An approach to audio amplifier design

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First of a series of three articles in which the fundamentals of audio amplifier design are re-examined, taking account of recent studies in psycho acoustics and circuit techniques. A recent design will be discussed and some experiments related.

In 1883 Lord Kelvin wrote, "I often say that when you can measure what you are speaking about, and express it in numbers, you know something about it; but when you cannot express it in numbers your knowledge is of a meagre and unsatisfactory kind; it may be the beginning of knowledge but you have scarcely in your thoughts advanced to a stage of science whatever the matter may be."

The major difference between a science and any other area of knowledge and thought is that with science, semantic errors can be avoided by reducing all concepts to a numerical form which can give a universally understood meaning, and, most important, allow a value to be predicted which can be experimentally verified.

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In evolving any theory the investigator is always left with the problem of isolating concepts and parameters, and where there is unusual complexity, as for example in interactions involving human beings, it is made difficult because of the number of ideas that must be involved, and the extreme numerical range that any investigation must produce.

The state of affairs that exists in audio design is that, although certain aspects of performance can be totally described, there is no accepted method which describes the overall performance as judged by the listener.

The ideal situation is one in which the complete audio chain, be it microphone to ear or perhaps record to ear, can be given a figure of merit which relates to its acceptability by a percentage of the population (Fig. 1). However, the problems here are many, not the least that this figure of merit may be time variable due to overall rising standards.

In an earlier article[1] I put forward the idea that subjective sound quality should be considered in terms of things going wrong —that is, a measure of the unpleasantness determined from a weighted sum of critical parameters.

It is fairly well accepted that overall sound quality is not equally disturbed by all the possible shortcomings and it is also accepted that there is a threshold below which a particular shortcoming may not be noticed, at least until one of the others has been improved.

These notions are of fundamental importance to the production of an effective design method, and the implications are that:

1. Linearity and hence superposition cannot be assumed in discussing degrees of aural unpleasantness.
2. The necessity for a compromise of subjective ideals, due to engineering limitations, results in the need to optimize all the parameters in a way that may not coincide with their individual maxima or minima.

In Fig. 2 I have redrawn the simple model which relates sound quality to the mean recording level in a tape recorder; here undesirable effects arise at low levels from noise and at high levels from progressive overloading. The model illustrates intuitively the way in which a trade-off is made and how the best result does not coincide with minima of the dependent variables.

Consider for a moment the record-playing chain of Fig. 3; here some of the variables affecting the final musical impression are isolated. The impression, apart from artistic considerations which can be dominant or destructive, depends on the passing of years t, temperature T, the quality of transduction by the cartridge and its impedance, tracking weight, mass and compliance of stylus. Also included is the amplifier transfer function, loudspeaker transfer function, the absolute level of the signal in the amplifier (e) and loudspeaker (E and P), the room acoustics, sound pressure level, mood m and disposition towards the listening event D. All of this is confused by the fact that the sensitivity to shortcomings in the system or its components is not constant with any individual, or between individuals. Knowing

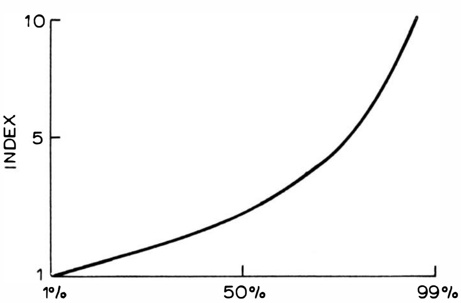

Fig. 1. Percentage population acceptability for more than 95% of the time.

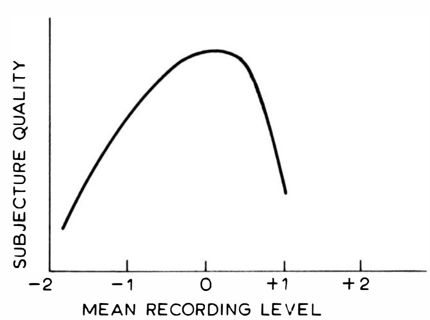

Fig. 2. Subjective quality as a function of mean recording level.

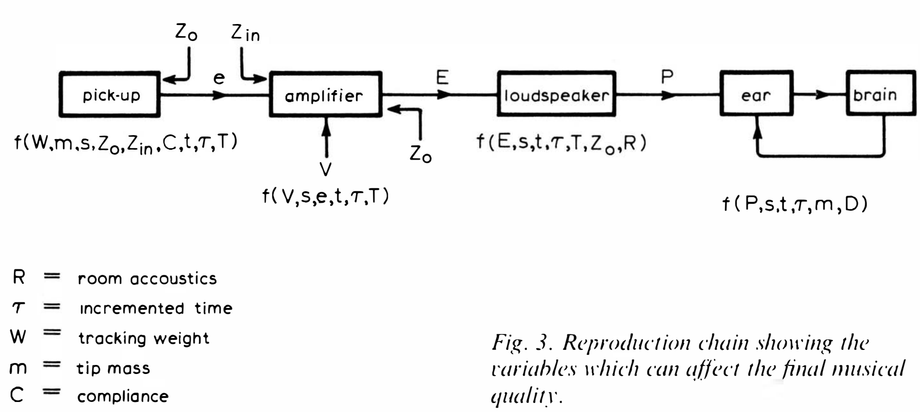

Fig. 3. Reproduction chain showing the variables which can affect the final musical quality.

the techniques of mathematical programming, it is possible to make useful analyses and predictions in problems of just this complexity. In a practical situation there will be a set of constraints which can demonstrate the need for a trade-off between different levels of unpleasant result. Thus a balanced result is obtained from an objective function $O(z)$, which is minimized using the empirical weightings C_α, C_β etc. This is shown in Fig. 4 in a general form. However, it would seem that the real problem is not the availability of tools to produce a design but a serious lack of psychoacoustic data and the consequent agreement on what aspect of it is important.

Constraints	Variables α	β	γ	δ
$A =$	a_{11}	$+ a_{12}$		
$B =$		b_{22}	$+ b_{23}$	$+ b_{24}$
$C =$		c_{32}		$- c_{34}$
etc.				
Min. $O(z) =$	C_α	$+ C_\beta$	$+ C_\gamma$	$+ C_\delta$

Fig. 4 General form of analysis.

When a designer is faced with the problem of producing an amplifier to a price,the most important facts to be established commence with the broad defining specification, and then there is a statement about the trade-off of distortions and other parameters in the chosen configuration. One could perhaps say that total harmonic distortion $D\%$ will reduce according to cost £Z in the following way:

$$Z = a \exp [Q.(D)^{-1}] + b.D^{-2} + c.D^{-1}$$

where a and Q relate to component cost, b to testing and c to production. Similar relationships could be proposed and tested for all parameters, and interaction analysis will show an overall cost-performance relationship which can in turn be applied to known percentage population preferences. A preference function $p(D)$ representing, for example, the probability that $D\%$ of distortion is detectable by a random population selection could be tested starting out with the form

$$p(D) = \alpha \exp (D-y) \quad D \leqslant y$$

A starting point

So far it has been suggested that a scientific approach is needed to establish for an audio system a figure of merit which can be related to the subjective reaction. Whilst showing that a very complete analysis can be achieved, provided that the correct information is selected and applied, the problems of complexity and variability remain associated with such a project.

It seems that the only road to a useful figure of merit is to accept the concept of "collective subjectivity" as factual and then to attempt to isolate its parameters and effects, assigning, as far as possible, measures of significance.

For example, it would seem reasonable to assume that the first two propositions to establish when discussing any one parameter, e.g. noise, are the level at which it becomes perceptible and the level at which it becomes objectionable—or impossible to neglect. Further work can then substantiate or challenge these results and in addition improve the accuracy of the curve fitting.

In these articles I discuss known parameters relating to amplifier design which can be of significance, attempting to assign to them a degree of importance based on my own work and the work of others. A more complete discussion of the figure of merit concept and a recent design experiment follow. It is not my intention to propose a finalized quality rating, but rather to make a few steps in this direction. At the same time I will point out how such a rating may be derived, in the hope of encouraging new work and discussion on this subject.

System considerations

In contemplating the reproduction of music the ideal is that the sound field, as perceived by the listener, should approach as closely as possible the original event, or at least the balance engineer's version of it.

To recreate a sound field it is necessary to produce all the essential detail of the original acoustic waveform at the appropriate loudness. Now it does seem that an accurate recreation is impossible using loudspeakers, even if they have ideal distribution characteristics. However, it is not possible or necessary in this discussion to consider reverberant sound fields set up by loudspeakers in rooms or the special problems of two or four channel systems.

Consider the problem in its simplest form; the original event is picked up, say, using the dummy-head microphone technique and conveyed through a system to a pair of headphones. In this chain there will be two or four electro-mechanical or electro-acoustic transducers possibly exhibiting non-minimum phase characteristics, resonances etc. It is also possible that the amplifier blocks and other links in this chain will have a historical design approach.

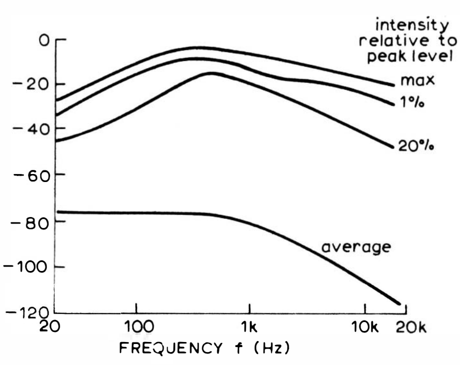

Fig. 5. The energy distribution (in arbitrary units) in an extended musical event.

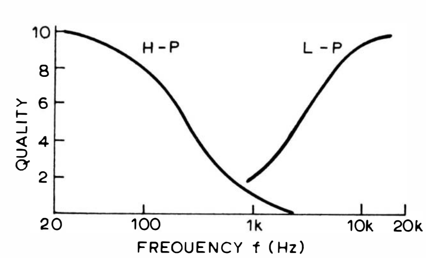

Fig. 6. The effect of frequency range upon the reproduced quality of music.

It seems clear that the criteria will be common for any part of the chain, namely to preserve as far as possible the integrity of the original signal. This implies that the fundamental design criteria be first determined and then applied to every element to ensure success.

Such a conclusion allows a more specific concentration on single elements in the chain, in the knowledge that the general principles derived will be applicable in all instances, provided the correct assumptions are made.

Amplifier design

The current attitude to audio amplifier design is reflected in the DIN 45500 standards, and an amplifier which nowadays would be considered to be very good will have a specification as follows:

1. Output power in excess of 40W each channel.
2. Power bandwidth 20Hz–30kHz±1dB.
3. Very low noise and hum, say −80dB.
4. Total harmonic distortion less than 0.1% at all frequencies and power levels in the bandwidth.
5. Intermodulation distortion, however measured less than 0.1%.
6. Low output impedance, say 400mΩ.

The starting point of a truly "scientific" design approach should be to accept the existing requirements, note the areas of weakness and if necessary build up a new design hypothesis.

For me, the practical starting point is that, say, ten amplifiers of different design all with the above specification when compared in a listening test show serious qualitative differences. Given this situation we are now interested in establishing the nature of the differences and from that evolving a figure of merit.

The bandwidth of the ear under the best possible conditions is generally a maximum of 22kHz and since musical events are known to have energy distributions as shown in Fig. 5 it seems reasonable that a system bandwidth of 20Hz–22kHz should be considered sufficient, together with an amplitude response within 0.5dB over this range. Snow[2] described experiments showing how the quality rating he had evolved varied with bandwidth (shown in Fig. 6). From his experimental results, limitation of the bandwidth became objectionable during the whole test cycle when a low frequency cut-off of 1kHz was applied.

It may be that a quality rating based solely on this one parameter is inadequate, but, as I hope to show later, the value of quoting bandwidth as ≈22kHz or >22kHz *per se*, is limited. What can be of overriding importance is the origin of the bandwidth limitation.

A cornerstone of the theory of sound reproduction has been Ohm's Auditory Law which states that the ear tends to analyse the components of a complex sound regardless of their phase relationships. Thus the ear is inclined to operate as an on-line Fourier analyser and this transformation of the waveform is considered to be adequate information.

Twenty years ago the specification for an audio amplifier would suggest that it should amplify all the frequencies of a musical signal equally, without adding any new frequencies. This is, if you like, the credo of the design philosophy based on frequency response and includes the notions of total harmonic distortion (t.h.d.) and intermodulation distortion (i.m.d.).

However, it seems that the "frequency response" viewpoint is very constraining since if one starts out with an idea set in a single frame of reference—in this case the $j\omega$ plane—it is easy to lose sight of the objective. We do not necessarily want to amplify all the frequencies of music equally—especially without regard to phase. What is required is to amplify an audio waveform of acoustic origin in such a way that the ear can detect no degradation.

For many years it has been accepted in audio engineering and psycho-acoustic circles that the ear-brain combination does not perform this frequency analysis in the way Ohm suggested; but rather analyses in terms of the waveform. It has been shown [4,5,6] that the qualitative characteristics of a complex sound depend on the phase relationships of the component harmonics. In fact, more recent work has made it clear that the ear has very specific sensitivities to waveform differences [7,8].

It may be thought that if an amplifier has a response $|F(j\omega)|$ = constant, between 20Hz and 20kHz then it will automatically reproduce all waveforms correctly; however, this is not a sufficient performance description. It is also necessary that the system be minimum phase, making it necessary to eliminate certain all-pass networks in common use. Helmholtz was the first to say that the "quality of musical perception of a complex tone depends solely on the number of partial tones and in no respect on their difference in phase".[9]

As a phase difference must be interpreted as a time delay between the component parts of a signal, it is clear by induction that sufficient phase shift in a system must eventually become audible as a result of moving these components with respect to each other in time. This can be deduced from:

1. The ear's ability to differentiate small time and amplitude differences as confirmed by directional acuity.

2. In practice such large phase shifts as occur in long telephone lines, render speech unintelligible unless phase and delay correction is introduced[10, 14].

In addition, recent experimental findings by Madsen[8] are summarized as follows.

