 and R8 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 27)* although dc offset will suffer by a factor of about 3 (dc gain of the oscillator). R10 matches R3 for improved dc stability; and the network R11, C9 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, C5, 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. C5 may be 0.1µF ceramic, or 0.47µF mylar. Layout is important; especially take care to avoid ground loops as discussed in the section on amplifiers. If high frequency instability still occurs, add the R12, C10 network to the output.

Figure 27 shows the use of the LM377 to drive a small 60 Hz 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 (R1 = R3 and R6 = R8). High frequency stability is assured by increasing closed-loop gain from approximately 3 at 60 Hz to about 30 above 40 kHz with the network consisting of R3, R4 and C3. The interstage

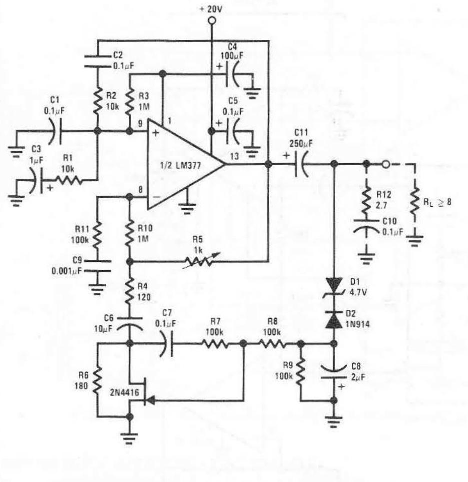


FIGURE 26. Wein Bridge Power Oscillator

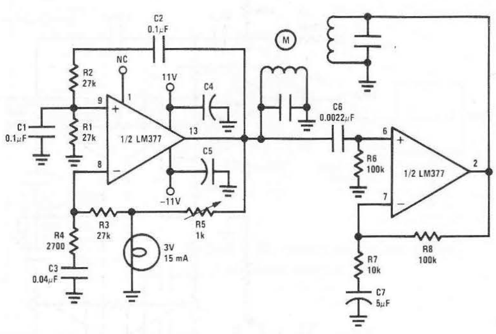
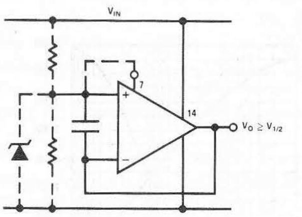
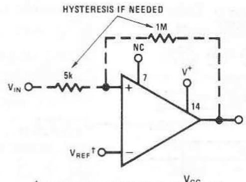


FIGURE 27. Two-Phase Motor Drive

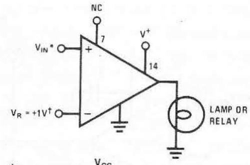


Supply Regulator



† Reference supply must sink $I = \frac{V_{CC}}{6}$ mA unless bias disconnect circuit is used.

Power Comparator



†V_R must sink $I = \frac{V_{CC}}{2k}$

*TTL compatible.

Lamp or Relay Driver

FIGURE 28. Miscellaneous Applications

coupling C6 R6 network shifts phase by 85° at 60 Hz 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 60 Hz with shunt capacitors. This circuit will drive 8Ω loads to 3W each.

MISCELLANEOUS APPLICATIONS

A number of non-audio applications come to mind, such as a dual power supply regulator, power comparator, and relay or lamp driver as shown in *Figure 28.* The degree of practicality of any of these will be limited by the special characteristics of the LM377/LM378/LM379 chip. Limitations are a higher than usual input-offset voltage and temperature drift (not troublesome in the intended capacitor coupled applications). As the devices are not unity-gain stable, a shunt capacitor across the inputs is needed in low gain applications such as the supply regulator. The output saturation voltage is 2 to 3V, thus internal power dissipation is non-negligible in relay or lamp driver applications.

As a lamp driver, the LM377 is limited to those applications where its dissipation is outweighed by the advantage of the internal current limiting. The real advantage of this current limiting is that the lamp driver cannot be destroyed by a shorted lamp *if* the lamp common terminal is ground (the LM377 will not survive an output short to supply).

In any of these applications, recall that the internal turn-on circuit will supply current out the (–) input until inputs are raised to $V_{CC}/2$.

