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AN AUDIO POWER AMPLIFIER FOR ULTIMATE QUALITY REQUIREMENTS

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ABSTRACT

A 50 W/4 ohm complementary symmetry audio amplifier concept is described. The amplifier is designed to meet ultimate quality requirements with particular emphasis on lack of transient intermodulation distortion. Excellent phase and amplitude linearity is obtained using low feedback value, class A operation for most of the signal time, and a fully symmetric dc-coupled high frequency design.

In the conference session a sound demonstration of transient intermodulation distortion effects will be given.

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Introduction

It is better to overestimate than to underestimate the capability of the ear to detect amplifier imperfections. Recently a number of theories and experiments concerning phase linearity requirements, transient intermodulation effects, and psychoacoustics have shown that some of the basic "truths" of audio amplifier design should be revised.

The purpose of this work is to try to meet the new, drastically different requirements. The approach chosen is to avoid all apparently "elegant" tricks frequently employed (bootstrapping, positive feedback, distortion compensation, higher feedback for dc than ac, gliding bias etc.) and to produce a state-of-the-art amplifier for ultimate quality requirements.

Transient intermodulation distortion

This type of distortion, occurring at non-stationary signals, is caused by the feedback [1,2] and seems to plague many commercial amplifiers [2]. When an input signal, having a sufficient amplitude and a higher frequency than the power amplifier open-loop cut-off frequency, is presented in the input, overshoots in the internal loop drive voltage of the feedbacked amplifier are produced. Depending on the value of feedback, these overshoots may be several hundred times greater than the nominal value of this voltage. Consequently, they will be clipped in the driver stages of the amplifier if sufficient overload margins do not exist, and will then produce momentary 100% intermodulation bursts. Because the usually encountered audio signals have their maximum slopes around the zero signal level, these distortion bursts usually occur in this region. The audible effect resembles high-frequency cross-over distortion and according to recent studies [3], the ear seems to be extremely sensitive to it.

To avoid transient intermodulation distortion, the open-loop frequency response of the power amplifier should exceed the preamplifier frequency response. This open-loop response is generally governed by the lag compensation network of the amplifier. The amount of lag compensation necessary to stabilize a given amplifier, in turn, is directly proportional to the amount of feedback used. The designer therefore has a difficult trade-off dilemma: in order to decrease the usual steady-state harmonic and intermodulation distortions, he would like to use as much feedback as possible. This would, however, drastically decrease the open-loop frequency response owing to the necessary heavy compensation, and thus increase transient intermodulation distortion.

The optimum feedback has a value where all these distortion phenomena contribute equally to the audible distortion sensation. The recent tendency to use higher values of feedback than necessary has radically upset this balance.

The various means of minimizing the transient intermodulation distortion include [4] :

- good high-frequency design along with the use of lead compensation instead of lag compensation within the feedback loop;
- use of low value of feedback;
- if lag compensation must be used, it should be placed in the input of the amplifier [4,5] in order to get passive attenuation of transients instead of active clipping.

The first of these methods requires gain reserves from the amplifier stages in order to make effective lead compensation possible, and the second requires very good open-loop linearity, because heavy feedback is not present to decrease static harmonic and intermodulation distortion. After these requirements are fulfilled, the power amplifier small-signal frequency response will be of the order of 1 MHz, and the limiting of the frequency response to 20 kHz [6] must be accomplished with a passive filter in the preamplifier.

Phase and amplitude linearity

Recent psychoacoustic experiments have convincingly shown that the ear is sensitive to phase relationships of the components of an audio signal [7,8,9,10]. It seems that only 10° deviation from exact phase linearity can be tolerated in the audible frequency range for the most critical requirements in direct field listening conditions. If phase correction networks are not used, this implies an amplifier closed-loop small-signal frequency response in excess of 200 kHz.

The important frequency response parameter is the power bandwidth, which requires redefinition. At present it is measured at the distortion level of 1% (European DIN 45500 requirement) or even as the amplifier limiting level. Especially at the critical high frequencies, severe distortion is usually present far below the limiting level, and the occurrence of two or more high-frequency signals will cause strong difference intermodulation products to fall within the audible frequency range.