1. The ear is sensitive to phase differences between frequency bands.

2. The sensitivity threshold is raised by a factor of three in reverberant surroundings where the sound source is a loudspeaker compared with results obtained using headphones.

3. The ear seems to prefer the frequency content of negative pressure transient wavefronts showing the significance of absolute phase.

4. In listening room conditions using a carefully constructed test signal a 10° phase shift between extreme frequencies was detectable.

Stodolski[7] suggested that an audio system which maintains a 3dB tolerance in amplitude-frequency response should also maintain a 17° tolerance in phase shift; he also showed that a 180° absolute phase error is aurally equivalent to 11.5% intermodulation distortion.

Now whilst it is relatively simple for an amplifier designer to achieve a maximum phase shift of 1° in the audio band with conventional parameters being considered, when I discuss some further aspects of musical realism I will show that it is a more complex problem than that; in addition all sorts of questions are raised about tone controls and filters.

On the basis that it is better to over- rather than under-estimate the acuity of the ear, it seems reasonable in the face of so much experimental evidence to agree that a figure of merit concept should also contain a measure of both phase deviation and phase smoothness. The only remaining problem is to propose the perceptual thresholds.

These arguments tend to convey that the quality of reproduction is principally affected by the accuracy with which the original acoustic waveform is recreated at the ear, but this is a point to return to.

Linear theory shows that in minimum-phase systems the steady-state function $F(j\omega)$ is related to $f(t)$ by the Laplace transform in a specific and simple way. The transfer function of an amplifier is said to be linear when complete correspondence exists between input and output and an important consequence of linearity is that superposition can be held as true.

It is customary and convenient to measure any departure from linearity as the extent to which new frequency components appear in the output of an amplifier, excited by n sinusoids where $n \geqslant 1$. The resulting measurement which is conventionally the r.m.s. sum of these new frequencies will be either t.h.d. for $n = 1$ or i.m.d. $n > 1$.

In 1947 it was suggested[12] that a good design objective was a maximum of 0.1% harmonic distortion since, first, it represented a readily achievable goal which was better than supposed necessary and, second, it left room for a deterioration of performance in service. (It should be pointed out that this objective referred to class A amplifiers using tetrode valves and having a moderate amount of negative feedback.) This level of performance would appear to be high, and in the light of other published work there is no ground for dismissing it. Olsen[13] showed that for reproduced music in a 15kHz bandwidth the levels of distortion necessary to produce the reactions perceptible, tolerable and objectionable were 0.75%, 1.8% and 2.4% respectively, in a system producing predominantly second-harmonic distortion.

However, no one can now suggest that 0.1% t.h.d. is a criterion by which the goodness of an amplifier can be judged: one only has to listen to a signal containing 0.1% 7th or 9th harmonic to realise that this is definitely audible. More recent investigation has shown that the ear is more sensitive to distortions according to their order, that is, 0.1% third harmonic is more significant than 0.1% second, and so on.

D. E. L. Shorter suggested[14] that the best correlation between objective and subjective tests on the order of harmonic distortions was obtained using the weighting $n^2/4$, thus the fifth harmonic would be 6.25 times as significant as the second harmonic. On the other hand, in a very thorough investigation Wigan[15] suggested that a distortion criterion C_t would be better defined as:

$$C_t = \sum_{2}^{n} n^2(p_n - t) \text{ for } (p_n - t) > 0$$

Here n is the harmonic number, p_n the percentage of the nth harmonic and t the threshold harmonic percentage in the experimental conditions. One of the problems of making use of Wigan's criterion is that it is very sensitive to the value of t, which was thought, in his experiments, to be between 0.1% and 0.5%. The two measures converge for values of $p_n \to t$; however, I feel that for the purposes of this discussion it will be sufficient to use Wigan's weighting with the arbitrary value for $t = 0.1\%$.

It is easy to be led astray at this point. I have said that the ear is sensitive to defects in waveform reproduction, and it is known that amplitude non-linearities can also degrade the sound. Whilst it is convenient to measure the steady generation of harmonics, it need not necessarily be this particular effect which annoys. For example, other measurements which could be applied to quantify the non-linear amplification of a waveform are

1. The "time rate of departure of the signal from normality" as proposed by Wigan.
2. The percentage of time of deviation.
3. The r.m.s. value of deviation.
4. The peak value of deviation.
5. The measurements used in p.c.m. networks e.g. p.a.r.†

However, as far as possible, existing methods of measurement should be used and a starting point established by proposing values for the two thresholds of perception and total unpleasantness of 0.1% and 2% weighted t.h.d. respectively.

So far, no allowance has been made for transient phenomena, and it is in this parameter, perhaps more than any other, that differences between amplifiers can be detected. In deciding how to demonstrate at a *Wireless World* lecture the inadequacy of the basic specification

1. bandwidth 20Hz – 22kHz ± 2dB
2. weighted t.h.d. 0.1%
3. very low noise and hum

the following system was evolved. Linearity and hence superposition suggest that there is no reason why the audio signal should be handled by one amplifier; therefore it was proposed that the signals should be carried by a triple path amplifier, the parallel sub-amplifiers approximately covering the ranges 20Hz–990Hz, 990Hz–1010Hz, 1010Hz–22Hz.

When comparing this amplifier and another, more conventional, one (both were

†Peak to average ratio.

fed into the same very high quality power amplifier) the difference was very marked. The three-band amplifier was horrendous, with voice reproduction sounding as though it had travelled along a metal tube.

However, both amplifiers met the basic specification; my explanation for the result was that the three-band amplifier exhibited a serious transient fault at around 1kHz, the impulse response of the middle amplifier showing ringing and overhang.

This example was chosen to illustrate the inadequacy of the outline specification and is not as exotic as it may first seem. In any audio chain resonance is inevitable and it is usual that more than one be evident at the extremes of the frequency range, although there are exceptions. It is also usual to find that an amplifier will, under some conditions, exhibit a natural frequency ring when excited by an impulse.

To many, the terms transient response and transient performance are synonymous with square-wave performance and it is necessary at the outset to carefully distinguish the point of discussion. In a linear minimum phase network of first order response the rise time t_r in response to a unit step input is completely related to the bandwidth B by $t_r = \frac{0.35}{B}$.

In general the impulse response $g(t)$ can be related to the frequency domain transfer function $F(s)$ by Laplace transformation; therefore, provided the system performs linearly, the rise time of an amplifier can be deduced.

It has been thought that an audio amplifier should be designed to have as fast a rise time as possible ($<1\mu s$). This implies a frequency response extending to several megahertz. When one is faced with the situation of being told that a response to 1MHz improves the audible quality beyond that given by an amplifier having a response to 25kHz, when it is known that the system reproduces signals like Fig. 5, restricted to 20kHz by a 4th or 5th order roll-off, then it is clear that there are other mechanisms at work.

The value of square-wave testing of equipment is that it can show up

1. Frequency, phase and amplitude performance at a glance.
2. Transient misbehaviour, e.g. ringing or overshoots.
3. Slew-rate limiting.

Ringing and overshoots may excite similar problems in transducers or later amplifier stages and are best minimized. In a system which handles square waves in a linear fashion the best response shape to obtain minimal overshoot is also that which has a maximally flat phase response, i.e. the Bessell.

It is in vogue to measure the performance of a power amplifier when it is delivering square waves into a reactive load which simulates a loudspeaker, and the two most common effects noted are slew-limiting and ringing. The ringing gives an indication of amplifier stability and, although there is no agreement whether or not this has an effect on the reproduced sound, it is probably best to avoid it as much as possible.

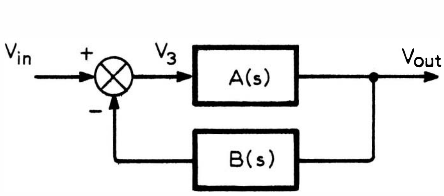

Fig. 7. A feedback amplifier configuration.

Negative feedback

It is a common assumption that all one has to do to produce an audio amplifier is to design any rough old circuit and pull the whole thing straight with negative feedback. In fact this technique could quite possibly permit an achievement of the simple specification which has been evolved so far; although obtaining good distortion figures may not be so easy. Thus to reiterate this specification:

1. Frequency response 20Hz – 22kHz $\pm$ 1dB + 10° phase
2. Power 40W
3. Weighted distortion less than 0.1% anywhere
4. Low noise and hum
5. Fast rise-time
6. Low output impedance.

I gain a definite impression that the words hi-fi and negative-feedback are generally accepted as being synonymous, and that enough negative feedback can reduce all undesirable effects. It is well known that using operational amplifier design techniques a t.h.d. of less than 0.002% is quite possible.

Consider the amplifier of Fig. 7. Classical feedback theory states that the gain will be reduced by the feedback factor F, where $F(s) = 1 + A(s)B(s)$. In addition any distortions and noise within the loop will be reduced by the same amount and the bandwidth increased as shown in Fig. 8.

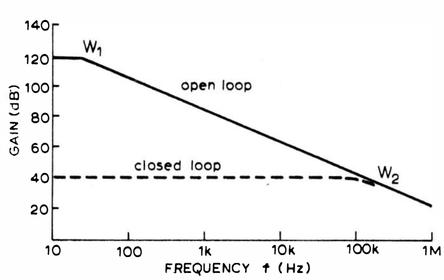

Fig. 8. Bandwidth increase with addition of feedback.

However, as is often forgotten, classical theory makes the following provisos:

1. The transfer function $A(s)$, $B(s)$ must be monotonically continuous and linear, which it is not in the event of clipping or crossover.
2. The feedback must be accurately negative at all times.
3. There must be no forward transfer of signal along the feedback path B.

The immediate implications are that distortion within the loop can only be reduced by the factor $F(s)$ if that distortion is already very small and hence $A(s)$, $B(s)$ does not deviate much from its nominal value.

In addition the theory of stability of negative feedback loops makes it clear that in a practical situation it is not possible for the feedback to be negative at all times, and hence the forward characteristics $A(s)$, $B(s)$ may have a response dictated by stability considerations. We have therefore an indication that negative feedback is not quite the acme first suggested.

Consider, for example, an amplifier of 40dB open loop gain at ω_0 with a 20dB feedback factor. It would be expected that the distortion at ω_0 would be reduced by 20dB or ten times. However, let us consider this statement in more detail. If a distortion occurs on any part of the waveform then v_3, the so-called error signal, will contain frequency components much higher than ω_0, so the effectiveness of the loop in reducing distortion at ω_0 will depend very much on its ability to detect and correct errors at a faster rate than this. Thus an important parameter when designing for low weighted distortion figures would seem to be the open loop frequency response. Two conclusions arise:

1. Negative feedback only reduces distortion by the predicted amount if the feedback is accurately negative and the distortion is very small in the first place.
2. Negative feedback will only reduce distortion at ω_0 by the predicted amount if the open-loop response has not begun to decay by ω_0.

Why do we use negative feedback? The usual reasons are given as a means of accurate calibration and stabilization of gain, to provide an extension of amplitude/frequency response together with linearisation of phase response, a reduction of the effects of open-loop distortion and a way of defining input and output impedances. Admittedly it is a very powerful design tool, but the object of introducing the subject of negative feedback is to discuss its particular shortcomings as judged by the listener.

Scroggie[16] gave a marvellous example of how negative feedback can make matters worse. An amplifier was considered which had a transfer characteristic:

$$V_{out} = 100V_{in} + 100V_{in}^2$$

and with a peak V_{in} of 0.4V this results in 20% 2nd harmonic distortion at a fundamental output of 40V pk. Applying 40dB of feedback reduced the sensitivity, reduced the maximum output to 30V pk and the distortion became 13.2% 2nd, 7.4% 3rd, 3.3% 4th . . . a weighted distortion very much more than 20%! Perhaps the most interesting aspect of amplifier feedback design for audio concerns the performance of the feedback loop under transient signal conditions.

A typical audio amplifier will comprise a pre-amplifier which may have three, four or more stages, of which two will normally have heavy overall feedback in the form of equalisation and tone controls. This is followed by a power amplifier which has a very high open-loop gain; that is, the maximum amount of overall negative feedback to minimize t.h.d.

One consequence of choosing a high overall loop gain is that stability requirements dictate that this gain be rolled off somewhat early in the audio band and it is common for commercial power amplifiers to have the first pole between 100Hz and 4kHz. This is usually effected by lag compensation in the forward path.

Transient intermodulation distortion occurs in amplifiers which employ overall negative feedback over several stages when a large enough signal is presented to the input of the amplifier at a frequency which is above the open-loop break point but is in the audio band. This type of intermodulation distortion occurs because the feedback is not operative during the open-loop rise time of the amplifier. The result is very large overshoots appearing in the error signal and depending on the particular open-loop response and feedback factor. These overshoots can be several hundred times the value of the steady-state error signal. Unless extreme precautions are taken these overshoots will cause clipping or severe overloading of the input at intermediate stages of the amplifier, and the amplifier will produce bursts of 100% intermodulation distortion.

Because the amplifier can be clipped internally, the particular circuit arrangement used can often result in transient intermodulations lasting much longer than the open-loop rise time. This mechanism has been understood for some time[17] and is analysed in some detail by Otala[18]. Figs. 9 and 10 show typical error signals in a power amplifier in response to an input step function. Here the open-loop response is 2kHz and the input is restricted to 20kHz.

It has been shown that the ear is very sensitive to this form of distortion which, in its effects, is very similar to cross-over distortion. The most rapid changes of voltage tend to occur around the zero crossing and both types of distortion produce waveform deviations in this sensitive area[19].

It is interesting that transient distortion has been largely overlooked yet its effects are quite audible. In the third part of this series of articles I will describe some interesting experiments on this problem.

To reduce steady-state distortions to a minimum it has been usual to increase the amount of negative feedback. A consequence of this is that it then becomes necessary to move the open-loop pole to a lower frequency and so inevitably transient intermodulation distortion (t.i.d.) becomes more and more likely.

I feel sure that this particular distortion mechanism is as much responsible for the notion of "transistor sound" as any cross-over problems, as it is usual for transistor power amplifiers to have more feedback and lower open-loop bandwidth than the valve counterparts.

