Audio power amplifier design

There is nothing so practical as a really good theory — LUDWIG BOLTZMANN

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Articles describing particular amplifier designs, or advocating specific solutions to design problems, abound in the literature, and it is evident that some quite conflicting views exist on certain topics — for example, concerning the amount of negative feedback that should be used. The present approach is of a fairly broad nature, and aims to elucidate and compare various familiar and unfamiliar circuit techniques in such a way that their advantages and disadvantages may be clearly and logically appreciated.

IN EXPLOITING the very great virtues of negative feedback, the problems and difficulties that arise are largely those associated with obtaining adequate stability margins under all conditions of operation. In a.c. coupled amplifiers, there are stability problems at both low and high frequencies, but the elimination of output transformers, together with the adoption of d.c. coupled circuitry in most modern designs, has virtually removed the low-frequency problems.

Negative feedback and slew-rate limits

Other things being equal, the larger the amount of overall negative feedback applied to an amplifier, the lower will be the distortion. However, other things are quite likely not to be equal, since, to achieve stability, it is usually necessary to introduce elements which start attenuating the forward gain, with rising frequency, at a frequency which has to be made lower and lower as the amount of overall feedback is increased. If *unsuitable techniques* are used for effecting this attenuation, increased distortion will be generated in the forward path of the amplifier *at high frequencies*, to an extent which may more than offset the advantages of the increased feedback. Indeed, drastic high-frequency internal overloading may occur, and once this has happened, the overall feedback is powerless to preserve the wanted output waveform.

The rudimentary amplifier circuit shown in Fig. 1 will serve to illustrate the point. Here the capacitor C attentuates the gain with rising frequency by making Tr_2 function as a Blumlein integrator. The current, I, supplied by the first stage includes, in addition to a component flowing to Tr_2 base, a component much larger at high audio frequencies flowing to C. At such frequencies, and with Tr_2 producing a large output voltage swing, the current demanded by C may severely tax the output capability of Tr_1 stage, and may, in the limit, cause Tr_1 to overload, i.e. cut off during part of the cycle. Whether or not this will happen can be determined quite simply, on a sine-wave basis, by calculating the current in C, which is, nearly enough, V_{out}/X_c. If the peak value of this current exceeds the d.c. working current of Tr_1, gross distortion will occur. Thus the critical condition for the onset of such distortion is

$$I_{dc} = \hat{V}_{out} \times 2\pi f C \qquad (1)$$

This relationship may be rearranged to give a convenient formula for the critical sine-wave frequency, f_{crit}, above which gross distortion sets in no matter how much overall feedback there is. Thus

$$f_{crit} = \frac{I_{dc}}{2\pi C \hat{V}_{out}} \qquad (2)$$

It is customary nowadays, in the above context, to employ the slew-rate concept, though it is by no means essential to do so. This concept has long been familiar to workers in other fields, particularly those of servo-mechanisms and radar. As applied to amplifier circuits, the basic relationship is simply that, for a capacitor

$$dv/dt = i/C \qquad (3)$$

Thus, with reference to Fig. 1 again, suppose the transistor Tr_1 is briefly cut off; then a current approximately equal to I_{dc} is left flowing in R_c and most of this also flows in C, producing a positive-going rate of change of output voltage

$$[dv_{out}/dt]_{max\ poss} = I_{dc}/C \qquad (4)$$

This is called the *output slew-rate limit* of the amplifier, or sometimes, in commercial practice, just the slew-rate. With the single-ended input stage of Fig. 1, the slew-rate limit for negative-going outputs will be much more rapid than the above, because Tr_1 can turn on much more current than it can turn off. But when a balanced long-tailed-pair input stage is used, as in most integrated-circuit operational amplifiers, the slew-rate limits in the two directions will be approximately equal.

The relationship (4) applies whatever the signal waveform may be. If, at any instant, the demanded rate of change of output voltage exceeds this value, the amplifier will fail to follow it properly. Thus, if an amplifier has an insufficient slew-rate limit, then, every now and