The power bandwidth should therefore always be measured at a specified distortion level. For professional quality amplifiers we propose the use of 0.2% total harmonic distortion as a measurement standard.

This measurement is fairly straightforward and simple, the only precaution being the disconnection of possible output filter elements (i.e. L_1-R_{35} and C_7-R_{34} in Fig. 1), which may attenuate the higher harmonics.

The definition of power bandwidth should then be the -3 dB points of the measured response; i.e. the half-power points. It should exceed the preamplifier frequency range in order to prevent distortion and amplifier overload due to high frequency components of the signal.

Other design criteria

The ear seems to be very sensitive to amplifier imperfections around the zero output level. Daugherty and Greiner [11] originally proposed that the amplifier should operate in class A for at least 50% of the signal time, and derived the corresponding A- to B-class transition level to be 17 dB below the full output power. This almost-class-A-operation completely eliminates cross-over distortion and greatly diminishes class-B-type unsymmetries, which are easily caused by improper design, or even in well designed fully complementary symmetry amplifiers by fundamental differences in the PNP and NPN power transistors.

It is common practice to employ higher feedback values for dc than for ac, motivated by the apparent increase of dc stability. However, this argument is not valid, because an sufficient dc stability can easily be achieved without this kind of tricks by proper use of circuit symmetry and temperature compensation. On the other hand, a high dc feedback value can cause exotic clipping and distortion effects, and infra-frequency rumble, because of open-loop B-class gain unsymmetries that force dc level adjustment transients through the feedback loop every time the signal amplitude changes. These effects usually manifest themselves in tone burst tests and noise cross-correlation tests, and are not discernible with steady-state sinusoidal signals.

It is therefore desirable to make the feedback value the same for ac and dc, at the same time eliminating a number of undesirable large capacitors, which always seem to be sources of trouble in class AB amplifiers.

The use of usual driver collector bootstrapping can give rise to unsymmetries and poor clipping behaviour. These problems can be eliminated by the use of symmetrical drivers which, apart from good linearity and ideal limiting characteristics, also lend themselves to integration of the power amplifier pre-stages.

The amplifier design

The circuit diagram of the amplifier is shown in Fig. 1. The basic design philosophy has been:

i to provide the required high open-loop linearity by using heavy local feedback in every amplifier stage. This minimizes the need for heavy overall feedback and at the same time increases the open-loop cutoff frequency of the amplifier;

ii to use the high-power stages T_{12-17} in a grounded collector configuration in order to increase their frequency response and linearity;

iii to use voltage drive in the crucial output and driver stages to force the transistors to operate with a cutoff frequency of almost f_T instead of f_β, which is usual in current-driven designs;

iv to use only lead compensation networks within the amplifier to compensate for the transfer function poles up to 1 MHz. This makes possible an unconditionally stable amplifier;

v to use a completely dc-coupled design with the same amount of feedback for ac and dc. This eliminates the usual dc level shift problem which can cause severe unsymmetrical clipping of transient signals;

vi to use in all stages a completely symmetrical design without bypass or bootstrap capacitors. This significantly improves the amplifier clipping and overload characteristics, makes clipping recovery instant and avoids dc level shift problems, as frequently encountered in conventional designs. Furthermore, it improves amplifier linearity especially at high frequencies, where the even harmonics usually become prime source of distortion;

vii. to use all stages in class A except the power transistors, which operate in class AB with a quiescent current of 600 mA and provide about 3 W output power in class A. This totally eliminates cross-over distortion and unsymmetries in the critical low-power region.

The value chosen for the overall feedback was 20 dB, and a closed loop gain of 32 dB was required to make the input sensitivity -6 dBV (0.35 V_{rms}). The total open-loop gain was then divided as follows:

13 for T_{1-2}, 3 for T_{4-5} and 11 for T_{7-8}

(measured differential in - differential out).
The three first poles of the transfer function occur at frequencies between 200 kHz and 1 MHz, and are lead compensated by C_4-R_{43}, C_2, and C_6. The total open-loop frequency response after lead-compensation has a -3 dB point at 1MHz. The final 6 dB/octave rolloff is produced by lag compensation network C_1-R_6, which is situated outside the loop [4,5]. The -3 dB point of this network is 100 kHz and the compensation is cut off at 1 MHz by R_6. The phase margin of the feedback loop is greater than 80° at 1 MHz.