The immediate conclusions to be drawn are:

1. Negative feedback has a clearly defined and limited use in audio amplifiers.
2. Attention must be paid to every feedback loop in the system to ensure that it does not produce t.i.d.
3. The power amplifier should have the lowest open-loop bandwidth, so the total system frequency response must be dictated in a controlled way by the pre-amplifier.
4. For ultimate quality, the minimum open-loop bandwidth is 20kHz and only enough negative feedback should be used to reduce steady-state distortions below the psychoacoustic thresholds or until the transient and steady-state distortions achieve the same significance.

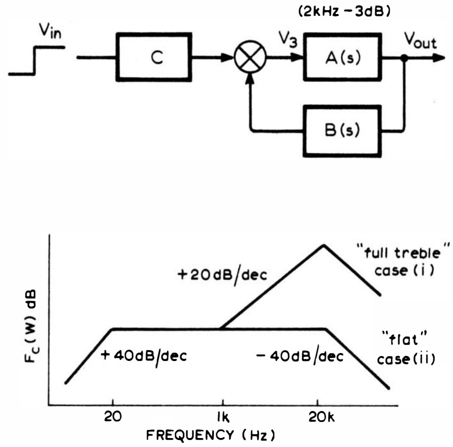

Fig. 9. Block diagram and response of a hypothetical audio amplifier.

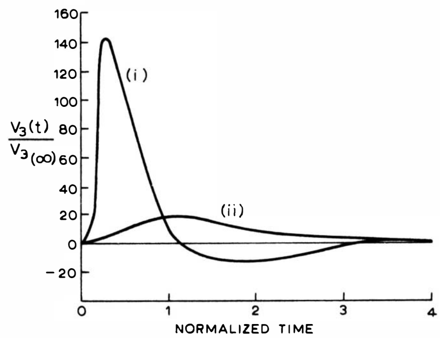

Fig. 10. Error signals produced in the amplifier of Fig. 9 with an input step function.

In Part 2 I shall continue the discussion of transient distortions and return to discussions of a figure of merit in the context of predictive design.

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Announcements

The first class for the City & Guilds **Radio Amateurs Course** (No. 765) for the 1972-1973 session begins on the 27th September 1973 at the North and West Farnborough Further Education Centre, St. John's Road, Cove, Farnborough, from where course details are available. There is also a Morse proficiency course beginning on 26th September.

The following are **courses for radio and electronics** enthusiasts offered at the Knaresborough Adult Education Centre, King James Road, Knaresborough, during the academic year 1973-74:

Tuesdays, beginning 18th September, "Morse Code For Radio Amateurs"

Wednesdays, beginning 19th September, "Electronics Workshop"

Thursdays, beginning 20th September, "Radio Amateurs Examination Course". All these classes are from 7.30-9.30 p.m. at a fee of £1 per term.

The 1973-74 edition of the annual publication **"A Compendium of Advanced Courses in Technical Colleges"** is available from the London and Home Counties Regional Advisory Council for Technological Education, Tavistock House South, Tavistock Square, London WC1H 9LR, price 70p, by post in the U.K. or from any of the Regional Advisory Councils for Further Education.

QFab Ltd, Milnathort, Kinross, Scotland, sister company to Kepston Ltd, manufacturers of electric resistance atmosphere furnaces, has begun specialization in the production of **magnetic screens** for manufacturers of electronic equipment.

Bosch Ltd, Rhodes Way, Watford, distributors of Uher equipment in the U.K., has announced that **Uher tape recording equipment** purchased in any E.E.C. country and still within the guarantee period offered in the country of original purchase will be accepted for repairs under guarantee.

Datron Electronics Ltd, has announced the appointment of REL Equipment & Components Ltd, Croft House, Bancroft, Hitchin, Herts., as their U.K. sales representatives for the Datron range of instruments, including r.m.s. digital voltmeters and r.m.s. to d.c. converters.

EMI has acquired the **cable television equipment** interests of Thorn Automation Ltd. The Thorn equipment complements the c.a.t.v. product range offered by the Telecommunications Division of EMI Sound & Vision Equipment Ltd., Hayes, Middlesex.

An Approach to Audio Amplifier Design

2. Measurement of characteristics, transient considerations, transient intermodulation index.

*by J. R. Stuart,*B.Sc.(Eng), M.Sc., D.I.C., M.I.E.E.*

In the first part of this series the discussion of an approach to the design of audio systems led to consideration of some of the characteristics that are currently measured for evaluation purposes, and to some that are not, but could be.

It is easiest of course to measure systems in terms of deviations from the specified ideal, and here I have concentrated on measurements made with standard equipment. Discussion of the design of audio amplifiers has so far isolated the following parameters for examination, and an attempt has been made to apply a level of significance to each.

Steady state:
1. Frequency/amplitude and phase responses.
2. Steady-state harmonic and intermodulation distortions.
3. Specific waveform distortions e.g. crossover, clipping.

Transient:
1. Rise-time.
2. Stability.
3. Transient intermodulation distortions (t.i.d.).

*Lecson Audio Ltd. www.keith-snook.info

Necessarily any discussion of these parameters is intimately involved with negative feedback design, and I return to this subject in this article.

Sensitivity of the ear to changes of slope around zero. The two distortion effects which, apart from serious overloading, seem to be the most objectionable to the ear are crossover and transient intermodulation distortion. Both effects can arise in any class AB or class B amplifier. The causes of both types of distortion are totally different and the designer can handle them independently. In general the values of these distortions are also independent of each other, but in the case of t.i.d. the presence of crossover distortion due to the gain deviation in the crossover region could aggravate the situation. It may be significant that both the distortions appear to have a similar subjective effect—possibly as both create a slope error (usually around the zero point)—although it may be an indirect consequence of these.

Measurement of amplifier open loop characteristics

In conventionally designed negative feedback audio amplifiers, and particularly in power amplifiers, the open loop frequency response is normally determined by lag compensation applied to the loop, although in some cases transistor characteristics may dominate, e.g. when output transistors are operated in the common-emitter mode.

The problem of measuring the open-loop transfer characteristic is that usually the a.c. feedback loop is associated with d.c. feedback, and thus d.c. feedback is essential to the correct operation of the amplifier; also the measurement technique should not substantially alter the compensations and working conditions from those of closed-loop operation.

Figs. 11(a) and (b) show two common circuit configurations with all the usual lead and lag combinations. The technique proposed for measuring the forward characteristic is to divide R_f into three elements as shown in Fig. 11(c), and to use two by-pass tantalum capacitors C_a whose value is such that the attenuation between points A and B at the measuring frequency is at least 20dB more than the measured gain. As power amplifiers normally have a closed-loop gain of at least 20dB, the error in this measurement is likely to be less than 2dB. Note that this technique maintains fairly well the impedances at points A and B and

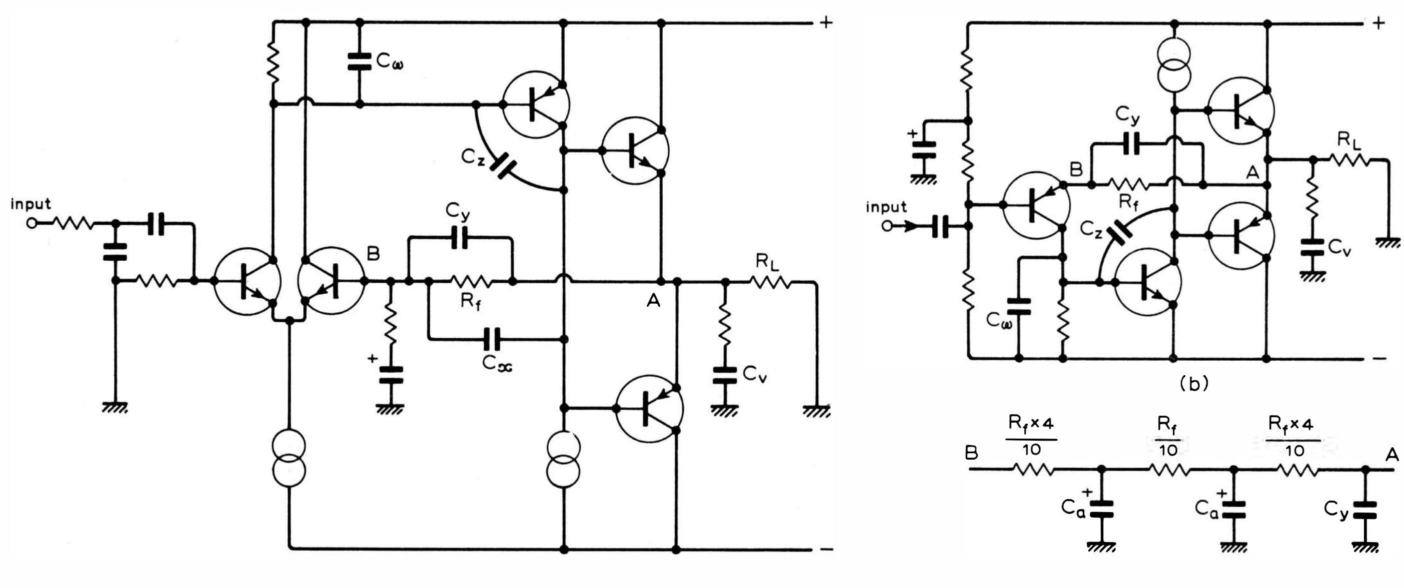


Fig. 11 (a), (b). Two typical amplifier configurations used to establish methods of measuring open and closed loop circuit performance. (c) Substitution of feedback resistor R_f by test jig circuit.

does not interfere with the conservative compensation of C_x or the loading of C_y.

This connection, with careful attention paid to stray capacitances, will allow the measurement of open-loop gain and phase (compensated and uncompensated) and open-loop distortion.

Some transient considerations

At the end of the first part in this series, a list of conclusions leading to the avoidance of t.i.d. were given in terms of negative feedback-loop design considerations. To obtain the best reproduction from an amplifier, careful attention has to be paid to other transient distortion mechanisms which include waveform distortions like ringing and overshoot and low-frequency rumble effects.

One very interesting property of negative feedback loops is that whereas usually their action is to reduce distortion appearing at the output, waveforms inside the loop may be subject to very large amounts of distortion—much more than in the open-loop condition. I have quite often measured distortions within a loop 1000× the output distortion when only 30dB of feedback is being used.

Low frequency transient effects can then arise if this distorted signal is able to change the d.c. working points of any stage in the amplifier, e.g. modify the average current in a decoupled stage as in Fig. 12. This is further aggravated, or may be induced, if the design allows the application of more d.c. than a.c. feedback. This is commonly done in an attempt to achieve an apparently high d.c. stability. In practice, small differences in the gain of the two halves of an amplifier can result in large very low frequency error signals being generated which can cause clipping or aggravate other distortions. The effect, which can be considered as a dynamic offset, is shown up by tone-burst testing.

Transient response, distortion and stability. The synthesis of a successful system, and of single-loop forward characteristics, relies heavily on judgements made in balancing the above three characteristics.

Curve 1 in fig. 13 is an uncompensated amplifier response shown as a Bode plot. Were this amplifier connected for a closed-loop gain of GdB it would be unstable. This is made more clear in the $j\omega$-plane Nyquist plot of Fig. 14.

There are three important ways in which this amplifier may be stabilised. The first is to very conservatively compensate it to achieve unconditional stability by rolling off the gain at a very controlled −20dB/decade. This makes it necessary to introduce a dominant pole, sufficiently low in frequency, say at ω_c, to ensure that either the loop gain $A\beta$ has fallen to 0dB by the second open-loop breakpoint ω_β, or that it is possible to add some lead compensation to the original response—or the feedback network—to allow a little less drastic compensation. The results of these suggestions are illustrated as curves 2 and 3 in Figs. 13 and 14. One implication is that unless the designer can find more gain in the forward characteristic $A(s)$, or is prepared to add a

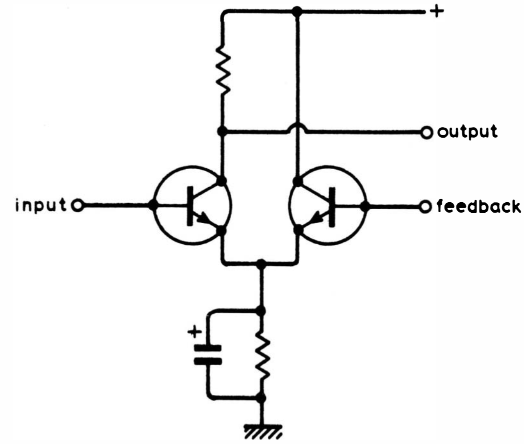


Fig. 12. Typical d.c. coupled amplifier.

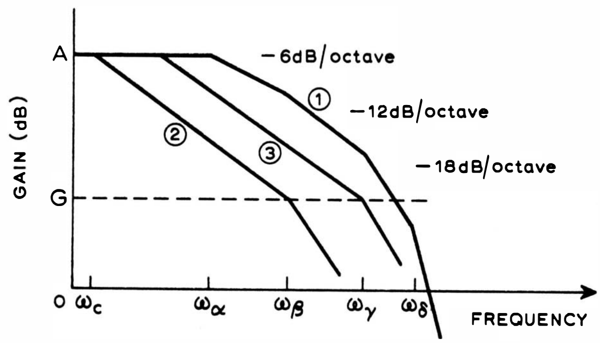


Fig. 13. Bode plot of a typical uncompensated amplifier.

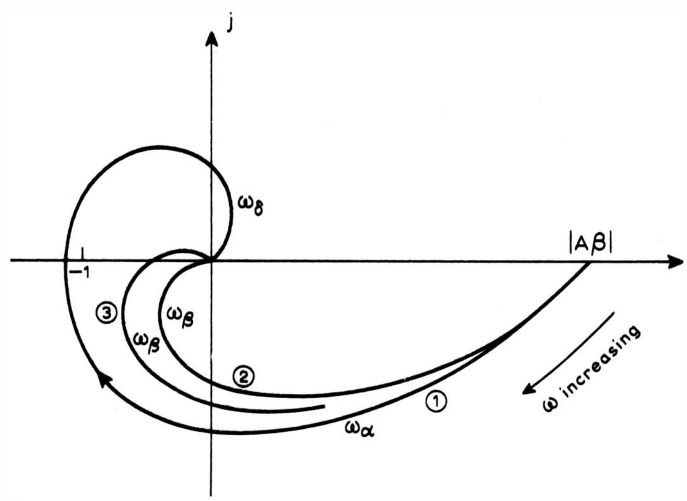


Fig. 14. Nyquist plot of amplifier example shown in Fig. 13.

