75W HI FI AUDIO AMPLIFIER WITH LOW TRANSIENT INTERMODULATION DISTORTION

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Until very recently, the trend in hi-fi audio amplifiers has been towards lower harmonic and intermodulation distortion numbers. Surprisingly, these ultra-low distortion amplifiers did not have the high sound quality that the distortion figures would indicate. It was soon discovered that the human ear was right and the traditional measurements were incomplete. What had been overlooked was transient distortion.

Transient distortion is high in an amplifier when there is the combination of high open loop gain and high negative feedback. Although this type of amplifier can be made stable, usually its relative stability is low due to a low gain and phase margin. This results in overshoot for a transient signal, which rings out in a dampened wave to its steady state. Another way to look at it is that the correction capacitances perform an integrating function and delay the signal. The leading edge of a transient signal is amplified with the open loop gain until the feedback takes effect. High transient amplitudes and even clipping are the result. If a second signal is present, it will momentarily disappear while the transient goes into clipping; this results in high transient intermodulation distortion (TIM).

The amplifier described here is designed to have low open-loop distortion and relatively low open-loop gain, therefore low transient distortion. This is achieved by using local negative feedback. Also, there is only 23 dB of loop negative feedback. In addition to the lag-type frequency rolloff that is natural in an amplifier, there is a lead-type rolloff in the feedback leg which acts as a phase correction and increases the phase margin of the loop, thereby increasing the relative stability. The result is an amplifier that has no visible overshoot on transients.

CIRCUIT DESCRIPTION

The amplifier is a complementary design to provide minimum open-loop distortion. The front end consists of two difference amplifiers *(Figure 1);* each amplifies the input signal for one side of the push-pull output. The right sides (bases of Q2 and Q4) serve as tie points for both ac and dc feedback. Local feedback is provided by the emitter resistors (R7, R8, R11, R12) and base resistors (R3, R4, R19, R20); the collector resistors are chosen at 2.7 kΩ to hold gain low and maintain good frequency response. The input impedance of the amplifier is high (>30 kΩ) because R2 need supply only the difference in Q1 and Q3 base currents and can, therefore, be high without creating a large offset voltage. Any offset voltage at this point is reflected into the output and causes dc to flow into the load; therefore, Q1 through Q4 should be matched (15% beta match at I_C = 5 mA). Beta matching the remainder of the transistors further reduces offset and distortion.

Input capacitor C1 is large (10 μF) and, together with the high input resistance, gives an extremely low frequency rolloff (0.5 Hz); C1 is used to block off any dc that might come from the input. Large amplitudes at frequencies above 70 kHz require an additional rolloff capacitor C2 in the amplifier input to avoid dynamic crossover distortion in the output transistors. The value of C2 depends on the output resistance of the preamplifier, about 1 nF is a good approximation.

The predrivers Q5 and Q7 have large emitter resistors for local negative feedback. Resistors R25 and R28 are used to lower the collector impedance which would otherwise be extremely high thereby causing high gain that would result in larger negative loop feedback and early open-loop rolloff. Transistor Q6, thermally connected to the heat sink of one of the output transistors, is used as a V_{BE} multiplier to reduce crossover distortion. Capacitors C7 and C8 stop any parasitic oscillations which are small but undesirable in a low distortion amplifier. The idling current is set by R26 to about 20 mA.

The output is a Darlington configuration with fast drivers Q8, Q9 and rugged output transistors Q10, Q11. Emitter resistors R33 and R34 supply local feedback. The combination of R32 and C11 provides off-drive for the bases of the output transistors Q10, Q11 to speed up the turn-off. This raises the onset of large signal dynamic crossover distortion from ~30 kHz to ~70 kHz.

Resistor R35 and capacitor C12 provide a constant high frequency load for the amplifier. Certain speakers (electrostatic) may require an inductance in parallel with a damping resistor at the output; 10 μH and 10 Ω are suitable values. In this case, the feedback should be taken directly from the speaker to compensate for the voltage drop across the LR combination.

The feedback ties back into the two difference amplifiers. DC is coupled back directly; for ac, the blocking capacitor C6 is chosen for low-frequency rolloff (1.5 Hz). This lag-type low-frequency rolloff is the only rolloff within the feedback loop and, therefore, the amplifier is unconditionally stable at low frequencies. At medium frequencies, the amplifier gain is set to R22/R23 = 22. Capacitor C5 parallel to R22 provides a lead-type high-frequency rolloff which is phase-correcting and provides better transient stability.