The amplifier is an inverting one with the same feedback value for ac and dc. Dc stability is ensured by using an input stage with a very good temperature stability. A dual transistor with an offset voltage drift of 3 µV/°C maximum is used for T_1 and T_2. With a maximum temperature excursion of ± 50°C, the output offset voltage is below ± 6 mV, which is negligible compared with normal output signals. Resistor R_{13} is to be adjusted for zero dc output voltage.

For good linearity and a high cutoff frequency of the output stage, the bases of transistors T_{12} and T_{15} must be driven as symmetrically as possible and from a relatively low impedance. This is accomplished with transistors T_{7-11} and R_{18-27}. Transistors T_7 and T_8 are a normal differential pair. T_9 and T_{10} form a current mirror circuit, familiar from integrated circuit technology. The current flowing through transistor T_{10} has the same amplitude as the current through transistor T_9. Therefore transistors T_8 and T_{10} perform current pumping in a symmetrical manner. The load for these two current sources is formed by a simulated zener diode (T_{11}, R_{23}, R_{24} and R_{25}) and two resistors (R_{26} and R_{27}). The simulated zener diode keeps a constant bias voltage between the bases of T_{12} and T_{15}. The amplitude of the voltage swing on these two bases equals the current difference between T_8 and T_{10} times R_{36} and R_{27} in parallel. Transistors T_{12} and T_{15} are therefore driven from a symmetrical source with an impedance of 1.1 kΩ. Rather large emitter resistors R_{18}, R_{19}, R_{21}, R_{22} are needed for good linearity, small gain and linearization of unequalities in base-emitter voltages of the transistors T_{7-10}. This will give rather high voltage losses which prevent the output stages from being driven over the full power supply range. The solution is to use second power supplies with higher voltages for the pre-stages. In this design $\pm$ 30 volt is used for the pre-stages and $\pm$ 24 volt for the output stages.

Transistors T_{12} and T_{15} work in class A. When the output transistors are driven in class B, and are switched off alternately, T_{12} and T_{15} remain conducting, so that the charge in the base-emitter junctions of the output transistors can easily flow away through the low output impedance of T_{12} and T_{15}. As a result the switching behaviour of the output transistors is very good.

The quiescent current in the output stage is 600 mA. With 4Ω load, class A operation is extended to about 3 Watt output power. This is 13 dB below the maximum rated output power, fulfilling adequately the requirement of having the amplifier more than 50% of the signal time in class A [11].

The output transistors are cheap plastic-package types BD203 and BD204 (f_β > 25 kHz, f_T > 3 MHz). For power dissipation reasons, two of each are connected in parallel.

The high-frequency transients in the supply lines are suppressed with capacitors C_{8-15}. This suppression is aided by placing damping resistors (R_{39-42}) in series with some of the capacitors. The resulting resonant circuits formed by the supply line inductances and the capacitor-resistor networks are over-critically damped.

The amplifier short-circuit and overload protection circuit is a conventional two-transistor network between the power transistors and the driver bases. To make the circuit diagram clearer, this protection circuit is not shown in Fig. 1.

Mechanical construction

The mechanical construction of the amplifier is shown in Fig. 2. The components are directly mounted on the heat-sink, which has 0.75 °C/W thermal resistance to the ambient. The amplifier quiescent power dissipation is 31 W and the maximum class AB total power dissipation is 60 W. The thermal stability is excellent, and because of the relatively high quiescent power dissipation the transistor junction temperature variation is rather small in operation.

Measurements

A survey of the measurement results is given in table 1.