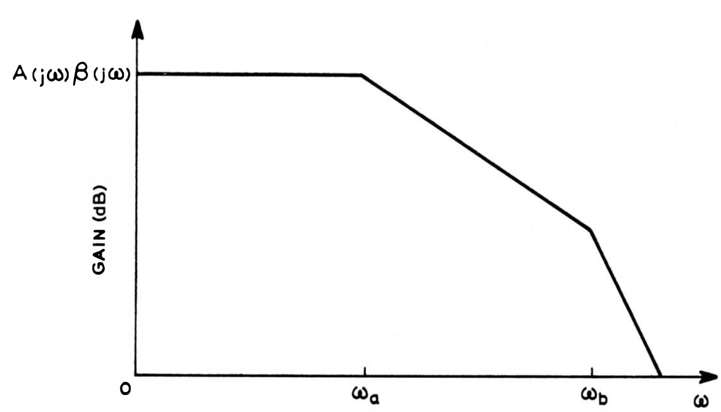


Fig. 15. Loop gain characteristic of a marginally stable amplifier.

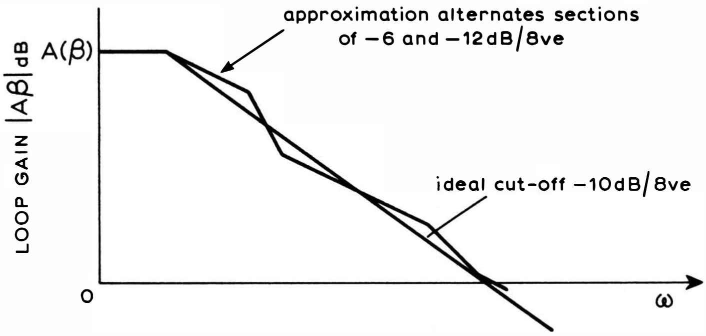


Fig. 16. An approximation to the ideal Bode loop gain characteristic.

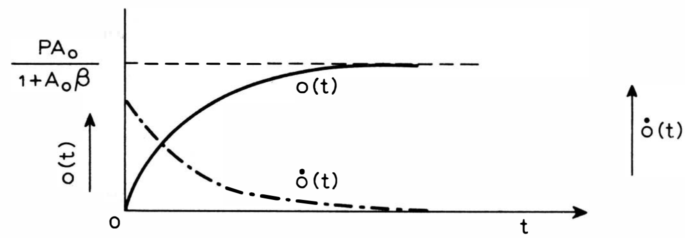


Fig. 17. Response of an amplifier with ideal Bode plot of Fig. 15, to a step function.

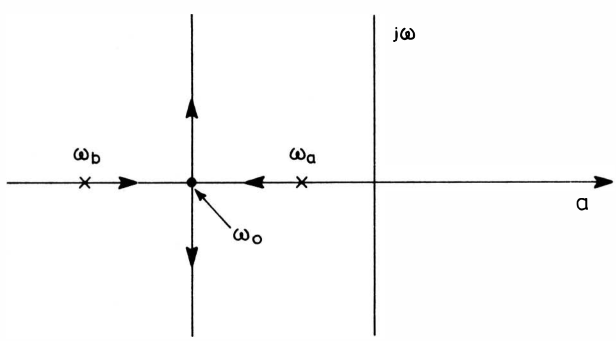


Fig. 18. An s-plane plot showing the effect of negative feedback on the natural frequencies of the two pole system shown in Fig. 17.

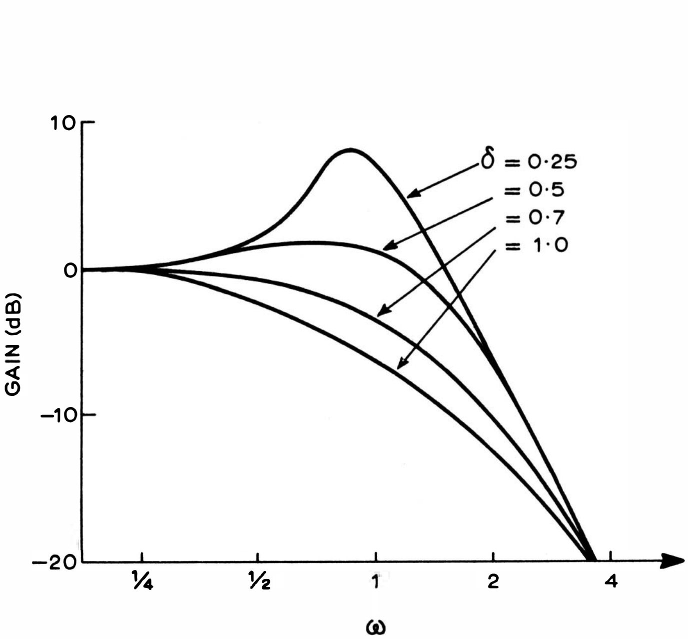


Fig. 19. The peaking of frequency response as a result of the application of more feedback which, in turn, reduces damping.

zero to the feedback network at ω_β, then unconditional stability can only be achieved by accepting that the feedback will become inoperative at ω_β.

An alternative method is to accept a marginal stability and hence permit a loop gain characteristic $A(j\omega)$, $\beta(j\omega)$ like that shown in Fig. 15. A characteristic of marginal stability is that such an amplifier may have a transient response which exhibits an overshoot or an oscillatory mode, and a peaking in the frequency response.

The third technique, of course, is to approximate to Bode's ideal loop gain characteristic as shown in Fig. 16. The first two cases will now be considered:

Unconditional stability. An amplifier whose ideal forward characteristic (see Fig. 16)

$$A(s) = A_o \cdot \frac{a_o}{s+a}$$

is connected as a feedback amplifier where $\beta(s)$ = constant β. The closed loop gain $G(s)$ is given by

$$G(s) = \frac{aA_o}{s+a(1+A_o\beta)} \qquad (1)$$

showing that the low frequency gain is reduced by the factor $(1+A_o\beta)$, along with noise and steady distortions, while the bandwidth has been increased to a $(1+A_o\beta)$. Here aA_0 is the gain bandwidth product.

The response of this circuit to a step function, value p, is given by

$$o(s) = \frac{p}{s} \cdot \frac{aA_o}{s+a(1+A_o\beta)}.$$

This gives by inverse Laplace transformation the output $o(t)$.

$$o(t) = \frac{pA_o}{1+A_o\beta}[1-\exp-(a(1+A_o\beta)t)]$$

These are sketched in Fig. 17 along with the impulse response $\dot{o}(t)$ where $\dot{o}(t) = g(t)$. It can be seen that no overshoot or ringing is possible in this circuit, and this kind of step response is one which many audio designers try to achieve.

Marginal stability. This is the condition where $A(j\omega),\beta(j\omega)$ is arranged so that when $|A(j\omega),\beta(j\omega)| = 1$ there exists a phase margin, θ, such that $\pi/2 > \theta > 0$. (Note: for conditional stability the relationship could be $\pi/2 > \theta > -\pi/2$.)

The phase margin θ is determined from $\theta = \pi - \angle A(j\omega_1),\beta(j\omega_1)$. The most general open-loop response that can be used successfully for closed-loop gains $\geqslant 1$ is the two-pole system shown in Fig. 15 and the open-loop response

$$A(s) = \frac{A_o \cdot a \cdot b}{(s+a)(s+b)}.$$

When the loop is closed this gives the form:

$$G(s) = \frac{A_o \cdot a \cdot b}{s^2+s(a+b)+ab(1+A_o\beta)} \qquad (2)$$

The effect of negative feedback on the natural frequencies of this network is shown in the s-plane plot of Fig. 18. Notice that with sufficient feedback the two poles coincide at $s_o = \sqrt{(1+A_o\beta)ab}$.

The damping factor

$$\delta = \frac{1}{2Q} = \frac{a+b}{2\sqrt{(1+A_o\beta)ab}}.$$

Note that the damping is inversely proportional to $\sqrt{1+A_o\beta}$ so, as would be expected, the application of more and more feedback reduces the damping, increases the peak in the closed-loop response and introduces an oscillatory mode to the step response for $\delta < 1$,* see Fig. 19 curves. This arises because the roots of the denominator of equation 2 become imaginary. Commonly used criteria to determine whether the transient response of such a circuit will be adequate are gain and phase margins, and the amount of peaking shown in the steady-state frequency response. To some extent, the values chosen depend upon gain and phase stability, and variability within the design, but rules of thumb suggest that $\delta \geqslant 1/\sqrt{2} \geqslant Q$, and suggest gain and phase margins of 15dB and 60°.

Reactive load conditions. One problem in the design of an audio power amplifier is that it may be expected to drive loads varying from pure inductance through resistance to pure capacitance, and may also be loaded with a tuned circuit.

So far as stability is concerned the inductive load does not concern us much† as it is assumed that the open-loop transfer characteristics referred to are measured or calculated under resistive loading.

Loading a practical amplifier with a pure capacitance, to simulate perhaps electrostatic loading, will result in the introduction of a new pole in the transfer function at ω_n. It can be seen from Fig. 20(a) that the effect of this on an unconditionally stable amplifier is that for ω_n, $\omega_{n1} \angle \omega_a(1+A_o\beta)$ a ringing may appear in the closed loop step response, whereas little change will occur in the step response if $\omega_{n1} > \omega_a(1+A_o\beta)$.

In an amplifier of 2nd order response working into a resistive load, the effect of adding the pole $\omega_n < \omega_x$ can be to make the amplifier unstable; or for $\omega_n > \omega_x$ little change should be noted. See Fig. 20(b).

Capacitive load condition. One technique often used to ensure stability and/or freedom from overshoot into capacitive loads is to deliberately over compensate the open-loop response so that at no time does ω_n fall below the closed-loop cut-off frequency. This technique is unnecessarily punitive and will obviously result in a lower open-loop break-point.

Another technique is to modify the output circuit so that although ω_n can be $\ll \omega_x$, the change to the open-loop response is small, ensuring that at $|A(j\omega)| = \beta^{-1}$ the slope of $A(j\omega)$ is the same as in the resistive termination. Such a modification involves a resistor and inductor in series with the output. This is shown in Fig. 20(c).

The last technique to take account of this type of loading involves adding a zero to the

*Critical damping

†Note: the inductive load is important when considering low frequency design.

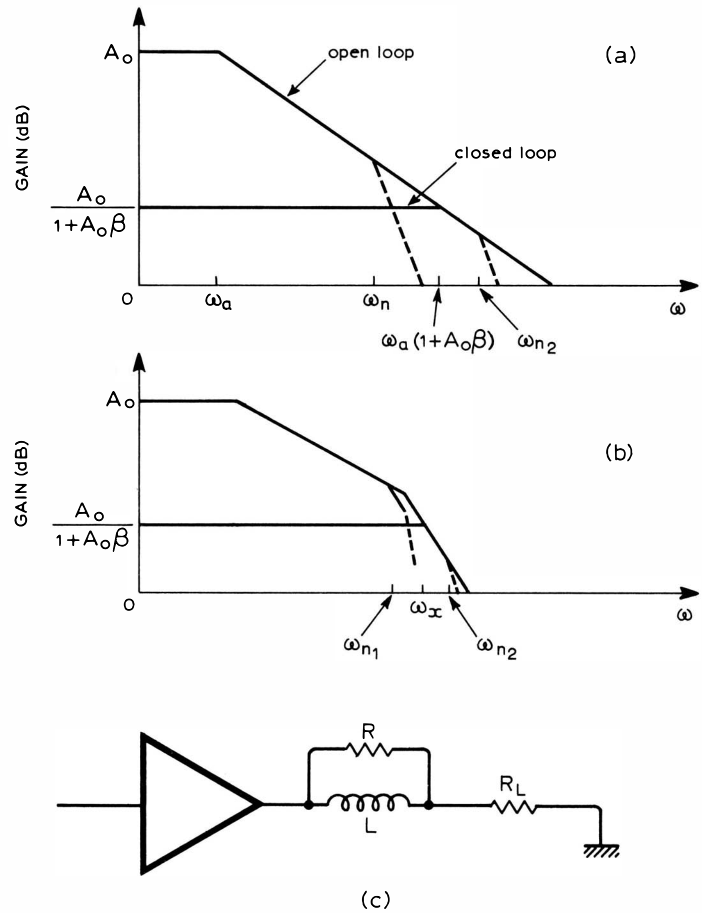


Fig. 20. Effect on single pole (a) and two pole (b) amplifiers of driving into a purely capacitive load. (c) Output circuit modification to improve stability with a capacitive load.

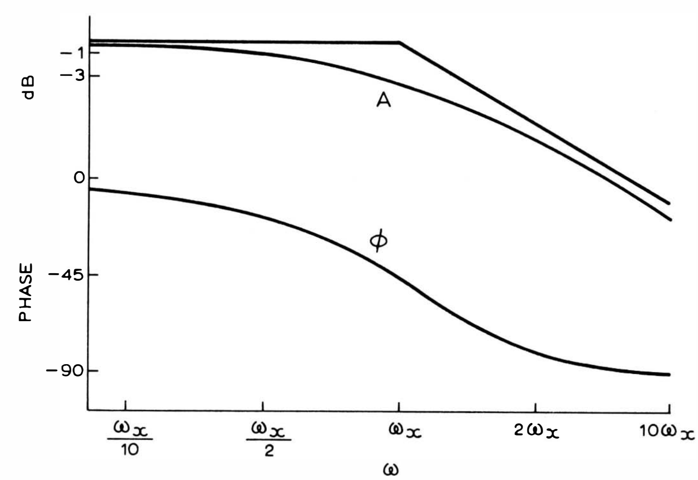


Fig. 21. Loop gain and phase characteristics of single pole amplifier.