The supply voltage at the front end is ac-decoupled by R30 and C9 on one side and R31 and C10 on the other. The diodes D1 and D2 prohibit the discharge of C9 and C10 into the output thus providing voltage for the front end to avoid switch-off noise when the amplifier is turned off.

FAIRCHILD

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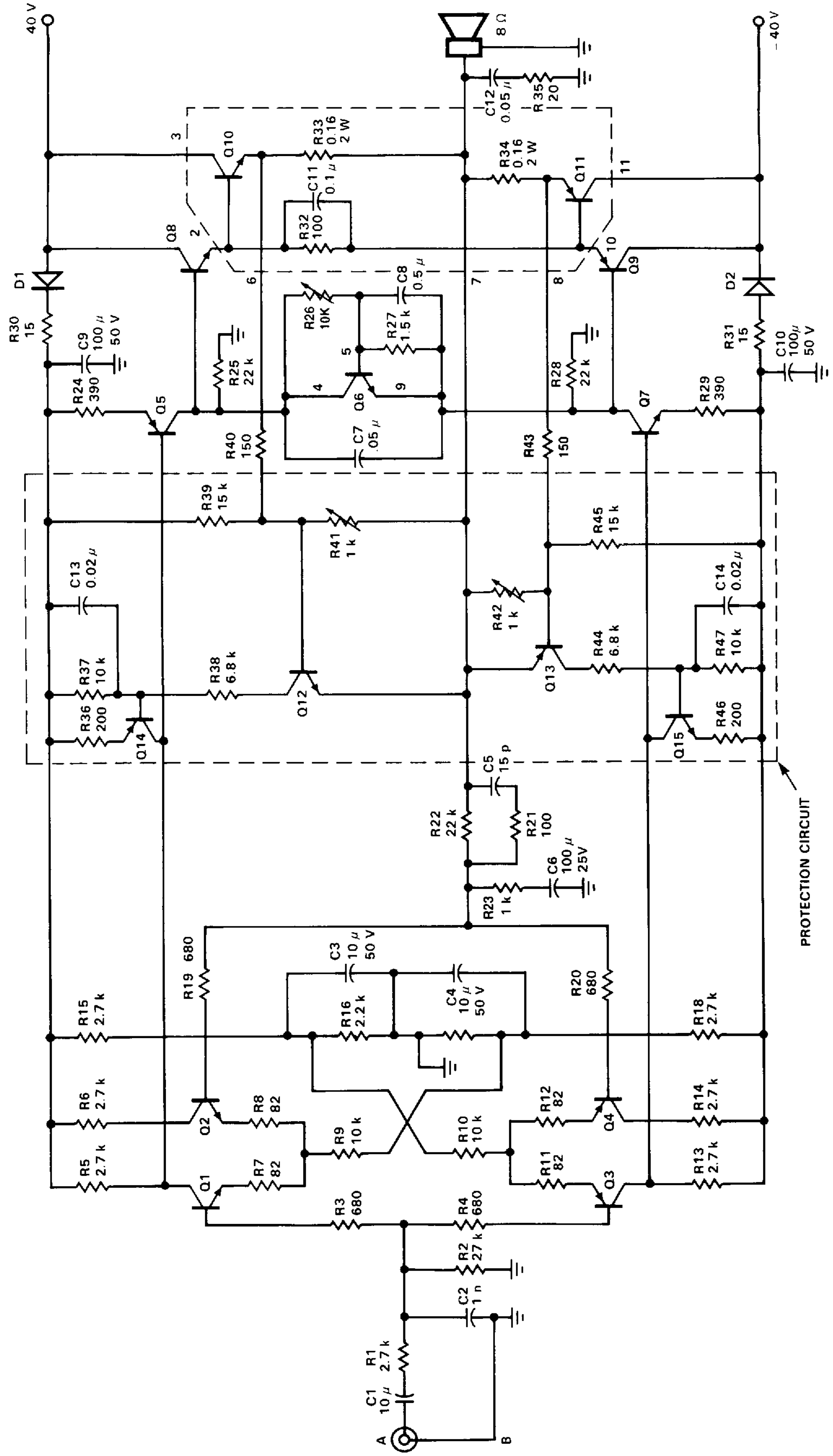