TABLE 1

- rated output power	: 50 W_{rms}
- nominal load impedance	: 4 Ω
- output impedance	: 50 mΩ for 1 kHz 70 mΩ for 20 kHz
- damping factor	: better than 50
- input impedance	: 3400 Ω
- input sensitivity	: 0.35 V_{rms} for rated output power
- feedback value	: 20 dB
- small-signal frequency response	
0-20 kHz	: flat within 0.05 dB
1.0 MHz	: -3 dB
- power bandwidth for 0.2% harmonic distortion	: 35 kHz
- phase characteristic 0-20 kHz	: flat within 3.5°
- slew rate	: 100 V/µs
- quiescent current (output stage)	: 0.6 A
- class A to B crossover point (4 Ω load)	: 2.9 W
- harmonic distortion, 1 kHz, 50 W	: < 0.2%
- intermodulation distortion, 7 kHz/250 Hz (1:4)	: < 0.15%
for 25 W_{rms} power level of the 250 Hz component	
- noise level	: 106 dB(A) below full rated output power.

All measurements were done with short-circuited L_1.

Amplifier small-signal frequency and phase responses are shown in Fig. 3. The -3 dB point is 1 MHz, at which frequency the phase shift is 100°. About 45° of this phase shift is caused by first-order rolloff, and 55° by a propagation delay of approximately 150 ns. The frequency response is flat within -0.05 dB and the phase response within 3.5° in the frequency range 0-20 kHz.

The power bandwidth is shown in Fig. 4 for different criteria. With the usually employed criterion of visual clipping, the amplifier has a power bandwidth of 500 kHz. With the criterion of 0.2% total harmonic distortion proposed by us, the power bandwidth becomes 35 kHz, which is just sufficient for a signal source upper cutoff frequency of, say, 30 kHz.

The harmonic distortion is shown in Fig. 5 for different frequencies. As can be seen, the distortion at low-power levels for frequencies up to 10 kHz is infinitesimally small. No traces of cross-over distortion can be found due to class A operation. The distortion figures up to 30 kHz are exceedingly good, especially when they are obtained with only 20 dB overall feedback. To measure the extremely small distortion values a HP3590A waveform analyzer was used, and the signal source was a Philips PM5125 signal generator followed by a Wandel & Golterman 1/3-octave filter bank. The same set-up was used to measure the intermodulation distortion, which was below 0.15% up to maximum output power of the amplifier. The test signals used were 7 kHz and 250 Hz in the ratio 1:4.

Figs. 6a and 6b show the square-wave responses at 1 kHz and 100 kHz, clearly demonstrating the good amplitude and phase linearity and the unconditional stability of the amplifier.

Figs. 7a and 7b show a burst-tone test for frequencies of 1 kHz and 100 kHz, respectively. The results are faultless.

When used in conjunction with a preamplifier having an upper cutoff frequency of 60 kHz maximum the amplifier produces no internal transient overshoots and consequently does not exhibit transient intermodulation distortion at all.

All these measurements were done with short-circuited L_1. Even then the amplifier is unconditionally stable for capacitive loading up to 10 μF (maximum tested).

A compensation test as in Fig. 8 was also performed. Because the amplifier is an inverting one, a position of potentiometer P can be found where the linear component of the output signal exactly compensates the input signal at the "test" point. The resulting signal at the "test" point then shows the total discrepancy between the input and the output signals. This momentary discrepancy signal, as scaled up to the output level, was below 0.01% for powers below 5 W and for frequencies in the audio range, and was mainly caused by incomplete phase compensation (R_1, R_2 and C_1 in Fig. 8). Near clipping the maximum discrepancy was of the order of 0.1%, and was caused by class B unsymmetries, and the previously mentioned incomplete phase compensation. Sinusoidal, noise, and music signals were tested.

Conclusions

Using available low-frequency audio power transistors it is possible to construct an audio amplifier exhibiting exceedingly good phase characteristics, sufficient power bandwidth, no transient intermodulation distortion, and vanishingly small harmonic and intermodulation distortion. The amplifier is free from usually employed circuit tricks, and because of its symmetrical design and the absence of large capacitors it should lend itself to partial integration. The price paid is a relatively high quiescent power dissipation and the use of four power supplies, but they should not constitute severe limitations to the applicability of the design in cases where ultimate quality is required.