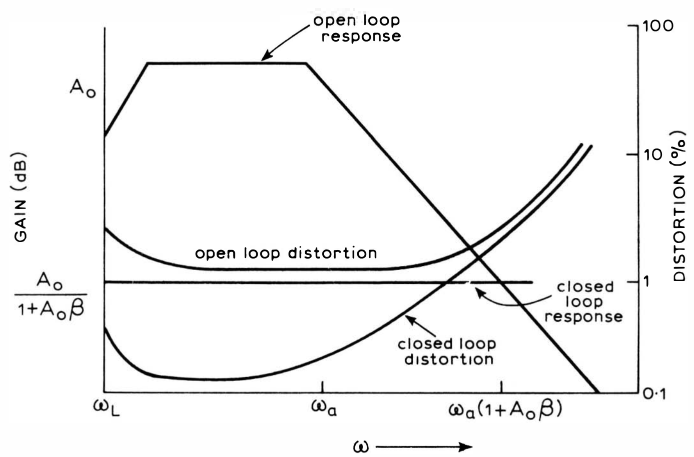


Fig. 22. The projected effect on distortion of the amplifier loop characteristic.

feedback network; hence $\beta(s)$ is of the form:

$$\beta(s) = \frac{s+m}{m} \cdot \beta$$

This of course has the effect of adding a zero to the closed-loop transmission, and if m is chosen to be equal to ω_n for the largest value of capacitor load envisaged, say $2\mu F$, then the transient response and stability will not suffer provided $\omega_n > 50kHz$ (for an audio amplifier).

Phase response

Neglecting for the moment the example of an amplifier with a second order open-loop response, let us consider the responses of the single pole example shown in Fig. 21 having its pole at ω_x. If we consider the closed-loop response of the system we can see that this also described the closed-loop responses where

$$\omega_o(\text{closed loop}) = (1+A_o\beta)\omega_o$$

The specification for a high quality amplifier quoted in the first part of this series contained the suggestion that the amplifier as a whole should have phase tolerances of $\pm 10°$. Given that a passive filter was shown to be desirable in the pre-amplifier, for transient distortion reasons, and that this will introduce a phase lag, it can be seen that the permissible phase shift in any one stage is small, and ideally $-2°$ at 20kHz in a typical system. Note that this and the requirements for stability apply equally to the low frequency design. An amplifier with a response of the form

$$G(s) = \frac{A_o \cdot a}{s+a(1+A_o\beta)}$$

will have a phase shift of $-2°$ at 20kHz for $\omega_x = \sin\omega_a(1+A_o\beta) = 570kHz$ and $-6°$ for $\omega_x = 200kHz$.

This result suggests that the phase requirement places fairly strict bounds on the gain-bandwidth product for any design; e.g., if we wish to have a phase lag of $-2°$ at 20kHz and to use 40dB of feedback then the open-loop pole $\omega_a = 5.7kHz$ min.

In the second order case the requirement for a given phase lag will be a greater open-loop $-3dB$ bandwidth due to the more rapid open-loop phase shifts.

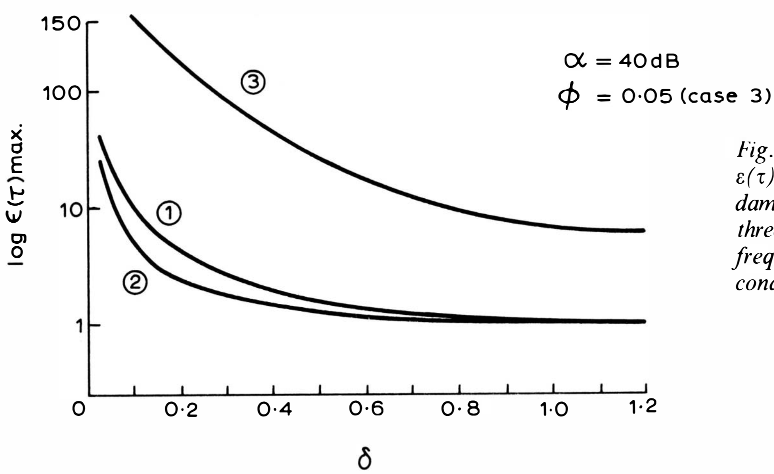


Fig. 23. Values of $\varepsilon(\tau)$ plotted against damping factor for three different frequency response conditions.

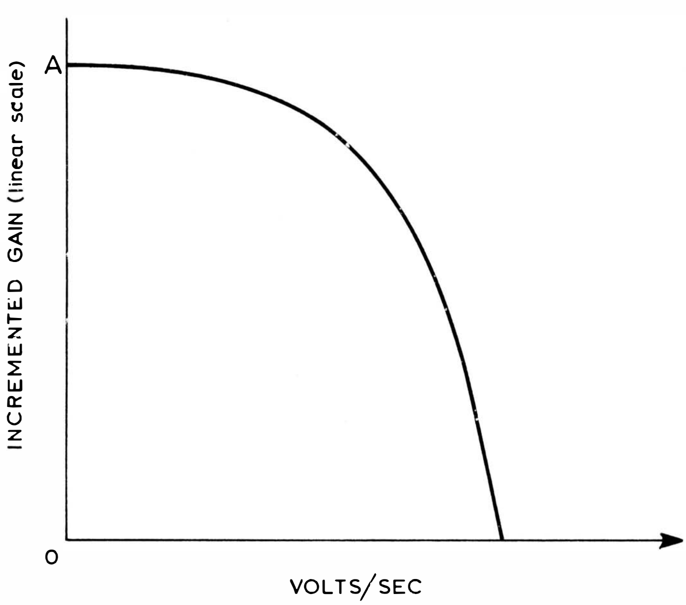


Fig. 24. Model of performance as a function of input slew rate.

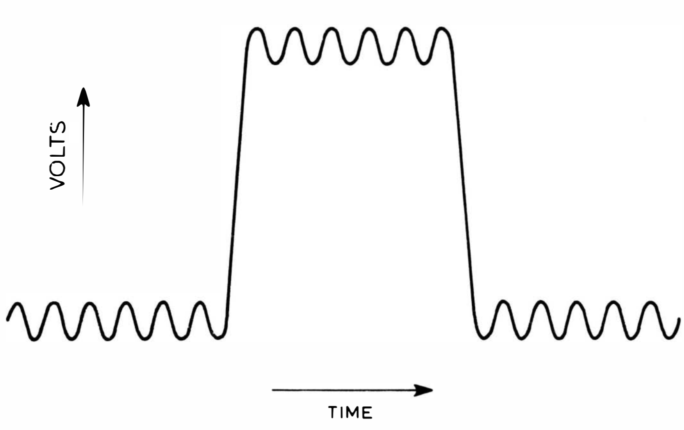


Fig. 25. Test signal used to probe incremental gain.

Distortion reduction

A point was made in the first part of this series that a weighted steady-state distortion of 0.1% at any frequency in the audio band was a good and sufficient goal at which to aim. In addition, general discussion of the mechanism of negative feedback loops produced three points:

1. The steady-state distortions, amenable to reduction by feedback, are not reduced for a signal of frequency ω_1 by the amount $(1+A(j\omega_1),\beta(j\omega_1))$ but by an amount related to $(1+A(jn\omega_1),\beta(jn\omega_1))$ for n, integer $\geqslant +2$. This accounts for any reduction of loop gain at harmonic frequencies and phase shifts at these frequencies.
2. The weighted steady distortions in an amplifier where $A(j\omega) = \text{constant}$, $\omega = 0$ to $\pm\infty$, will only reduce by the amount $(1+A\beta)$ if the distortions are already small, so that $A\beta(v)$ has small limits; otherwise not only will the t.h.d. not reduce, but the weighted t.h.d. can be expected to reduce less, due to a changed harmonic structure being introduced.
3. It is likely that negative feedback will reduce the maximum unclipped output. Fig. 22 illustrates the expected effect of the loop characteristic on the distortion of an amplifier. Obviously the only two points the designer must carefully consider are the extremes of the range—as would be expected. The arguments of course apply equally to intermodulation distortion.

It can be seen that for any particular specification, e.g. $\not> 0.1\%$ weighted t.h.d. in the band 20Hz–22kHz, the open-loop gain and phase response will be an important consideration, a wider bandwidth giving a neater solution to the t.h.d. specification and making it necessary to obtain a given distortion characteristic, to use less low frequency gain and hence achieve better stability, and phase response.

System transient analysis

In analysing the transient performance of a system, that is pre-amplifier plus power amplifier, the power amplifier has been considered as having a first order open-loop response—or at least a second order response where $\omega_b \gg \omega_a$ (Fig. 15). This has been done for simplicity; however, the conclusions are general. Three conditions have been envisaged:

1. Pre-amplifier response declines at -6dB/octave above ω_1.
2. Pre-amplifier response declines at -12dB/octave above ω_1.
3. Pre-amplifier response is boosted by $+6$dB/octave above ω_2 and decays at -6dB/octave above ω_1.

In each case the power amplifier has an open-loop pole at ω_o and has a response

$$\frac{e_o}{e_1} = G(s) = \frac{A \cdot \omega_o}{s+\omega_o(1+A\beta)}.$$

The error signal ε is described by the relationship

$$\frac{\varepsilon}{e_1} = K(s) = \frac{s+\omega_o}{s+\omega_o(1+A\beta)}.$$

Put $(1+A\beta) = \alpha$ in case 1, then, the total input–output response is given by:

$$H(s) = A \cdot \frac{\omega_1}{(s+\omega_1)} \cdot \frac{\omega_o}{(s+\alpha\omega_o)}$$

and in response to step function value p, the output is given by:

$$e_o(s) = \frac{p}{s} \cdot \frac{A\omega_o\omega_1}{(s+\omega_1)(s+\alpha\omega_o)}$$

and the error signal:

$$\varepsilon(s) = \frac{p}{s} \cdot \frac{(s+\omega_o)}{(s+\omega_o\alpha)} \cdot \frac{\omega_1}{s+\omega_1}$$

Taking partial fractions we deduce that

$$\frac{\varepsilon(s)}{p} = \frac{1}{\alpha s} + \frac{\dfrac{\omega_1(1-\alpha)}{\alpha(\omega_o\alpha-\omega_1)}}{[s+\omega_o\alpha]} + \frac{\dfrac{\omega_1-\omega_o}{(\omega_o\alpha-\omega_1)}}{[s+\omega_1]}$$

Setting $\omega_o/\omega_1 = \delta$ and $\omega_1 t = \tau$ and using the reverse Laplace transform we get

$$\varepsilon(\tau) = \frac{p}{\alpha}\left[1+\frac{(1-\alpha)}{(\alpha\delta-1)} \cdot \exp(-\delta\alpha)\tau - \frac{\alpha(1-\delta)}{(1-\alpha\delta)}\exp-\tau\right] \quad (5)$$

By differentiating $\varepsilon(\tau)$ and setting this signal to zero we get the stationary points. Other than at $\tau = 0,\infty$ there is a value of τ for which $\varepsilon(\tau)$ is a maximum given by:

$$\tau = \frac{1}{(1-\alpha\delta)} \ln \frac{(-\delta)}{\delta(\alpha-1)} \quad (6)$$

The value of $\varepsilon(\tau)$ for various values of α and δ are plotted in Fig. 23. A similar analysis shows for case 2,

$$\varepsilon_2(\tau) = \frac{p}{\alpha}\left[1+\frac{(1-\alpha)}{(1-\alpha\delta)}\exp(-\alpha\delta\tau) + \left\{(\delta-1)\tau - \frac{1-\delta(2-\alpha\delta)}{(1-\alpha\delta)}\right\} \frac{\alpha}{1-\alpha\delta} \cdot \exp-\tau\right] \quad (7)$$

and for case 3

$$\varepsilon_3(\tau) = \frac{p}{\alpha}\left[1+\left\{\frac{(1-\alpha)}{(1-\alpha\delta)} \cdot \frac{(\alpha\delta-\phi)}{\phi(1-\alpha\delta)}\right\} \exp(-\alpha\delta\tau) + \left\{(\delta-1)(\phi-1)\tau - \phi(\delta\phi-1) + \frac{(\delta-1)(\phi-1)}{(1-\alpha\delta)}\right\} \cdot \frac{\alpha}{\phi(1-\alpha\delta)} \exp-\tau\right]$$

where $\phi = \omega_2/\omega_1$.

The values of $\varepsilon(\tau)$ are also shown for cases 2 and 3 in Fig. 23. It is clear that the form of $\varepsilon(\tau)$ allows quite large overshoots to occur in the feedback loop and these can have a peak value much larger than the steady state error signal for the highest frequency ω_1. It is interesting to consider the reason for this, and once again the surprise can be traced back to the classical feedback model. In the open-loop amplifier, the time for the output to rise from 10% to 90% of the final value was given as $0.35/2\pi\omega_o$ sec. Now when the loop is closed, to achieve the original output level, instead of this condition, the effective input signal is increased in an attempt to increase the output slew-rate.