Fig. 1 75 W Audio Amplifier With Load Line Protection

Table 1. Recommended Transistors

TRANSISTOR		TYPE	POLARITY	$V_{CEO(sus)}$ Min V	h_{FE} Min @ I_C A		f_T typ MHz	REMARKS
Front End	Q1, Q2	2N5961	npn	60	135	1 m	150	low noise high gain down to μA, gain linearity
	Q3, Q4	PN4250A-18	pnp	60	250	100 μ	70	
Multiplier	Q6	2N5961	npn	See Above				
Pre driver	Q5	2N5400	pnp	120	40	10 m	200	high voltage, rugged
	Q7	2N5830	npn	100	80	10 m	200	
Driver	Q8	FT317	npn	100	35	1	35	high voltage, fast 40 W, TO-220
	Q9	FT417	pnp	100	35	1	25	
Output	Q10	FT324	npn	140	20	5	6	200 W TO-3, rugged SOA; 50 V, 3.0 A medium speed
	Q11	FT424	pnp	140	20	5	6	
Protection Circuit	Q12	2N5831	npn	140	80	10 m	200	high voltage rugged
	Q13	2N5401	pnp	150	60	10 m	150	
	Q14	PN4250A-18	pnp	See Above				high gain
	Q15	2N5961	npn	60	150	10 m	100	

OVERLOAD PROTECTION

Excessively large input signals or a momentary short circuit in the output will overstress the output transistors and destruction may result. This can be avoided by using a protection circuit to clip off excessive signals that would exceed the safe operating area of the output transistors. The signals are clipped along a load line that starts at 5 A output current for zero voltage across the output transistor and goes linearly to 80 V at zero current. This voltage – double the supply voltage – is required because the speakers normally constitute an inductive load.

For protection of output transistor Q10, the current-sensing resistor R33 determines the 5 A limit, R39 and R40 set the 80 V limit, and potentiometer R41 adjusts the protection to the $V_{BE(on)}$ of transistor Q12. *Figure* 2 shows the load lines for one output transistor. V_{CC} is the supply voltage for one side of the amplifier, R_L is the load line for a resistive load, in this case 8 Ω. The inductive load lines for an inductive speaker are ellipses of decreasing amplitude and increasing phase-angle ϕ for increasing frequency. The envelope is a straight line connecting the coordinates 0, I_p and 2 V_{CC}, 0. Speaker systems using crossover networks have similar impedances. The protection circuit clips off any signal beyond this line but does not shut off the signal altogether.

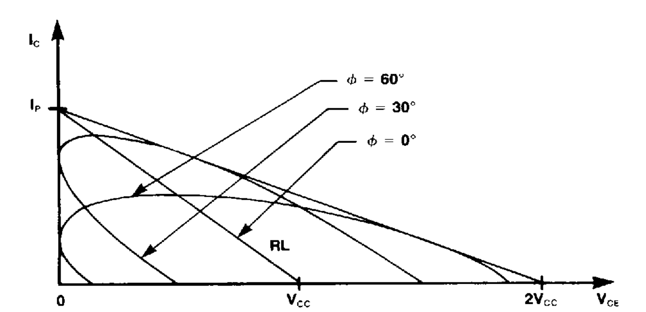

Fig. 2 Output Transistor Loadlines

The load line (envelope of the load ellipses) can be expressed:

$$\frac{I_C}{I_P} + \frac{V_{CE}}{2V_{CC}} = 1 \qquad (1)$$

The base voltage of the load sensing transistor Q12 can be expressed by:

$$V_{BE} = I_C R33 \frac{R41}{R40 + R41} + V_{CE}\frac{R40}{R39} \times \frac{R41}{R40 + R41} \qquad (2)$$

which is a good approximation for R33 ≪ R40 ≪ R39. *Equation* (2) can be written as:

$$\frac{I_C}{V_{BE}\dfrac{R40 + R41}{R33 \times R40}} + \frac{V_{CE}}{V_{BE}\dfrac{R40 + R41}{R40} \times \dfrac{R39}{R40}} = 1 \qquad (3)$$

