

Transient Distortion in Transistorized Audio Power Amplifiers

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Abstract

This paper discusses a new kind of distortion mechanism found in transistorized audio power amplifiers. It is shown that this distortion arises from the multistage feedback loop found in most high-quality amplifiers, provided that the open-loop transient response of the power amplifier is slower than the transient response of the preamplifier. The results of the analysis are verified by measurements from a simulated power amplifier, and a number of constructional rules for eliminating this distortion are derived.

Introduction

An ordinary transistorized audio amplifier consists of a preamplifier and a power amplifier. The typical preamplifier incorporates two to eight stages with local feedback. The power amplifier has, however, usually a feedback loop enclosing three to four stages. The power amplifier generally determines the frequency response and the distortion of the whole amplifier.

For stationary signals, the harmonic distortion of the power amplifier decreases proportionally with increasing feedback, provided that the transfer function of the amplifier is monotonically continuous and that the gain is always greater than zero. (These assumptions are not valid, of course, in case of overload or crossover distortion.) With the same assumptions, the intermodulation distortion decreases similarly. The frequency response is also enhanced in proportion with the feedback.

It would seem, then, that feedback is highly beneficial to the power amplifier. The purpose of this paper is, however, to show that the usable frequency response of the amplifier does not necessarily become better due to feedback, and that, under certain circumstances, the feedback can cause severe transient distortion resembling intermodulation distortion.

These facts are well known among amplifier designers and have been discussed on a phenomenological basis (for instance [1]). They have not, however, received a quantitative treatment except in some special cases [2], [3].

Transient Signals in Amplifiers

Sound in general, and especially music, consists largely of sudden variations. The steep rise portion of these transient signals can be approximated with a unit step function, provided that the transfer functions of the microphone and the amplifiers are considered separately. We may, therefore, divide the amplifier as in Fig. 1. A is the preamplifier including the microphone, C is the power amplifier, and B is the feedback loop around it.

If resistive feedback is to be applied in the power amplifier, stability criteria necessitate its transfer function to have not more than two poles and a single zero in the usable frequency range. The transfer function without feedback can thus be approximated to be of the form

$$F_c(s) = A_0 \frac{s\omega_1}{(s + \omega_0)(s + \omega_1)} \quad (1)$$

where A_0 is the midband gain without feedback, and ω_1 and ω_0 are the upper and lower cutoff angular frequencies, respectively.

The transfer function of the signal source and the preamplifier can be arbitrary. Usually, however, it can be considered as having several poles and zeros, often multiple. In the following we will consider two special cases.

Case a: The transfer function is flat in the midband and has a -12 dB per octave rolloff in both the high-frequency

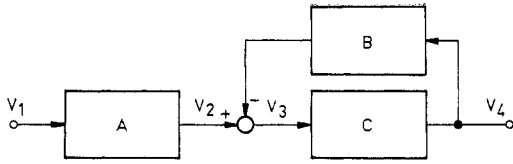
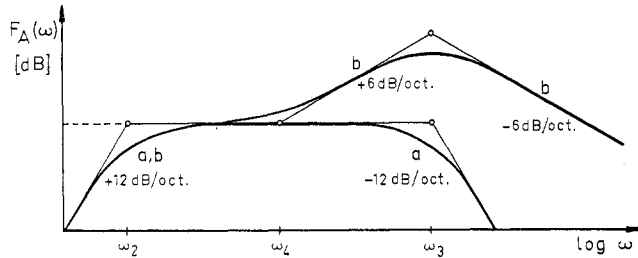


Fig. 1. The analyzed circuit. A is the preamplifier which includes the transfer function of the signal source. B is the feedback path around the power amplifier C.

Fig. 2. The preamplifier frequency response asymptotes used in the analysis. Asymptote *a* corresponds to a flat response and asymptote *b* corresponds to a case where the high-frequency tone control has been turned to maximum.



and the low-frequency ranges. This characteristic is shown in Fig. 2 with asymptote *a*.