The classical theory breaks down here, because in a practical amplifier the following conditions will be obtained:

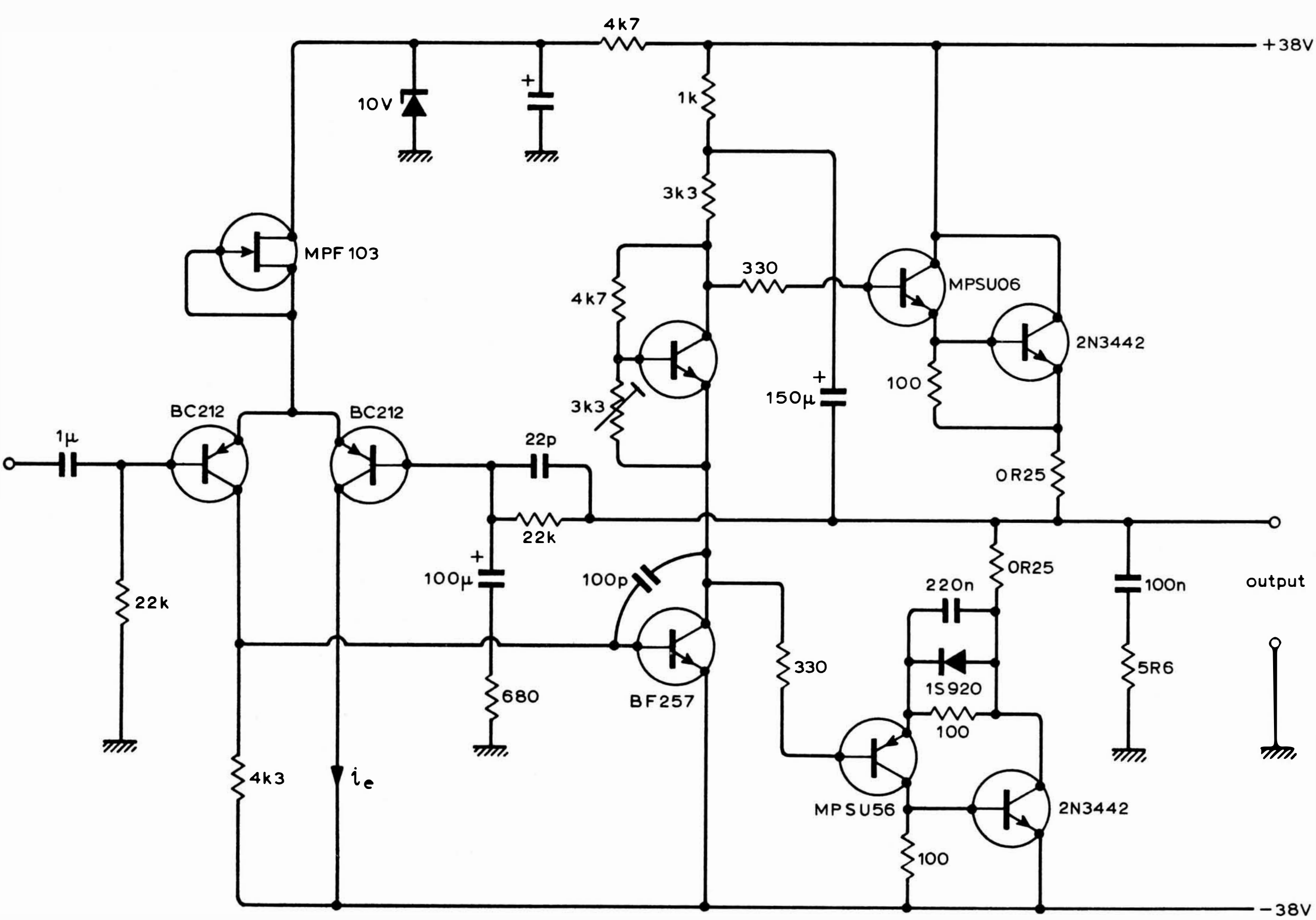


Fig. 26. Basic amplifier design produced to test transient distortion analysis.

- The amplifier will exhibit a time delay.
- The output slew-rate will not increase indefinitely in response to increased input signal level.
- In practice there will be a maximum slewing rate for the amplifier (S_{out} volts/sec) which is determined by the internal time constants and overload margins. At this point the relationship $S_{out} = AS$ totally breaks down.

This amounts to the feedback network being totally inoperative for a portion of the open-loop rise-time of the amplifier, which can in turn have three effects:

- The incremental gain of the amplifier is not controlled during this time.
- The resulting overshoot in $\varepsilon(\tau)$ can overload the early stages of the amplifier, resulting in a 100% intermodulation during the rise time.
- The overshoot may, depending on the amplifier recovery characteristics, cause the amplifier to latch-up, resulting in an intermodulation burst much longer than the rise-time. The burst is commonly several milliseconds in many commercial amplifiers, particularly if an amount of treble boost is applied as in case 3.

The transient distortion need not produce 100% i.m.d. since amplifiers tend to overload progressively. Fig. 24 is a model of performance as a function of the input slew rate S_{in}. The incremental gain will, of course, indicate the amplifier's ability to handle any other signal present before, during and after the transient. A useful test signal to probe this is shown in Fig. 25.

It will be remembered that the sensitivity of the ear to changes of slope was discussed earlier. It seems that there will be an equal likelihood of response to gain deviation in the vicinity of a transient as to disappearance of the incremental gain.

Transient intermodulation index

Transient distortion can be analysed statistically, the likelihood of such a distortion arising being dependent upon the probability that an overshoot will arise within the loop sufficient to cause a significant gain deviation. This can in turn be related to the product of two parameters, the transient intermodulation index (t.i.i.), which describes the maximum size of overshoot expected from the previous analysis, and the overload margins in the amplifier.

The transient intermodulation index has been proposed here to give a rough idea of the amplifier's susceptibility to this form of trouble, and has been normalised so that an amplifier considered to have a good quality, judged by listening experiments and having, say, 20kHz open-loop bandwidth with 40dB of feedback, is normalized to 0.1.

From the analysis it can be seen that the size of an overshoot could be approximately proportioned to F/δ where F is the feedback factor. The important thing to note here is that δ refers the signal bandwidth to the open-loop bandwidth and so this relationship can be applied to any stage in the amplifier.

An example: To test the analysis an amplifier was built as shown in Fig. 26. It is a very common circuit and there is nothing special about the way the forward compensation has been applied. The dominant open-loop pole is set by C_y and, as can be seen from the steady-state summaries in Fig. 27, the performance is nothing special. However, it is typical of the configuration which, due to device limitations, can never achieve the ultimate quality of alternative circuits.

Fig. 27. Graphical summary of steady state performance of amplifier design shown in Fig. 26.

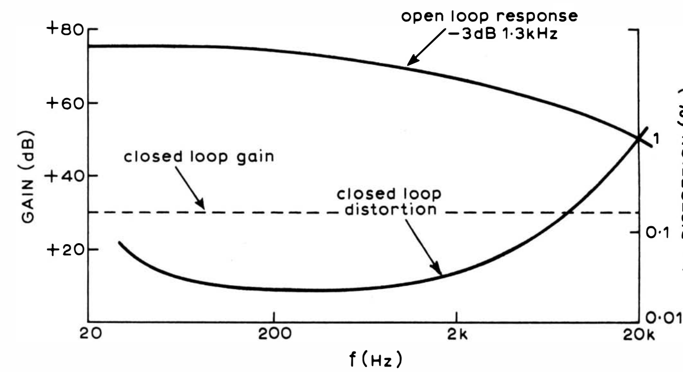
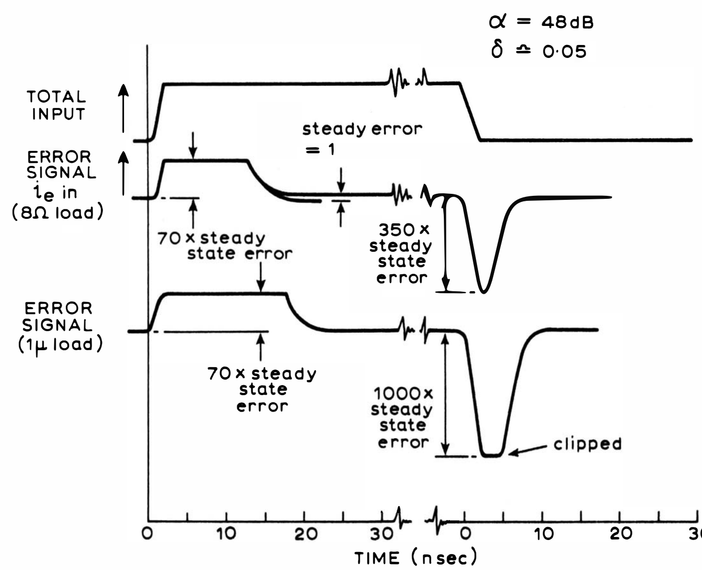


Fig. 28. Transient response to the test amplifier to square wave test signals for a resistive 8 load and a 1 F capacitive load.



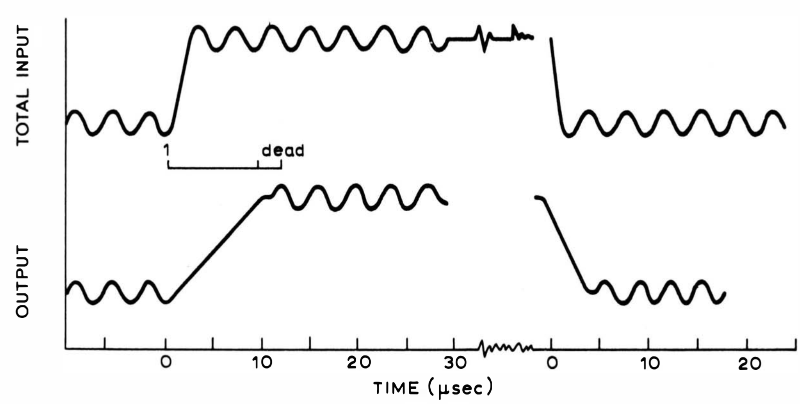


Fig. 29. Effect of combining a sine or square wave to demonstrate t.i.d.

Two experiments were carried out and in each case the input signal was passed through a network with a single pole at 44kHz. In the first a square wave was applied at 1kHz p.r.f. and 2μs rise and fall time and the error signal was measured for 20V peak to peak output. This waveform is shown in Fig. 28. Note that the overshoot is 350 x the steady-state error in the negative direction and that in the positive direction the error is clipped at 70 x the steady-state error.

Secondly, a 20kHz sine wave of 1V peak to peak was added to the input signal and, as Fig. 29 shows, a transient distortion occurred, which was considerably aggravated for both positive and negative excursions by the addition of a capacitive load.

Now obviously this can be much improved by increasing the overload margins of the first stage. Even when the situation

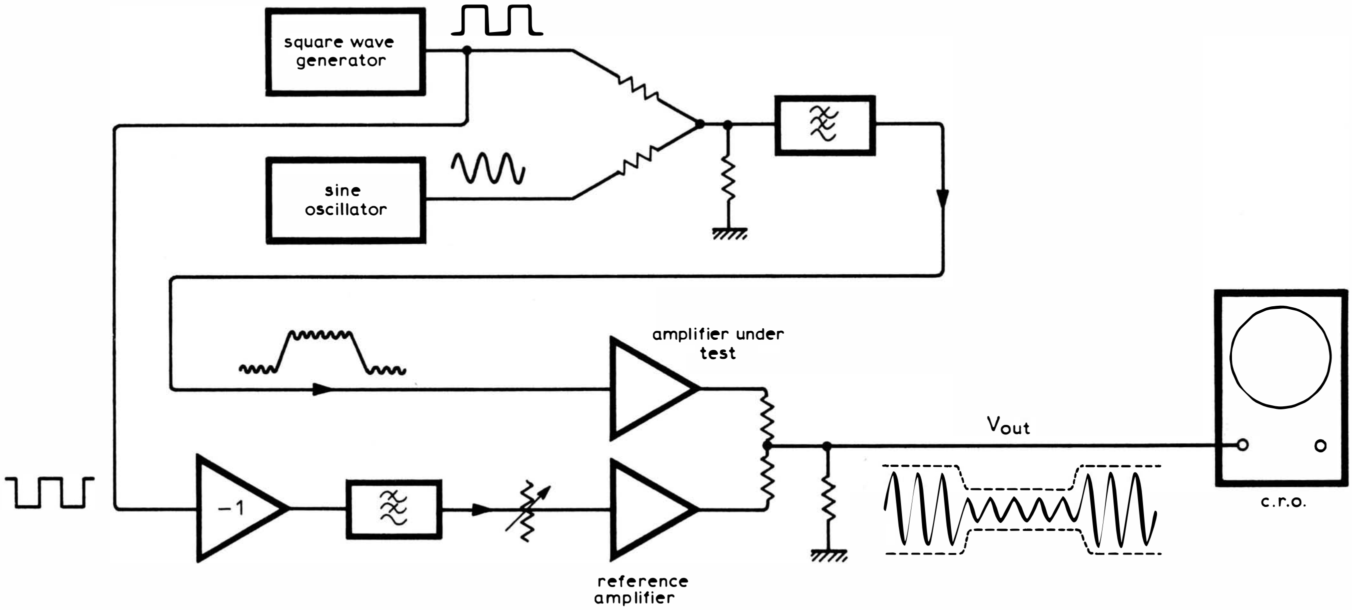


Fig. 30. Test rig used to measure t.i.d.

can be arranged so that the gain does not disappear, experiments indicate that the oscilloscope is not sensitive enough to show up the sort of gain deviations to which the ear responds. It is quite difficult to see 5% t.h.d. as a sine wave.

One technique I have used to increase the sensitivity of the sine plus square test is shown in Fig. 30. Here, the square-wave is modified by a phase characteristic equal to that of the amplifier under test, e.g. a duplicate amplifier, and the resulting signal is subtracted from the output waveform. The resulting signal allows 5% gain deviation to be seen.

The treatment of steady distortion

The transistor is basically non-linear in many respects, and the designer of any equipment must understand this completely to design predictively. It is convenient to consider the non-linearities of the transistor in three cases:

- Small signal conditions, high and low frequency.
- Large signal, low frequency.
- Large signal, high frequency.

The small signal condition is, however, best considered as a special case of the large signal, except that some useful analyses are available on the minimization of second-harmonic distortion in small signal stages. Owing to the general shape of the polynomial describing the basic relationship between collector current (i_c) and base current (i_b) the predominant distortion is normally second harmonic. At low frequencies the non-linearities are (Fig. 31):

- The exponential form of the relationship between i_b and V_{be}, and of V_{be} with temperature.
- Variations of h_{fe} and h_{FE} with collector current i_c and with collector emitter voltage V_{ce}. The variation of h_{fe} with high values of V_{ce}, which is due to narrowing of the base area (Early effect). Also h_{fe} is subject to variation with temperature.