Summarizing, the amplifier does not exhibit any known source of distortion, instability or undesired response, which would not be far below the psychoacoustical sensation thresholds.

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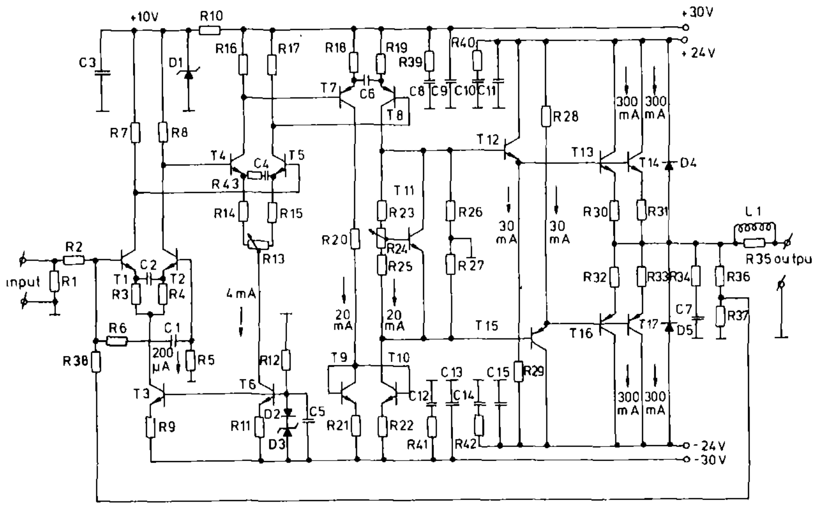


Fig. 1. Circuit diagram of the amplifier. In order to make the diagram clearer the short-circuit and overload protection circuitry is *not shown*. *The components* are as follows:

T1 & T2	= dual transistor pair BCY 87
T4 & T5	= " " " BCY 89
T3,6	= BC 107
T7,8,15	= BD 140
T9,10,11,12	= BD 139
T13,14	= BD 203
T16,17	= BD 204

D1	= BZY88C10
D3	= BZY88C13
D2,4,5	= BA145
C1	= 220 pF
C2,4	= 120 pF
C3,5,7,9,11,13,15	= 0.1 µF
C6	= 1.5 nF
C8,10,12,14	= 0.68 µF
L1	= 2 µH

R1,2,10	= 6.8k
R3,4 metal film	= 1k
R5	= 10k
R6	= 470Ω
R7,8	= 22k
R9	= 68k
R11	= 3.3k
R12	= 4.7k
R13 (pot.)	= 220Ω
R14,15 metal film	= 390Ω
R16,17	= 1.8k
R18,19,21,22 metal film	= 100Ω
R20	= 560Ω/1W

R23	= 1.5k
R24 (pot.)	= 1k
R25	= 470Ω
R26,27,36	= 2.2k
R28,29	= 820Ω /1W
R30,31,32,33	= 1Ω /2W
R34	= 12Ω /2W
R35	= 2.2Ω /2W
R37	= 390Ω
R38	= 39k
R39,40,41,42	= 1Ω /1W
R43	= 180Ω

Resistors $\frac{1}{4}$ W carbon film unless otherwise specified, small capacitors polystyrene, large capacitors polyester.