Case b. The transfer function in the low-frequency range is similar to Case *a*. A +6 dB/octave emphasis is applied in the high-frequency range starting at an angular frequency ω_4 and resulting in asymptote *b* in Fig. 2.

These two cases are, of course, arbitrary, but are considered as being representative: the first for the flat response case, and the second, for the worst case where the high-frequency tone control has been turned to maximum.

The transfer functions of the preamplifier are then for Case *a*

$$F_A(s) = \frac{s^2\omega_3^2}{(s + \omega_2)^2(s + \omega_3)^2} \quad (2)$$

and for Case *b*

$$F_A(s) = \frac{s^2\omega_3^2(s + \omega_4)}{(s + \omega_2)^2(s + \omega_3)^2\omega_4} \quad (3)$$

where ω_2 and ω_3 are the low-frequency and high-frequency cutoff angular frequencies, respectively, and ω_4 is the break frequency of the tone control.

Supposing the input voltage of the preamplifier to be a step function

$$V_1(s) = \frac{v_p}{s}$$

and the feedback path *B* to be resistive, the output voltage of the power amplifier is in the frequency domain for Case *a*

$$V_4(s) = \frac{A_0 s^2 \omega_1 \omega_3^2 v_p}{[(s + \omega_0)(s + \omega_1) + \beta A_0 s \omega_1](s + \omega_2)^2(s + \omega_3)^2} \quad (4)$$

and for Case *b*

$$V_4(s) = \frac{A_0 s^2 \omega_1 \omega_3^2 (s + \omega_4) v_p}{[(s + \omega_0)(s + \omega_1) + \beta A_0 s \omega_1] \omega_4 (s + \omega_2)^2 (s + \omega_3)^2} \quad (5)$$

where β is the gain of the feedback path.

With the same assumptions the input voltage within the feedback loop is, for Case *a*,

$$V_3(s) = \frac{(s + \omega_0)(s + \omega_1) s \omega_3^2 v_p}{[(s + \omega_0)(s + \omega_1) + \beta A_0 s \omega_1](s + \omega_2)^2(s + \omega_3)^2} \quad (6)$$

and for Case *b*,

$$V_3(s) = \frac{(s + \omega_0)(s + \omega_1) s \omega_3^2 (s + \omega_4) v_p}{[(s + \omega_0)(s + \omega_1) + \beta A_0 s \omega_1] \omega_4 (s + \omega_2)^2 (s + \omega_3)^2} \quad (7)$$

In high-quality amplifiers the following is generally true:

$$\omega_0 \ll \omega_1 \quad \omega_2 \ll \omega_3.$$

Simplifying (4)–(7), accordingly, and adopting new normalized variables,

$$1 + \beta A_0 = \alpha, \omega_1/\omega_3 = \gamma, \omega_4/\omega_3 = \epsilon, \text{ and } \omega_3 t = T,$$

the inverse transforms of (4)–(7) will be in the time domain for Case *a*

$$v_4(T) = \frac{A_0 v_p}{\alpha} \left\{ 1 - \frac{e^{-\alpha \gamma T}}{(1 - \alpha \gamma)^2} + \left[\frac{\gamma T}{1 - \alpha \gamma} - \frac{\gamma(\alpha \gamma - 2)}{(1 - \alpha \gamma)^2} \right] \alpha e^{-T} \right\} \quad (8)$$

$$v_3(T) = \frac{v_p}{\alpha} \left\{ 1 - \frac{(\alpha - 1)e^{-\alpha \gamma T}}{(1 - \alpha \gamma)^2} + \left[\frac{(\gamma - 1)T}{1 - \alpha \gamma} - \frac{1 + \gamma(\alpha \gamma - 2)}{(1 - \alpha \gamma)^2} \right] \alpha e^{-T} \right\} \quad (9)$$