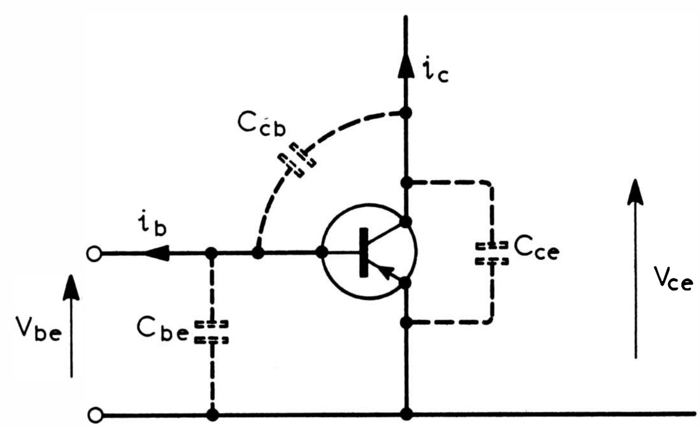


Fig. 31. Sources of non-linearity in a transistor.

At high frequencies the following effects are to be added:

- Variations of C_{be}, C_{cb} and C_{ce} with device temperature, V_{ce} and i_c.

Particularly for large signals this amounts, or can amount to, large changes in gain (10 to 1000 ×), of any stage during a signal cycle, and particularly so at high frequencies.

The fundamental freedoms available to the designer are:

- Choice of device to optimise the effects for any circuit.
- Choice of quiescent operating conditions (V_{ce} and i_c) and of large signal swings.
- Control of the source and load impedances. In particular, choice of the correct source impedance for a stage can be fundamental in reducing distortion.
- Choice of the amount of local feedback.

In part 3, further discussion of steady state distortions leads to a description of a recent design. The idea of a figure of merit is returned to and the results of experiments given together with some consequential proposals. www.keith-snook.info

An approach to audio amplifier design

3 System design, applying the figure of merit.

by J. R. Stuart, B.Sc. (Eng.), M.Sc., DIC, M.I.E.E.E.

In the second part of this series, the discussion of an approach to the design of an amplifier as part of a system led to a detailed analysis of the application of negative feedback loops. Highlighted in this analysis was the way in which the open loop characteristics of an amplifier need to be related to the closed loop operating conditions in order to achieve the correct compromise of phase, transient and steady-state distortions.

Steady-state distortions

The transistor parameters which contribute to non-linearity have been listed in part 1, as follows.

- The exponential form of the relationship between i_b and V_{be} and of V_{be} with temperature.
- Variations of h_{fe} and h_{FE} with collector current i_c, with collector-emitter voltage V_{ce} (Early effect), and with temperature.

At high frequencies other effects are in variations of C_{be}, C_{cb} and C_{ce} with chip temperature, V_{ce} and i_c. Apart from controlling quiescent conditions, the major freedom available to the designer in defining the forward or open-loop characteristics of an amplifier is the choice of source and load impedance for each stage and of the amount of local feedback to be applied.

*Lecson Audio Ltd.

The two most useful techniques for reducing distortion introduced by device non-linearities are local emitter feedback (in a common emitter amplifier) and the cascode configuration. Fig. 32 shows a simple common emitter amplifier with and without local feedback supplied by R_e, and the small signal equivalent circuit for each.

We have for the case with local feedback, the trans-impedance

$$R_b = \frac{V_0}{i_S} = \frac{h_{fe}R_LR_S}{h_{11}+R_S+R_e(h_{fe}+1)}$$

Setting $R_e \to 0$ gives the case of no feedback Fig. 32(a)

$$R_a = \frac{V_0}{i_S} = \frac{h_{fe}R_LR_S}{R_S+h_{11}}$$

By partial differentiation the sensitivity of R_a and R_b to device parameters can be shown, e.g. for change of h_{fe} for whatever reasons we have:

Case (a) no feedback

$$\frac{\delta R_a}{\delta h_{fe}} = \frac{R_LR_S}{h_{11}+R_S}$$

Case (b) with feedback

$$\frac{\delta R_b}{\delta h_{fe}} = \frac{R_LR_S}{h_{11}+(h_{fe}+1)R_e+R_S}\left[1-\frac{h_{fe}R_e}{h_{11}+(h_{fe}+1)R_e+R_S}\right]$$

This represents an improvement in gain stability of:

$$\frac{[h_{11}+(h_{fe}+1)R_e+R_S]^2}{(h_{11}+R_S)(h_{11}+R_e+R_S)}$$

Analysis will show the same improvement for many device parameters and a similar form of improvement in high frequency effects.

The cascode arrangement shown in Fig. 33 allows the common-emitter stage to be virtually freed from the Early effect and modulation of C_{ce}, as the device is allowed to operate at constant V_{ce}; this clearly also allows a higher bandwidth to be achieved by the stage for given source and load impedances as the Miller effect is considerably reduced.

Design of a system

The preceding arguments in parts 1 and 2 indicate that an amplifier designed to sound very good cannot necessarily be synthesized from the basic specification:

1. Output power in excess of 40W.
2. Power bandwidth 20Hz – 30kHz ± 1dB.
3. Very low noise and hum, say −80dB.
4. t.h.d. less than 0.1% at all frequencies and power levels in the bandwidth.
5. i.m.d., however measured, less than 0.1%.
6. Low output impedance, say 400mΩ.

However it seems reasonable in the light of the preceding discussions to propose a starting point specification for very good quality as below.

1. Output power in excess of 40W.
2. Power bandwidth 10Hz – 30kHz ± 1dB.
3. Very low noise and hum, say −80dB flat, −80dB C.C.I.R. weighted.
4. Weighted total harmonic distortion less than 0.1% at all frequencies and power levels; i.e. 10Hz – 20kHz, 0–40 watts.
5. i.m.d., however measured, less than 0.1%.
6. Low output resistance, say 400mΩ; 10Hz – 20kHz.

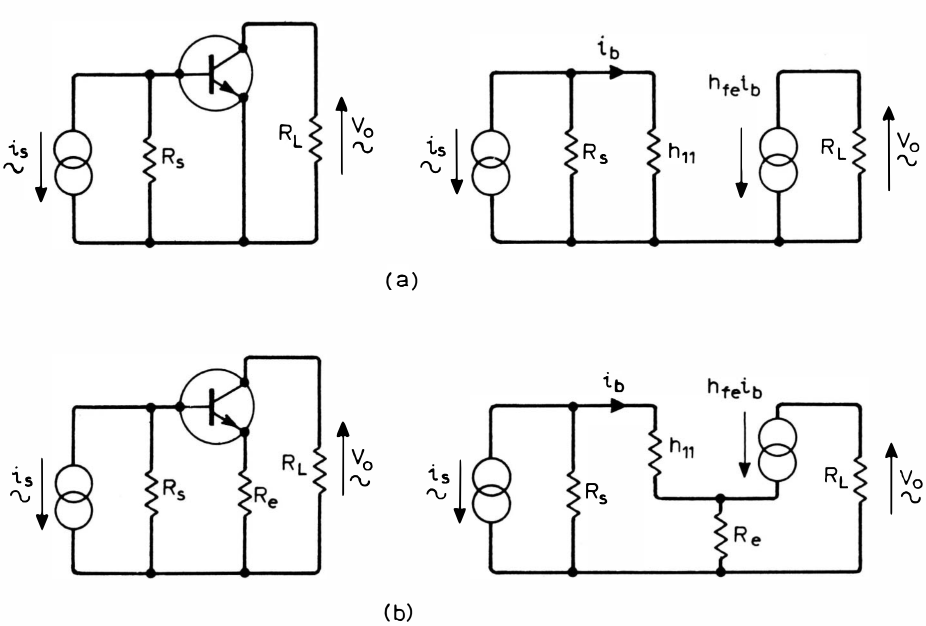


Fig. 32. Common emitter amplifier drawn (a) without feedback and (b) with feedback together with their small signal equivalents.

7. Open loop frequency response—any loop -3dB at 20kHz min.
8. Feedback factor -40dB any loop.
9. Phase accuracy $\pm 10°$ 20Hz–20kHz.
10. Accurate overload characteristic inside the loops.

A typical audio amplifier system will be as shown in Fig. 34; here three major negative feedback loops are isolated. These are around the low noise input amplifier, in which equalization may be applied, the tone control stage and the power amplifier. In addition there is the volume control and a stage of filtering which need not be achieved by feedback loops.

It has been shown earlier that for any single stage to have a phase shift of 2° at 20kHz then the minimum -3dB closed loop bandwidth for that stage is 570kHz; three such stages cascaded would have a total lag of 6°. It has also been demonstrated that it is not desirable to drive any audio feedback amplifier significantly above its open-loop bandwidth; therefore if the signal can be restrained to say 45kHz in the filter stage, then the open-loop response of the two stages following the filter should be as similar as possible, thus giving a guide to the feedback factor that can be applied for a given overload margin.

The choice of 45kHz for a passive roll-off is a compromise between the phase distortion introduced by such a filter and t.i.d. in the power amplifier. It is not in any way a magic number and may be different in every design.

At this stage the designer runs seriously short of information, in particular the extent to which phase shift can be traded off for incipient t.i.d., and this is discussed later. However, it seems reasonable to me that in view of the poor phase performance of parts of the audio chain outside direct control, e.g. the recording studio, and in view of the high apparent sensitivity of the ear to t.i.d., that it would always be preferable to err on the side of a lower passive roll-off and higher phase shift—but as a compromise—not a rule.

A recent design. A commercially available amplifying system* designed by myself is shown in block diagram form in Fig. 35.

A low-noise high overload input stage is followed by an active volume contol, filters and tone control; in each case the open-loop bandwidth and feedback factor, FdB, is shown. Care has been taken to ensure that no transient distortion effects can arise with an audio signal, and the signal bandwidth of the system is constrained to 45kHz with a third-order Bessell roll-off which introduces a lag of 12° at 20kHz.

It is clear from the arguments presented that, for an unconditionally stable characteristic in an amplifier which exhibits no transient distortion effects in the signal bandwidth, a low feedback factor is necessary. This is because any increase of feedback factor must be accompanied (in the general and practical situation†) by a reduction of open-loop bandwidth, ω_{OL}. The consequences of this are a rise of steady-state distortion starting below ω_{OL} and an increased possibility of t.i.d. Therefore, in order that the amplifier should also have a weighted t.h.d. of less than 0.1% at any frequency or power level, it was essential to achieve a low open loop distortion figure.

The final power amplifier design, which is shown in block diagram form in Fig. 36, uses a new configuration which is the subject of a British patent application.

Use of local stage feedback combined with a complementary form and output triples operating in class AB gives an open-loop bandwidth of 17.5kHz and distortion of 0.2%. The application of 32dB of feedback reduces the weighted t.h.d. well below 0.1% and gives an unweighted t.h.d. of 0.005% between 100Hz and 3kHz.

A figure of merit

Earlier I put forward the idea of a figure of merit which describes the quality of an audio chain or a link of that chain. This is a number derived from a weighted sum of undesirable characteristics, measured in terms of the critical parameters. This figure of merit (f.o.m.) may be time variant; that is, an amplifier may have for example a rating of 0.8 (1971) and 0.7 (1973).

It was further proposed that by using collective subjective results, any parameter could be assigned a measure of significance, and further that the starting points for each parameter would be the thresholds of perception and objection—the latter Mantel[3] calls "the threshold of non-neglectability".

Successive experiments may then show improved accuracy in the choice of parameters, defining thresholds and curve fitting between the thresholds.

In this article I propose to outline a workable f.o.m. which is based on current knowledge as outlined, in the hope that its defects can be improved upon by large-scale experimental work.

The working of an f.o.m. Let us consider that the figure of merit for a chain or item in the chain be M, where M is the probability that a person will not be able to detect a shortcoming in the sound. This could be restated as M = probability of non-detection of a shortcoming by a member of the population chosen at random. Therefore an ideal audio system would have an $M = 1$ and a poor system $M = 0$.

For each stage in the chain of n elements we propose m_i ($i = 1, 2 \ldots n$) such that the total figure of merit for the n cascaded stages is $M_T = \prod_{i=1}^{n} M_i \alpha_i$, where α_i is a weighting factor.

Each partial f.o.m. M_i is composed of a weighted product of factors believed to influence the quality of the sound, such that M_i shows the joint probability that any one factor may be detected as a shortcoming. Therefore in a simple example, if M_i considers only the terms

- $p(d)$ the probability of detection of d% weighted t.h.d., and
- $p(n)$ the probability of detection of ndB s/n ratio

then we may write

$$M_i = q(d).q(n)$$

where

$$q(d) = 1 - p(d)$$
$$q(n) = 1 - p(n)$$

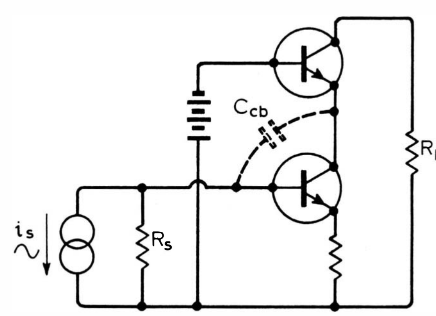


Fig. 33. The cascode circuit.

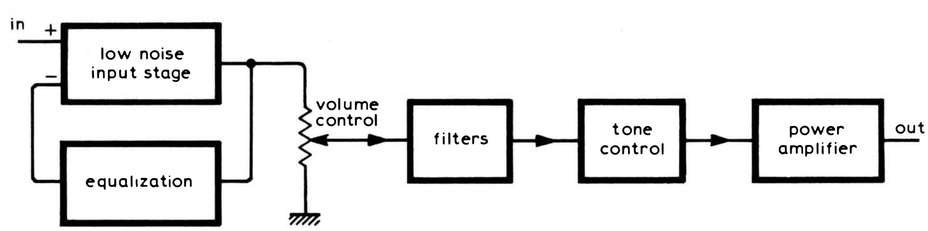


Fig. 34. Block diagram of a typical audio amplifier.