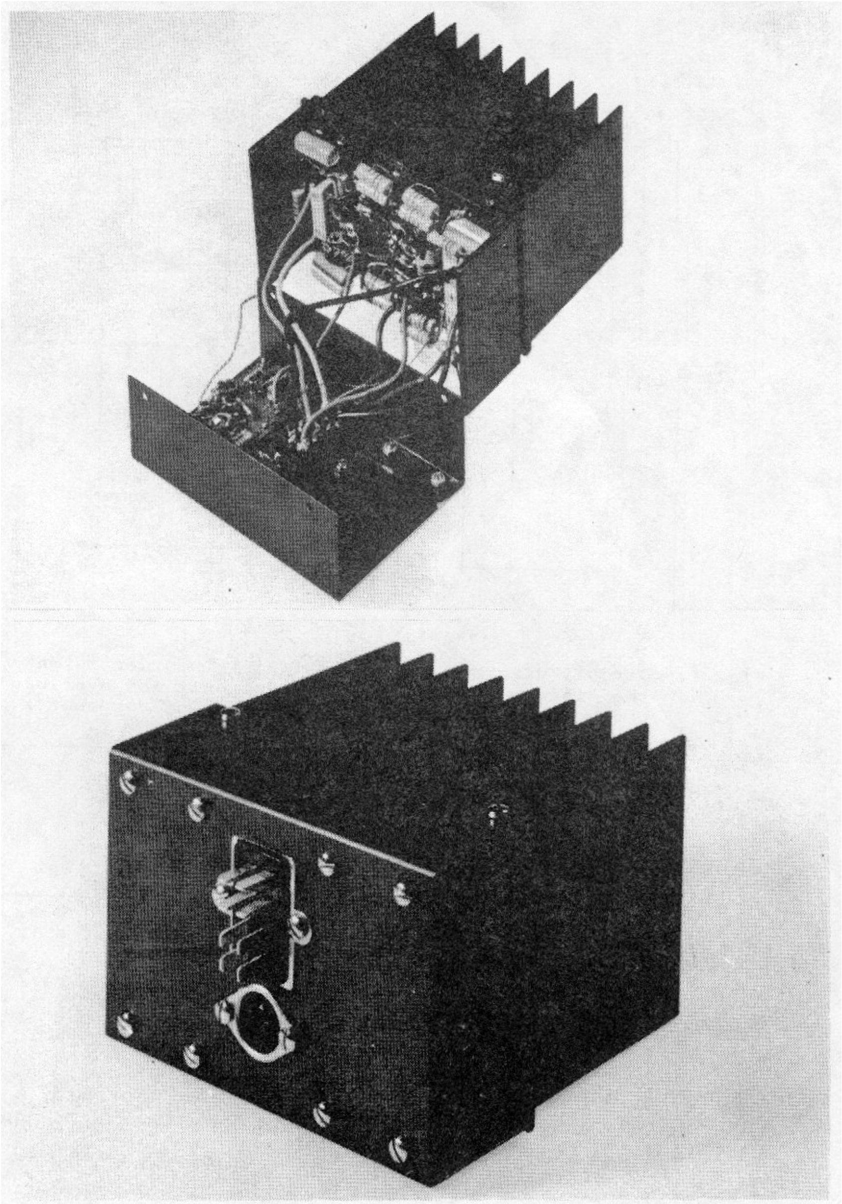


Fig. 2. The mechanical construction of the power amplifier. Transistors T_7-T_{17} are directly mounted on the heat-sink, which has 0.75°C/W thermal resistance to the ambient. The maximum total power dissipation is about 60 W. The input stages are placed on a small printed circuit board which is mounted in the cover. Amplifier dimensions are : 7.5 x 11.7 x 12.0 cm.

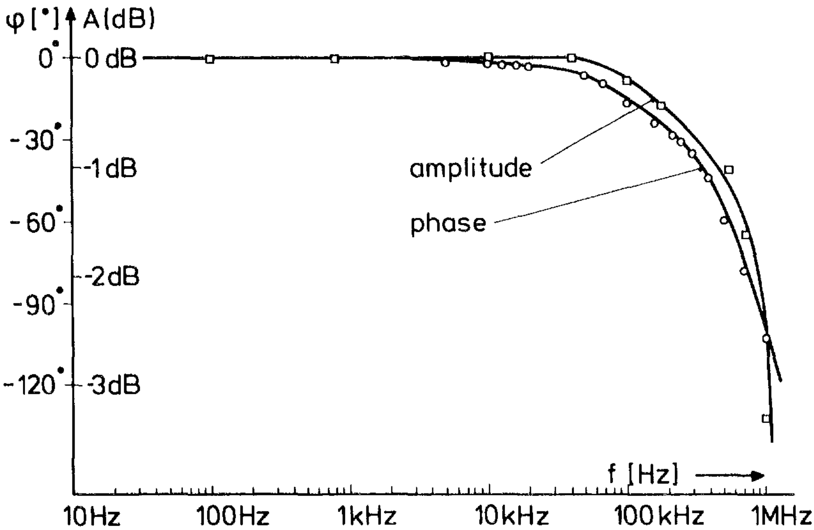


Fig. 3. Amplifier small-signal frequency and phase responses. The -3 dB point is 1 MHz. At this frequency the phase shift is 100°. About 45° is caused by a first-order rolloff, and 55° by a propagation delay of approximately 150 ns.