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. Several examples of complete audio systems are described. One is a 2-channel audio system for radio, phono, and tape playback. The other is rear channel amplifier pair for extracting "ambience" information from stereo signals and amplifying for 4-channel sound.

Figures 29 to 30 describe the complete electronic section of a 2-channel sound system with inputs for AM

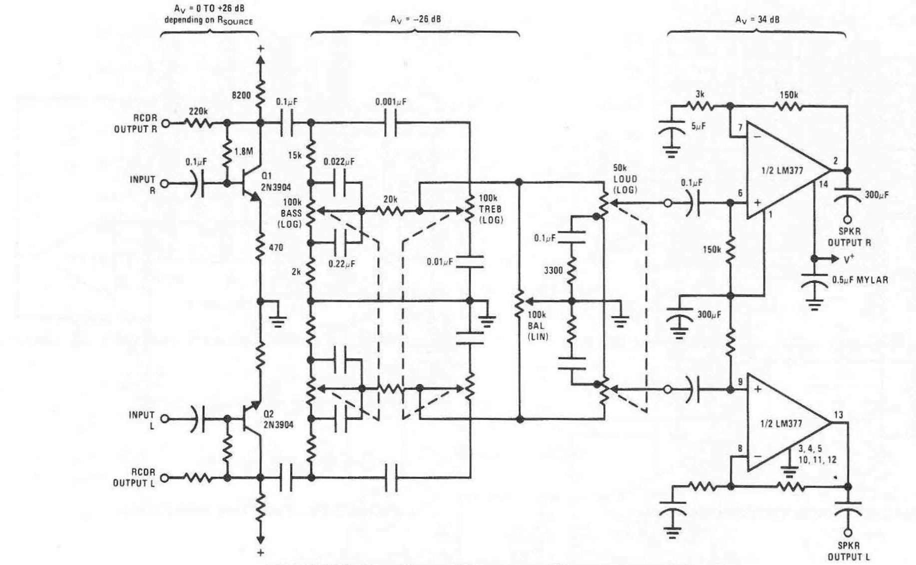


FIGURE 29. Two-Channel Power Amplifier and Control Circuits

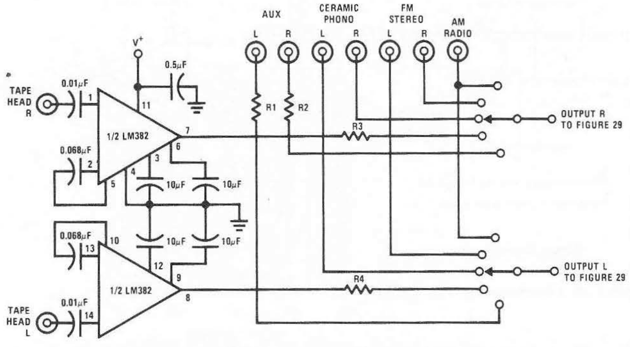


FIGURE 30. Two-Channel Tape-Playback Amplifier and Signal Switching

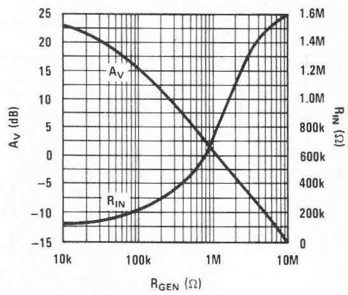


FIGURE 31. A_V and R_{IN} for Input Stage of Figure 28.

radio, stereo FM radio, phono, and tape playback. *Figure 29* combines the power amplifier pair with loudness, balance, and tone controls. The tone controls allow boost or cut of bass and/or treble. Transistors Q1 and Q2 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 Q1 Q2 would be lost. It is believed that this circuit represents the lowest parts count for the complete system. *Figure 30* 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 the data sheets.

Figure 31 shows the relationship between signal source impedance and gain or input impedance for the amplifier stage Q1 Q2. 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 R1 to R4 in *Figure 30*). 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.

The rear channel "ambience" circuit of *Figures 32 and 33* can be added to an existing stereo system to extract a difference signal (R − L or L − R) which, when combined with some direct signal (R or L), adds some fullness, or "concert hall realism" to reproduction of recorded music. Very little power is required at the rear channels, hence an 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 32*.