and for Case *b*

$$v_4(T) = \frac{A_0 v_p}{\alpha} \left\{ 1 + \frac{\alpha \gamma - \epsilon}{\epsilon(1 - \alpha \gamma)^2} e^{-\alpha \gamma T} + \left[\frac{\gamma(\epsilon - 1)T}{\epsilon(1 - \alpha \gamma)} - \frac{\gamma(1 - 2\epsilon + \epsilon \alpha \gamma)}{\epsilon(1 - \alpha \gamma)^2} \right] \alpha e^{-T} \right\} \quad (10)$$

$$v_3(T) = \frac{v_p}{\alpha} \left\{ 1 + \frac{(\alpha - 1)(\epsilon - \alpha \gamma)}{\epsilon(1 - \alpha \gamma)^2} e^{-\alpha \gamma T} + \left[\frac{(\gamma - 1)(\epsilon - 1)T}{\epsilon(1 - \alpha \gamma)} - \frac{(\alpha \gamma - 1)(\gamma \epsilon - 1) - (\gamma - 1)(\epsilon - 1)}{\epsilon(1 - \alpha \gamma)^2} \right] \alpha e^{-T} \right\} \quad (11)$$

The values of these equations have been computed and the results are shown in Figs. 3–8. It can be seen very clearly that the feedback loop input voltage v_3 includes an overshoot in both cases of preamplifier transfer function. In Case *b* this voltage also includes a negative overshoot.

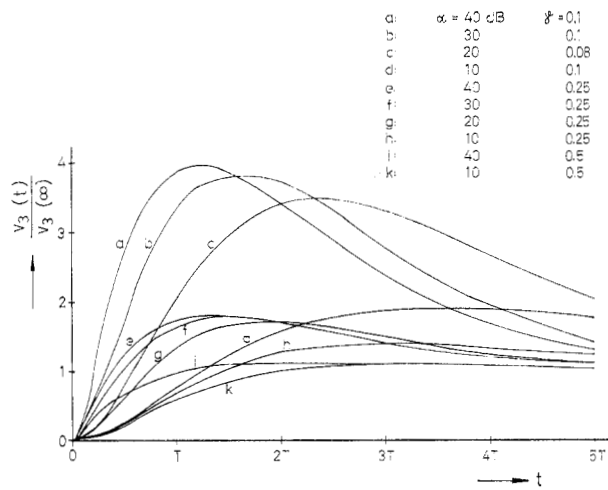


Fig. 3. The normalized input voltage $v_3(t)$ of the power amplifier as a function of the normalized time for several sets of parameters. Case a.

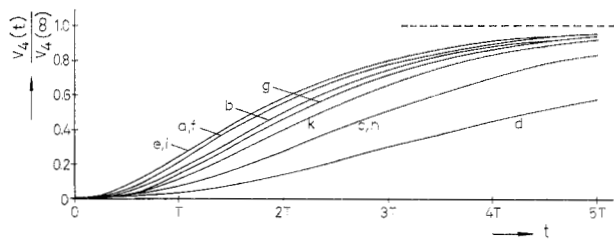


Fig. 4. The normalized output voltage $v_4(t)$ of the power amplifier as a function of the normalized time for several sets of parameters. Case a.

Fig. 5. The maximum value of $v_3(t)$ divided by the final value as a function of the cutoff frequency ratio γ . Case a.