Direct comparison with (1) yields

$$I_P = V_{BE}\frac{R40 + R41}{R40} \times \frac{1}{R33} \qquad (4)$$

$$2V_{CC} = V_{BE}\frac{R40 + R41}{R40} \times \frac{R39}{R40} \qquad (5)$$

In this case I_P = 5 A, $2V_{CC}$ = 80 V. For R33, 0.16 Ω is selected. Substituting in *Equation* 4, the following results.

$$V_{BE}\frac{R40 + R41}{R40} = 0.80 \text{ V} \qquad (6)$$

With this value substituted in *Equation* (5) and 150 Ω selected for R40, R39 = 15 Ω. Resistor R41 is a 1 kΩ potentiometer set so that the V_{BE} turn-on voltage of transistor Q12 fulfills *Equation* (6).

When Q12 is turned on, it activates Q14, which shunts the signal off Q5. Capacitor C13 is used to stabilize the protection loop and suppress oscillations. The high-frequency rolloff for this protection loop is at Q5, as it is for the whole amplifier. The protection for the lower side is complementary and works in the same manner to protect Q11.

PC BOARD LAYOUT

The pc board *(Figures 3, 4)* is laid out as symmetrically as possible with amplification going in one direction. Input and output grounds are decoupled to avoid ground-loop problems. The drivers Q8 and Q9 are TO-220 mounted upright and carry friction-fit heat sinks of about 20°C/W (Model 106B from Fab-Tec, Danbury, CT). Each output transistor is on a separate heat sink (θ = 2.7°C/W) mounted with thermal grease. Transistor Q6 is mounted snugly into a hole in one of the output heat sinks. Thermal grease is advisable, but if there is any danger that transistor Q6 could slip out of its hole and lose thermal connection, epoxy should be used.

In case it is necessary to compensate the speaker impedance with a series LR combination, the feedback can be taken directly from the speaker. This is accomplished by drilling a hole at point X and connecting the feedback to the speaker by a separate wire from point Y.