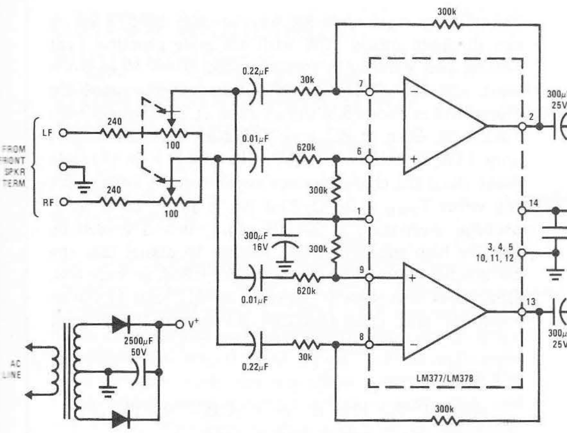


FIGURE 32. Rear Speaker Ambience (4-Channel) Amplifier

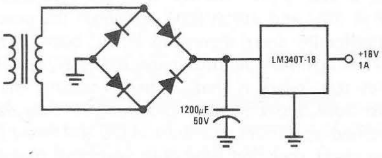


FIGURE 33. Alternate Power Supply for Ambience Circuit

APPENDIX: INTERPRETATION OF P_O vs P_D CURVES

The angled straight lines on the curves of *Figures A-1* and *A-2* 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 A-1* 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 A-3* 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 A-4* showing LM379

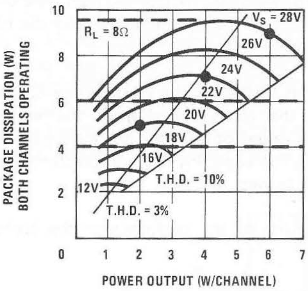


FIGURE A-1. Power Dissipation vs Power Output

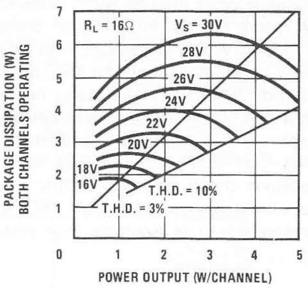


FIGURE A-2. Power Dissipation vs Power Output

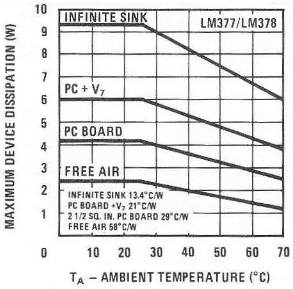


FIGURE A-3. Maximum Dissipation vs Ambient Temperature

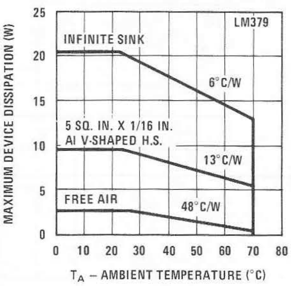


FIGURE A-4. Maximum Dissipation vs Ambient Temperature dissipation with a simple small heat sink. This heat sink is 5 square inches of 1/16" aluminum in a modified V shape which is clamped to the sink side of the LM379.

These practical limits are transferred to *Figure A-1* 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 OK so far. What this means is that, in bench testing, the LM377 must have better than PC mounting to meet data sheet specified operation; and the LM378 will likely not meet data sheet specified operation with any practical heat sink. The LM377 will meet specs with the PC board plus Staver heat sink. The LM378 will meet specs with a 7W heat sink, but not with normally available heat sinks. The LM379 will meet specs with the 5 square inch sink described above. 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 A-1*, 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 A-1* 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.

$$P = I^2 R$$

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

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

Now we have the actual limits at $P_{O(MAX)}$ = 4–6.2W at 8Ω or 8W at 16Ω. Trouble is we are limited to 5W at 28V, 8Ω or 5.5W at 30V, 8Ω and 4W at 16Ω by a 30V operating limit. Current limits could run higher than data sheet 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.

Here is a summary of the performance the customer may encounter.

TABLE A-1. 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	──	──┤					
	├──	──	──	LM378	──	──	──┤			
	├──	──	──	──	LM379	──	──	──	──	──┤