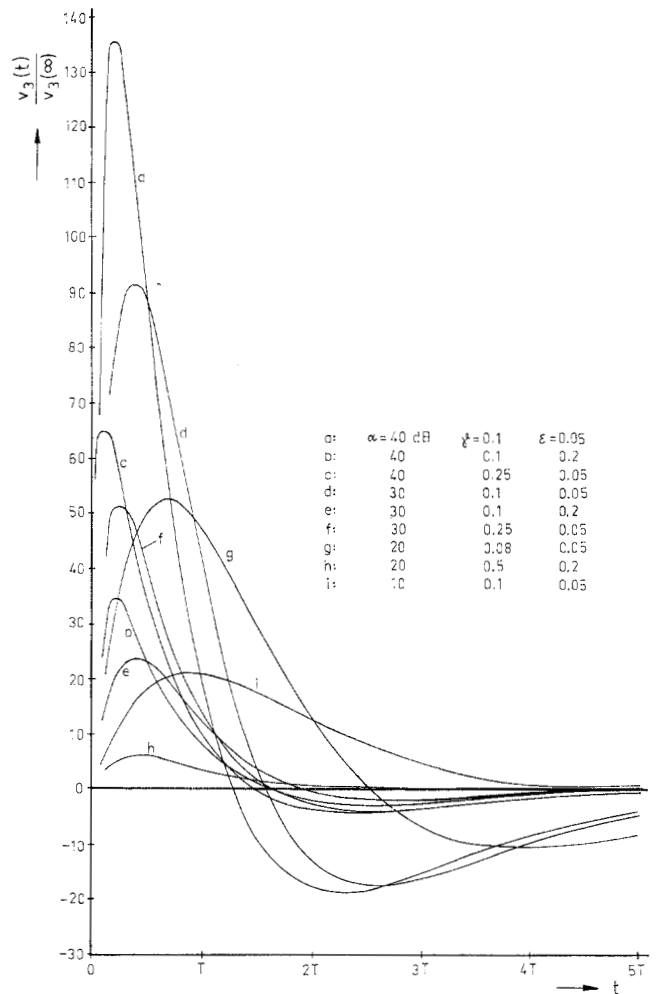
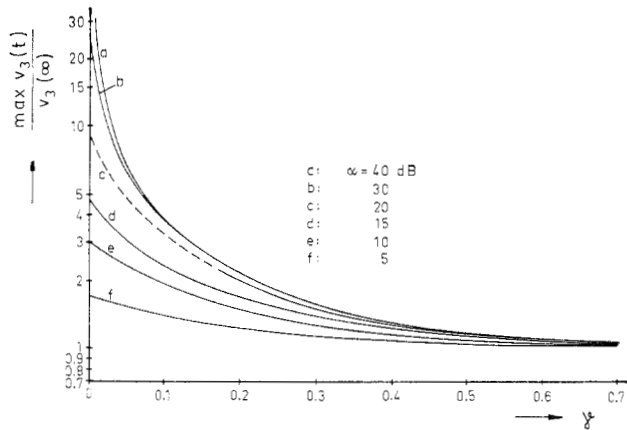
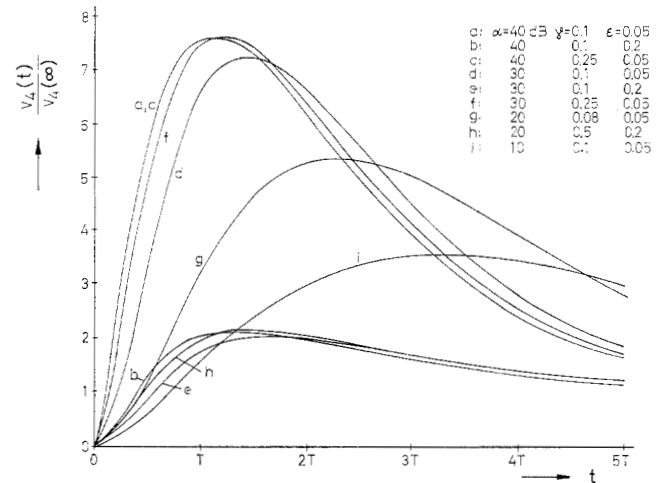


Fig. 6. The normalized input voltage $v_3(t)$ of the power amplifier as a function of the normalized time for several sets of parameters. Case b.

Fig. 7. The normalized output voltage $v_4(t)$ of the power amplifier as a function of the normalized time for several sets of parameters. Case b.