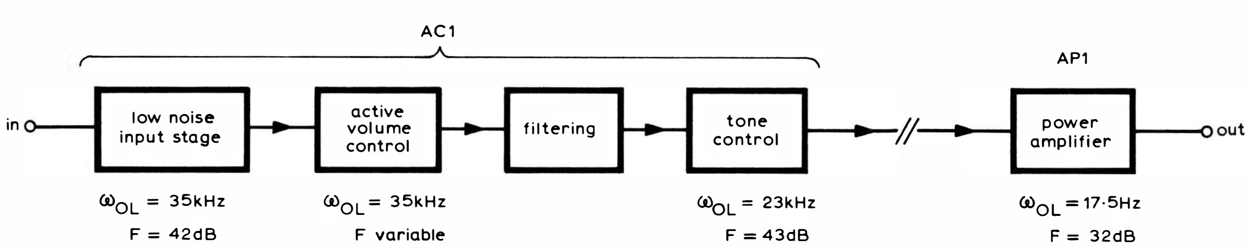


Fig. 35. Block diagram of the Lecson system showing the bandwidth and feedback factors for each section.

*The Lecson AC1 + AP1
†given output transistors e.g.

Now this is clearly a simple example and does not take account of perception thresholds or interactions of parameters and masking effects. It will not be sufficient to write, for example, $p(n)$ as the probability of detection of ndB signal to noise ratio, but possibly as a conditional probability of detection of weighted noise—say C.C.I.R. weighting—given a specific bandwidth.

So an f.o.m. which would be useful in the predictive design of audio components could be made up from tables of conditional probabilities and give a performance measure of universal use.

In this analysis I propose to use the thresholds of perception and non-neglectability ($p(x) = 0$ or 1) for all the parameters discussed so far and to discuss interpolation between these points.

In Table 1 a list is given of these parameters, and of thresholds which seem to be reasonable in the light of current knowledge.

Frequency response is treated by considering the two roll-off points—items 1 and 2—and determining a rough measure of $q(\omega_L)$ and $q(\omega_H)$ from Fig. 6 which are from results produced by Snow. Thus a response 20Hz – 20kHz has a partial M of 1 while 100Hz – 10kHz has a partial M of $0.9 \times 0.95 \approx 0.86$. Phase and amplitude linearity have been considered as being logarithmically interpolated in the absence of any other information—the same method has also been used by Mantel[3].

Steady state distortions are again interpolated logarithmically; this being chosen as a reasonable assumption in the absence of further knowledge. The whole basis of this experiment is to test the values and curves I have offered as a starting point.

I would suggest that at this starting point in the derivation of an f.o.m. the *a priori* measure of the likelihood of t.i.d.—the transient intermodulation index—be used, and the interpolation is as shown in the Table.

Signal to noise ratio is shown weighted according to the C.C.I.R. standard, and it is intended that this should include only hum and noise and not measures of crosstalk or other interfering signals.

Other parameters which have not been listed but are clearly essential when discussing elements of the chain other than amplifiers include a frequency modulation measurement to include wow, flutter and Doppler effects.

A review of the f.o.m. In the form proposed here it is possible to produce a single number which is intended to describe the subjective sound quality of a piece of equipment derived from objective measurements based on the following suppositions:

(i) it is possible to tabulate a conditional probability for the detection of any single shortcoming in terms of population.

(ii) that this probability will move in some way from 0 to 1 between the levels of perception and objection.

In order that a number may be derived, and that the behaviour of the f.o.m. may be investigated, I have used the thresholds discussed in this article, and tentatively proposed others with an interpolation. Clearly if such an f.o.m. is shown to give accurate results then it would be of great use to designers and users of audio equipment. However, in order that a f.o.m. of this kind can evolve, very extensive listening tests should be carried out. These are probably best controlled by and published through a respected journal such as *Wireless World*. [We are considering this.—Ed.]

The figures tabulated in Table 1 indicate that an amplifier which equals or betters the perception threshold for every parameter would have an $M_i = 1$. This rating would reduce to 0.9 with a low frequency cut off of 100Hz, or an amplitude deviation of 0.4dB or a phase deviation of 8° etc.

Table 1

Measurement	$p(x) = 0$ $q(x) = 1$	$p(x) = 1$ $q(x) = 0$	Interpolation
1. Amplitude-frequency response lower −3dB point ω_L	20Hz	1kHz	See Fig. 6 rating /10
2. Upper −3dB point ω_H	20kHz	1kHz	
3. Amplitude linearity $\pm L$dB*	0.25	30dB	$p(L) = 0.48 \log_{10} (4L)$
4. Phase linearity $\pm\theta°$	5°	100°	$p(\theta) = 0.77 \log_{10} (0.2\theta)$
5. Maximum weighted t.h.d. or i.m.d.* d%	0.1%	50%	$p(d) = 0.37 \log_{10} (10d)$
6. Transient intermodulation index (t.i.i.) ti	0.1	100	$p(d) = 0.33 \log_{10} (10ti)$
7. Rise-time $\tau\mu$s	5μs	1ms	$p(\tau) = 0.44 \log_{10} (0.2\tau)$
8. C.C.I.R. weighted s/n* n	70dB	30dB	$p(n) = (1-(n-30)/40)$
9. Cross-talk c	60dB	0dB	$p(c) = (1-n/60)$

*In the band 20Hz–20kHz or ω $-\omega$ whichever is the smaller. Note $0 \leqslant p(x) \leqslant 1$ only.

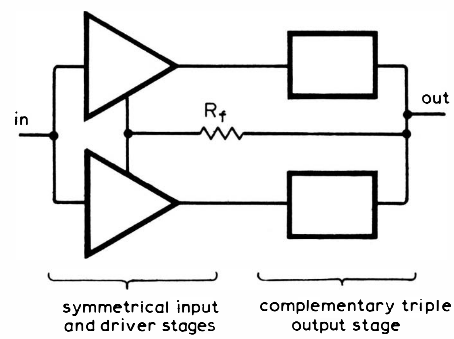


Fig. 36. Simple block diagram of the Lecson power amplifier.

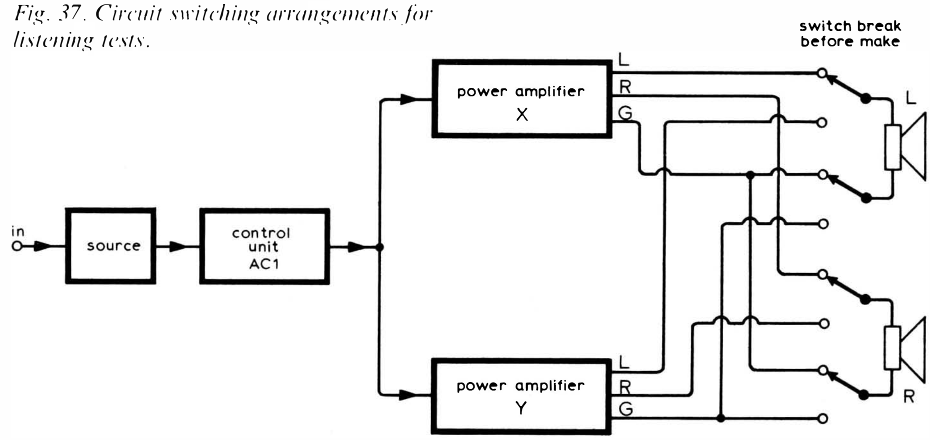


Fig. 37. Circuit switching arrangements for listening tests.

Some experiments

The author has recently carried out some listening experiments in an attempt to measure the significance of t.i.d. in high quality power amplifiers and, while the tests are not completed, some preliminary results have been obtained which are of interest.

The approach has been to use the basic Lecson AP1 power amplifier design and to vary only the open-loop bandwidth and feedback factor.

Three amplifiers were used:

1. The standard amplifier with an open loop bandwidth of 17.5kHz and feedback factor of 32dB as summarised earlier.

2. A modified version with an open-loop bandwidth of 4kHz and feedback factor of 40dB. This amplifier exhibited amplitude and phase responses identical to the first example, within the accuracy of the measurements (0.25dB, 2°), and showed t.h.d. results within 10% weighted of the first example.

3. A modified version with an open-loop bandwidth of 17.5kHz and a feedback factor of 6dB. This amplifier exhibited t.h.d. of 0.11% 50Hz – 3kHz, rising to 0.18% at

Appendix

The three amplifiers used in these listening experiments were all of very high quality showing an f.o.m. based on the routine of Table 1, of 0.9, 0.82 and 0.83, for the amplifiers A, B and C respectively.

Test 1. The test routine was performed using a panel of 8 listeners. Programme was derived from a very high quality disc player and monitor-standard loudspeakers employed (Spendor BC3).

Comparison A and B. On all programme material chosen, amplifier A was preferred by 87% of the listeners. The reaction of all listeners subjectively defined a clear difference, A being preferred for greater clarity at high frequencies. On switching to B the impression was obtained of a veil being drawn over the sound, particularly with strings or percussive material.

Comparison A and C. All listeners observed audible differences; C was preferred by 62% on all programme material and by 75% on folk music or percussive music. The overall impression was that C handled transient material very well but showed slight high frequency colouration, possible due to the weighted distortion.

Comparison B and C. Of the total audience, 75% preferred C on all material. Of particular interest in this test was that the two amplifiers showed subjectively different balance, C sounding to have less high frequency content than B. Also it was noted that B showed up background noise on the disc—hiss, clicks and pops—much more than C.

Test 2. A panel of 4 listeners using a high quality disc source and monitor-standard loudspeakers (Lecson HL1).

Comparison A and B. All preferred A for reasons of high frequency clarity.

Comparison B and C. This test produced confusion, no direct results were applicable as preference depended upon the material used. The faults of amplifier B on transient sounds seemed to be contrasted with a slight lack of clarity on high notes with amplifier C.

Tests 3 and 4. Devised as a control test for the comparison B and C. Two panels took part, consisting of three and seven listeners respectively. Again a disc source was employed and small loudspeakers used (Spendor BC1). In the first test C was unanimously preferred, to the second—as before, preference depended to an extent on the source material.

A working hypothesis to explain reactions to amplifiers B and C could be, that subjectively the amount of t.i.d. produced by B was as significant as any high frequency t.h.d. or i.m.d. produced by C. However these listening tests are only the beginning of a serious programme of tests which will aim to establish significance over a much wider range, and so these results can only be considered to be provisional. For example no attempt has been made to establish an f.o.m. for the loudspeakers used in these tests or to calculate or measure any interactions in the reproducing systems.

20kHz, 35W r.m.s. The distortion was such that the second harmonic was 40dB above any other so the weighted t.h.d. was below 0.2% at all times.

In each case the output impedance at the terminals of the amplifier was less than 250mΩ 20Hz – 20kHz, so any effect that a change of feedback factor may have had on this, was swamped by the 3m long loudspeaker leads used.

Three experiments were conducted, two formal, one informal. In each case the amplifying equipment was arranged as Fig. 37; only two amplifiers are used in any one test and both are driven continually by the pre-amplifier. Instantaneous comparison on programme is made by switching the loudspeakers between the two power amplifiers.

In accordance with the testing procedures laid down by Percy Wilson the participants had no knowledge, until the end of the experiments, of the nature of the differences between the amplifiers (if any) nor of the kind of subjective difference (if any) to expect. At no time was it asked which of the amplifiers sounded most natural, but simply "which of two, X or Y, do you prefer?".

The results of the tests are summarized in the Appendix. It is clear that, between amplifiers which are otherwise extremely good, despite relatively small changes to the t.i.i. performance, differences can definitely be detected by the ear as changes in the clarity and tonal balance of sound.

In a future article the author intends to describe further listening and objective tests and procedures in an attempt to quantify t.i.d. in absolute terms within the f.o.m. and with respect to t.h.d.

Conclusions

In these articles the author has attempted to study the relationships between objective tests made on amplifiers and the subjective results. Many aspects of amplifier performance have not been covered, the discussion concentrating more on distortions.

While it has been possible to outline in detail the rigorous compromises that face the designer of negative feedback amplifiers, the way in which each of the subjective effects trade-off is still not precisely known. A figure of merit calculation is given which makes an inquiring step in this direction, but it is clearly necessary that a programmed and controlled series of tests be carried out on a large scale.

Sixty Years Ago

From time to time over the years, successive editors of *Wireless World* have taken issue with the Post Office on the subject of licensing, especially when it has been considered that the Postmaster-General has tried to overstep the bounds of reason by claiming a proprietorial interest in the forces of nature. A correspondent in 1913 obviously felt very much the same way. . . . "We have heard lately of bedsteads and gas pipes being successfully used as substitutes for receiving aerials. Suppose I go a little further and discover that I get Paris, using only domestic appliances (such as a bedstead on an upper floor as an aerial, the wires of a piano suitably connected as a tuning coil, a nest of cake tins with buttered paper between them as a condenser, a piece of washing soda and a darning needle as a detector, and my tongue in place of the 'phones), must I obtain a licence from the Postmaster-General before I dare use such apparatus to get the time from E.L.? A few more discoveries(!) in "wireless" and we shall require to get a licence from the Postmaster-General before we furnish a house, and we shall have inspectors inspecting our pots and pans to see that they conform to the wireless regulations!"

Darts Game Calculator

Apprentices at the Guided Weapons Division of the British Aircraft Corporation, Bristol, have built an experimental automatic darts game calculating system which registers and keeps scores. It comprises a special dartboard with sensing devices, and a computer-controlled display unit which acts as the scoreboard. This unit, using 120 integrated circuits, shows the running totals for the competing teams and adds up each individual score.

The dartboard is designed with each segment internally divided and connected to the display unit. Impact of a dart on each segment causes an electrical signal to be sent to the computer in the display unit. The conclusion of an individual three-dart score is signalled by the removal of the darts from the board. The system is then re-activated by the next player standing on the throwing mat, under which is concealed a proximity detector. This causes a bulb on the display unit to be lit, showing that the system is ready to accept the next score.

Any variation of the game can be fed into the display unit before the game starts, so that the starting total could be set at, say, 1001 or 301 depending on the type of game to be played. If a double is required to start the game, then a "double" light is switched on and the system ensures that electrical signals from the dartboard will not alter the setting until the first double is obtained.

The apprentices were given just 13 weeks to design, build and test the project and were allowed to spend no more than £100 on materials.