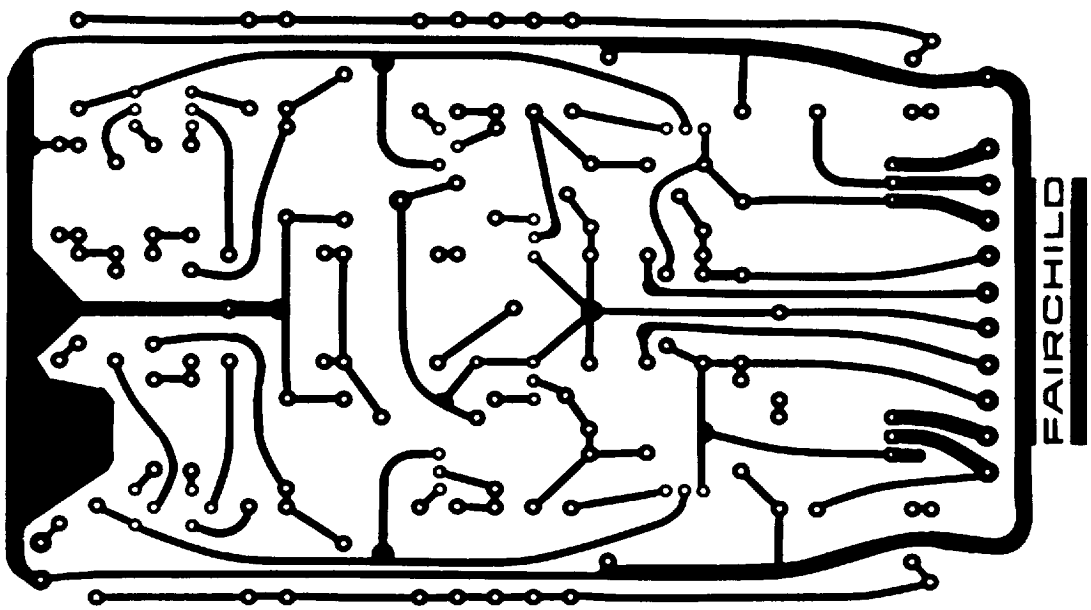

Fig. 3 PC Board, Circuit Side

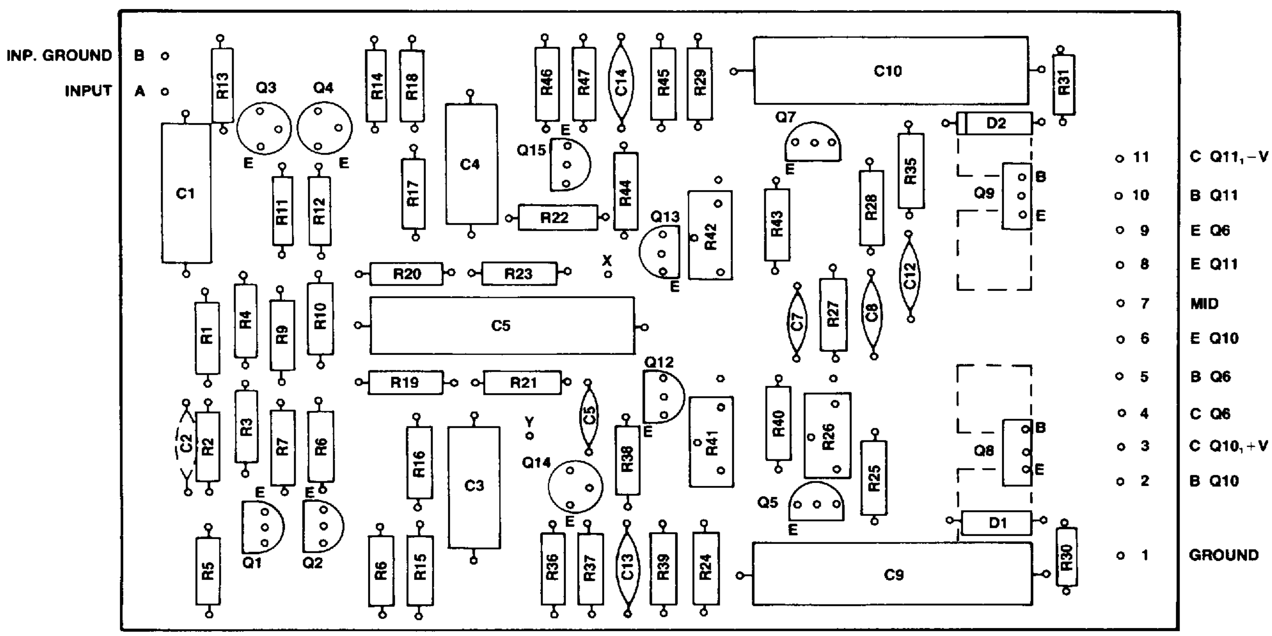

Fig. 4 PC Board, Component Side

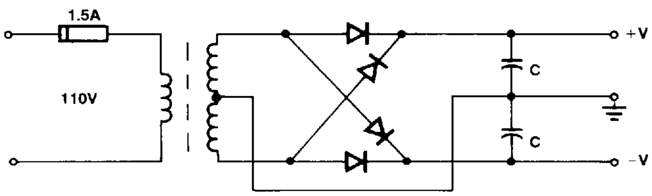

Fig. 5 Power Supply

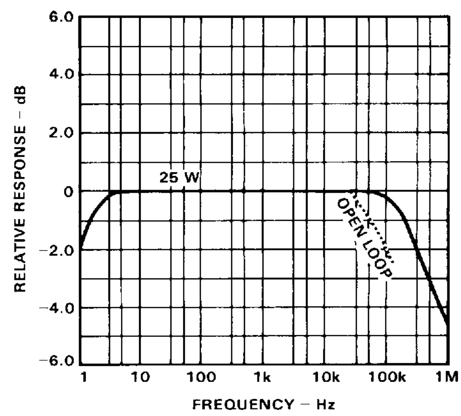

Fig. 6 Frequency Response

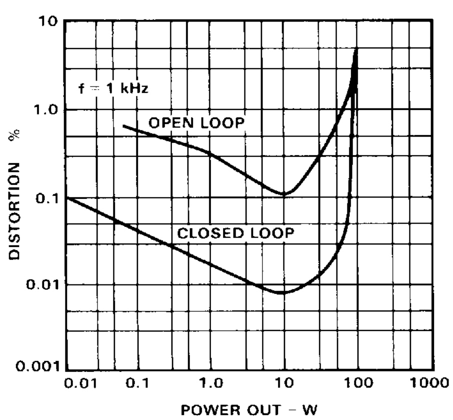

Fig. 7 Total Harmonic Distortion

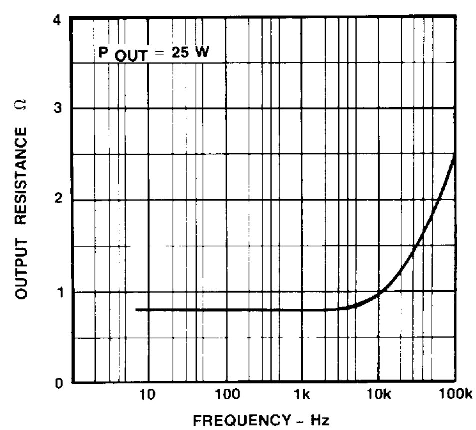

Fig. 8 Open Loop Output Resistance

POWER SUPPLY

An unregulated split supply is used *(Figure 5)*. The transformer is 110 V/56 V center tapped and should be able to deliver

$$P = \frac{V_{CC}^2}{2R_L} \times \frac{4}{\pi} = \frac{40^2}{2.8} \times \frac{4}{\pi} = 128 \text{ W} \qquad (7)$$

Generally, the lower the output resistance of the transformer, the better. The capacitances C were chosen 5000 μF; the bigger the C, the stiffer the supply.

PERFORMANCE

The basic design philosophy was to build an amplifier that works well without a feedback loop, and add a little feedback to get an extremely stable amplifier that has low distortion and wide bandwidth at the same time.