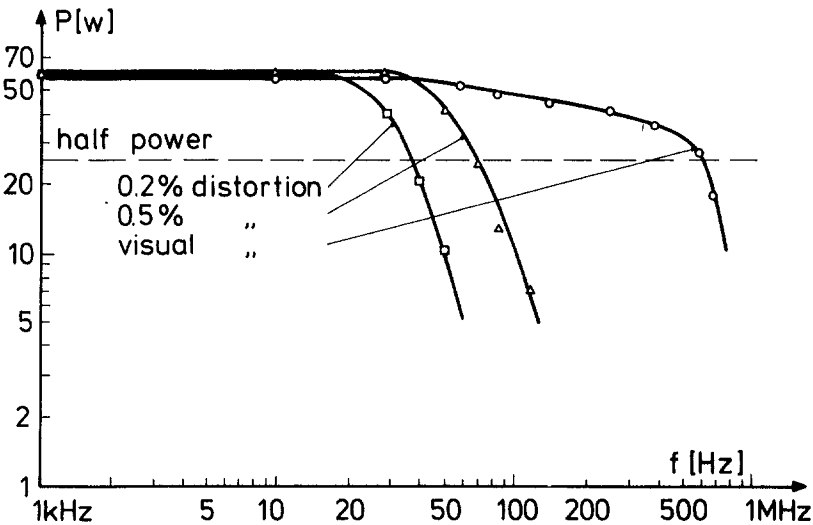


Fig. 4. Equi-distortion curves: power versus frequency for three different criteria. For the proposed criterion of 0.2% total harmonic distortion the amplifier has a power bandwidth of 35 kHz (half power point).

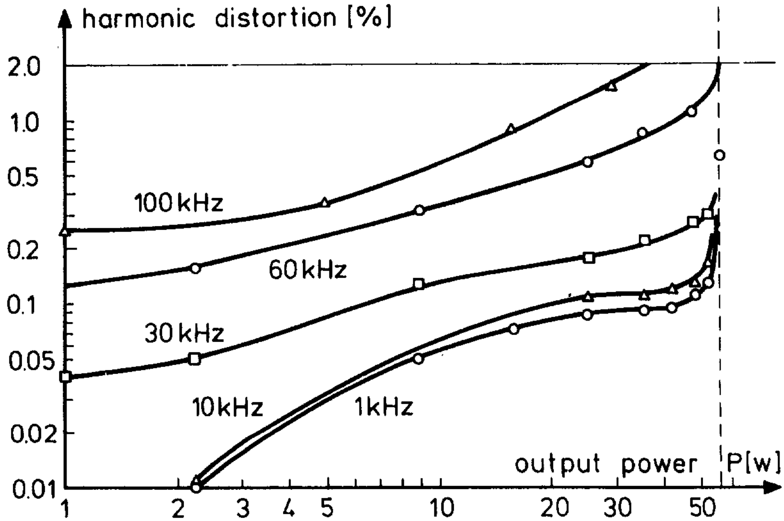
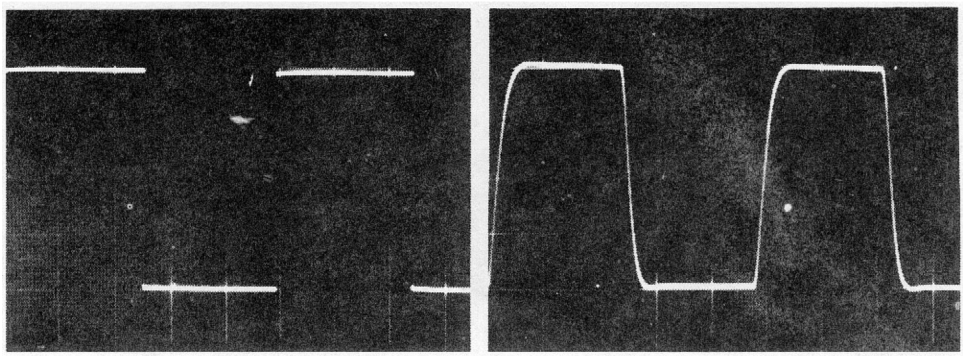