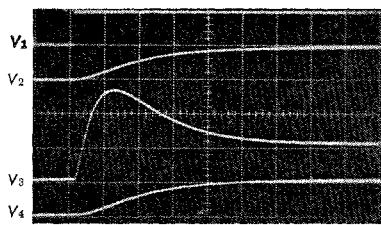


Fig. 10. Measured voltages of the amplifier with 30-dB feedback and cutoff frequency ratio $\gamma=0.13$ for Case a. All the voltages have been normalized so that the difference between the initial and the final value is 1 division. Horizontal scale: 1 division is 1.5 T.

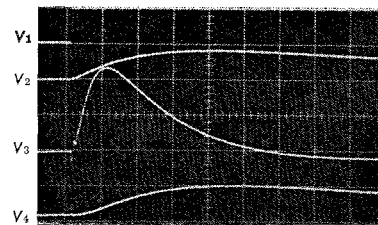


Fig. 12. Measured voltages of the amplifier with 30-dB feedback and cutoff frequency ratio $\gamma=0.1$ for Case b. $\epsilon=0.05$. Vertical scales: the difference between the initial and the final values. $v_1:1$ division, $v_2:0.2$ division, $v_3:0.03$ division, $v_4:0.25$ division. Horizontal scale: 1 division is 0.3 T.

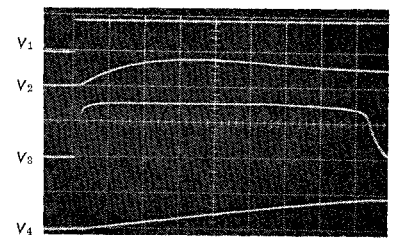


Fig. 14. The voltages of Fig. 11 with limiting with Zener diodes D_1 and D_2 at a level of $3v_4$ (∞). The scales are the same as in Fig. 11 except for $v_3:0.5$ division.

Fig. 11. Measured voltages of the amplifier with 30-dB feedback and cutoff frequency ratio $\gamma=0.5$ for Case b. $\epsilon=0.05$. Vertical scales: the difference between the initial and the final values. $v_1:1$ division, $v_2:0.2$ division, $v_3:0.2$ division, $v_4:0.4$ division. Horizontal scale: 1 division is 0.3 T.

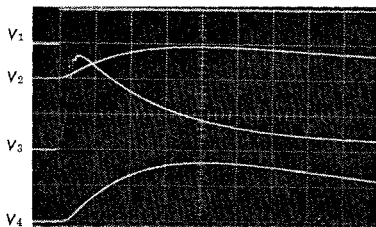


Fig. 13. The voltages of Fig. 10 with limiting with Zener diodes D_1 and D_2 at a level of $1.30 v_4$ (∞).

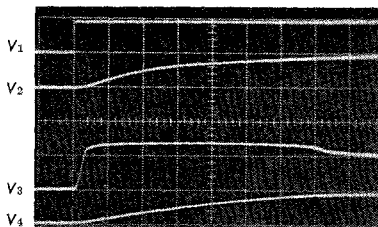
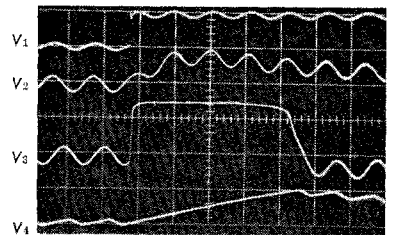


Fig. 15. A demonstration of the intermodulation distortion caused by the transient signal. The input to the amplifier is a sinusoidal signal and a transient. Due to transient limiting the sinusoid is clipped in v_3 and v_4 . This is equivalent to a momentary 100 percent intermodulation distortion. The parameters and vertical scales are as in Fig. 14. The horizontal scale: 1 division is 0.5 T. Note that in this figure the sinusoid has a very high frequency for the purpose of clarity. The level of it has therefore been increased when photographing v_3 and v_4 .



ments in the bias circuitry, for instance. The cutoff frequencies of signal sources and preamplifiers are usually much higher, 20 kHz to 100 kHz being typical. A typical value of γ may, therefore, lie somewhere between 0.1 and 1. For these values the overshoots may range from zero to several hundred times the nominal feedback loop input voltage.