The amplifier without feedback has a linear frequency response far above the audible range; it rolls off 1 dB at 70 kHz *(Figure 6)*. The harmonic distortion is below 1% *(Figure 7)*, and the output resistance typically 0.8 Ω *(Figure 8)*.

With the feedback connected, these performance figures improve drastically. The 1 dB frequency rolloff shifts up to 300 kHz and the rolloff at the low end is 1 dB below 3 Hz *(Figure 6)*. The harmonic distortion has improved by a factor of 10 and is below 0.1% up to 75 W output; in the middle range it is below 0.01% *(Figure 7)*. Distortion vs frequency is below 0.03% up to 35 kHz for 25 W output *(Figure 9)*. Due to the feedback, the output impedance has dropped to typically 20 mΩ *(Figure 10)*; the input resistance is high due to the double differential input, about 30 kΩ. The voltage gain is set to about 23, and the input voltage for full output is below 1.4 V_{rms} which can be handled by any commercial pre-amp. Transient stability of this amplifier is excellent; a 1 kHz square wave shows no visible overshoot or ringing.

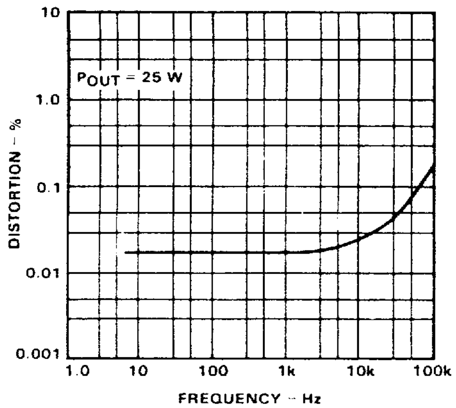

Fig. 9 Total Harmonic Distortion (Closed Loop)

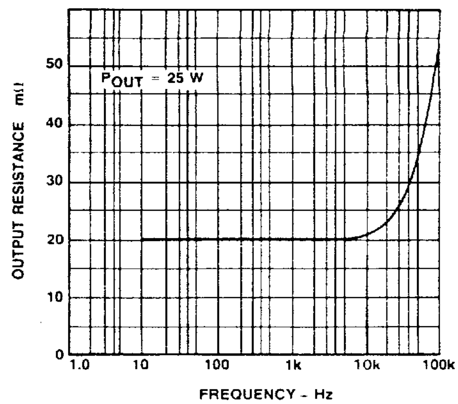

Fig. 10 Total Harmonic Distortion (Closed Loop)

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