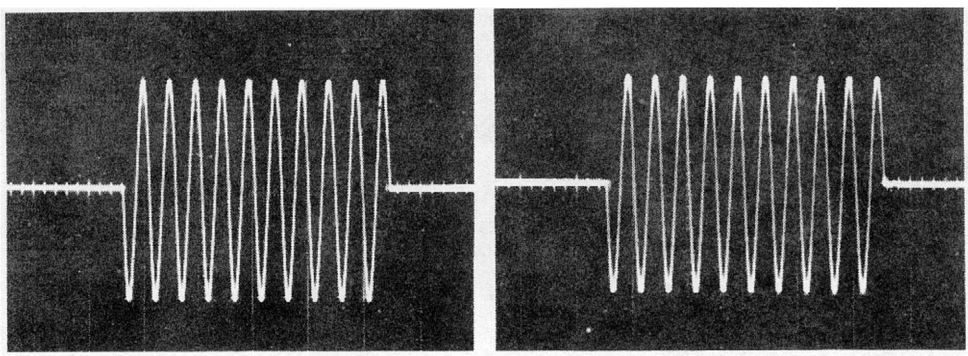
Fig. 5. Total harmonic distortion as a function of power for different frequencies. The distortion at low power levels for frequencies up to 10 kHz is infinitesimally small. No traces of cross-over distortion can be found due to class A operation in the low power range.



a) b)

Fig. 6. a) Amplifier square wave-response at 1 kHz, 20 V peak-to-peak output, 4Ω load.
Horizontal: 200 μs/div., vertical: 5 V/div.

b) Amplifier square wave-response at 100 kHz, 20 V peak-to-peak output, 4Ω load.
Horizontal: 2 μs/div., vertical: 5 V/div.



a) b)

Fig. 7. a) Burst tone test for 1 kHz sinusoidal burst, 20 V peak-to-peak output, 4Ω load.
Horizontal: 2 ms/div., vertical: 5 V/div.

b) Burst tone test for 100 kHz sinusoidal burst, 20 V peak-to-peak output, 4Ω load.
Horizontal: 20 μs/div., vertical: 5 V/div.

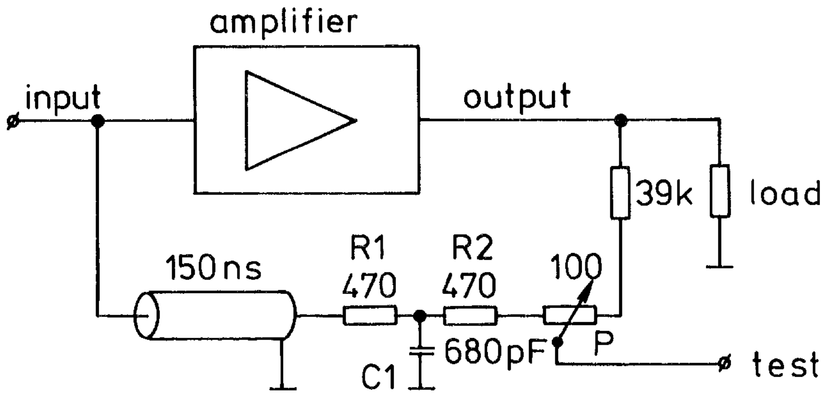


Fig. 8. Compensation test. Because the amplifier is an inverting one, a position can be found of potentiometer P where the linear component of the output signal exactly compensates the input signal at the "test" point. The momentary discrepancy signal, as scaled up to the output level, was below 0.01% for powers below 5 W and for frequencies in the audio range, and was mainly caused by incomplete phase compensation (R_1, R_2 and C_1).