Most present-day transistorized high-quality amplifiers are of the transformerless complementary symmetry type. Though simple in principle, this type of amplifier includes several distortion mechanisms. The minimization of the distortion requires careful design. With regard to transient distortion this type of amplifier is disadvantageous, because the upper cutoff frequency depends on the cutoff frequency of the power transistor current amplification or on the usual lag compensation circuit in the collector of the driver transistor. In both cases the overdrive condition with stationary signals is usually reached at the same instant in both the power transistors and the driver transistor. With the transient signal described above, the driver transistor reaches the overdrive condition first due to the overshoot of v_3 . This results in the clipping of the overshoot which, in the following pages, will be called the

transient distortion. In the simulated amplifier of Fig. 9 the clipping may be accomplished with Zener diodes D_1 and D_2 .

The Effect of the Transient Distortion

If the overshoot is clipped, the gain of the amplifier for other signals occurring at the same time is zero. This results in 100 percent momentary intermodulation distortion as can be seen in Figs. 13–15. It is obvious that the transient distortion perception is due to this intermodulation, as the increase of the transient rise time due to clipping can hardly cause anything other than a slight alteration of the transient sound “color”. Small-scale listening experiments seem to support this theory.

If the intermodulation is to be avoided with a given amplifier, the input signal must be made smaller in order to prevent the transient clipping. As the output power of the amplifier is proportional to the square of the input signal, it is easy to see that the distortionless output power of the amplifier will often be only a fraction of the nominal.

The overshoot is strongly dependent on the feedback

when $\gamma < 1$. The transient distortion thus increases with increasing feedback. As the harmonic distortion and the usual intermodulation distortion decrease with increasing feedback, there will be an optimum value for the feedback producing minimum subjective distortion sensation. This optimum value is dependent on the type of signal to be amplified. It must, however, be realized that the ordinary distortion measuring methods do not reveal transient distortion.

Conclusions

The results justify the following conclusions in the case of strong feedback and low value of upper cutoff frequency ratio γ .

- 1) To minimize the transient distortion it is advantageous to let the preamplifier limit the frequency response of the complete amplifier.
- 2) In the above case, it is the power amplifier frequency response *without feedback* that determines the desired pre-amplifier frequency response. Then feedback in the power amplifier does not necessarily enhance the usable frequency response of the complete amplifier.
- 3) If high fidelity reproduction requires a 20-kHz upper cutoff frequency in the amplifier, the power amplifier should reach it *without feedback*.
- 4) In the above case, the upper cutoff frequency of the power amplifier *with feedback* must be at least 20 kHz times the feedback loop gain. (For example, for an amplifier with 30-dB feedback, 630 kHz, 40-dB feedback, 2 MHz.)

- 5) There exists an optimum feedback for the amplifier. Any substantial alteration of the feedback from this value will increase the subjective distortion sensation.
- 6) The usual lag compensation techniques for obtaining stable feedback amplifiers are inadvisable. By reducing the upper cutoff frequency of the power amplifier this increases transient distortion.

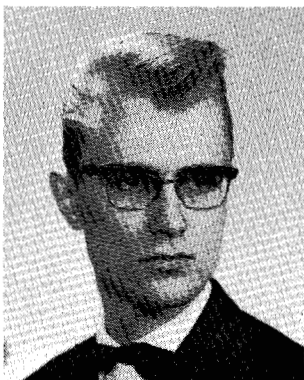
The above conclusions are valid within the validity of the approximations made in the previous analysis. They hold good for transistorized complementary symmetry audio power amplifiers especially when "slow" power transistors are used. The transient distortion can be minimized by using RF power transistors, local feedback in the power transistors (unbypassed emitter resistors), low value of generator impedance seen by the power transistor bases, and by dimensioning the feedback loop correctly.

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References

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