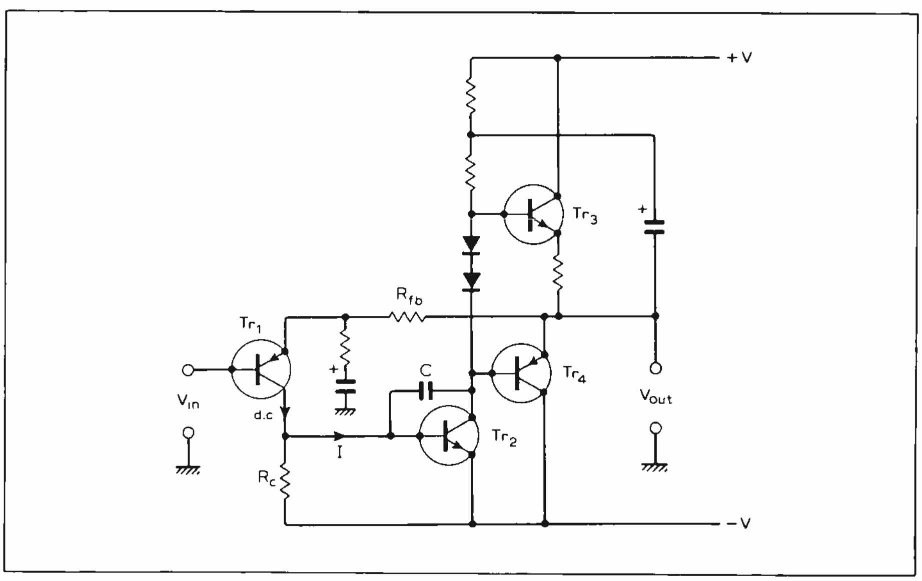


Fig. 1 *Rudimentary amplifier circuit in which the capacitor C gives rise to slew-rate limiting.*

then, on fast transients particularly, the slew-rate limit will be exceeded by the programme waveform. When this occurs, the amplifier gain will fall drastically, and all components of the signal being handled at that moment will be chopped, or modulated, by the transient. This effect, well known to enlightened designers of feedback amplifiers for decades, has nowadays, of course, become known as transient intermodulation distortion or t.i.d. (sometimes t.i.m.), as a result of several papers by M. Otala. Another, more recent, related term, due to W. G. Jung, is slewing induced distortion, or s.i.d.[1,2,3].

It is of interest to obtain the relationship between the general slew-rate limit formula (4) and the conditions which apply with sine-wave input. Substituting in (2) the value of I_{dc}/C given by (4) yields

$$f_{crit} = \frac{[dv_{out}/dt]_{max\ poss}}{2\pi\hat{V}_{out}}$$

$$\text{i.e.}\quad f_{crit} = \frac{\text{output slew-rate limit}}{2\pi\hat{V}_{out}} \qquad (5)$$

(This result can alternatively be obtained by differentiating the output voltage waveform, $v = \hat{V}\sin 2\pi ft$, and equating the peak instantaneous value of the differential coefficient to the slew-rate limit.)

In all the above, the slew-rate limit referred to is that of the amplifier *output* voltage, and this is the usual practice – especially in integrated circuit data sheets, where it is simply called the slew-rate. Thus, unless otherwise stated, slew-rate figures may be assumed to apply to the output of an amplifier. However, it is sometimes convenient to express them with respect to the input, which merely involves dividing by the amplifier's voltage gain. The corresponding equation to (5) for the input is

$$f_{crit} = \frac{\text{input slew-rate limit}}{2\pi\hat{V}_{out}} \qquad (6)$$

Consideration of (5) and (6) makes it evident that what is invariant is the quotient of the slew-rate limit and the peak sine-wave voltage at any selected point in the system. Hence, more generally,

$$I = I_o e^{\frac{qV_{be}}{kT}}$$

The peak voltage V is normally that for full output level. The quality of the slew-rate performance of an amplifier may thus be expressed by the slew-rate-limit figure given in *volts per micro-second per volt peak* of sine-wave signal. For example, f_{crit} = 20kHz corresponds to a figure of 0.126V/μs per volt peak.

It is of interest to consider what sort of output waveform would be expected from an amplifier suffering from slew-rate limitation, on sine-wave input. Suppose initially that the amplifier is basically as in Fig. 1, having a single-ended input stage which imposes a much more severe slew-rate limit for positive-going amplifier output voltage than for negative-going. Referring to Fig. 2(a), the sine-wave represents the wanted output waveform, and the broken line represents the maximum rate of change of output voltage of which the amplifier is capable, i.e. it represents the output slew-rate limit. The actual output therefore follows the wanted waveform from A to B, but after B it follows the path BCD before joining the wanted waveform again at D. The complete output waveform is thus as shown in Fig. 2(b). Fig. 3(a) shows some experimental waveforms obtained with a circuit having the basic configuration of Fig. 1, for two different degrees of slew-rate limitation overload on sine-wave input. Fig. 3(b) shows the output waveform for square-wave input, and is a typical result for an amplifier exhibiting unsymmetrical slew-rate limitation.

The waveforms of Fig. 4 were obtained using a type LM301AN integrated circuit operational amplifier as a unity-gain inverter. The 301 circuit, very broadly speaking, has a similar type of configuration to that shown in Fig. 1, but with a balanced long-tailed-pair input stage arrangement. The external stabilizing capacitor C, more often called the compensation capacitor, had a value of 30pF. It will be seen that, as expected, the slew-rate limitation is of a nearly symmetrical nature.

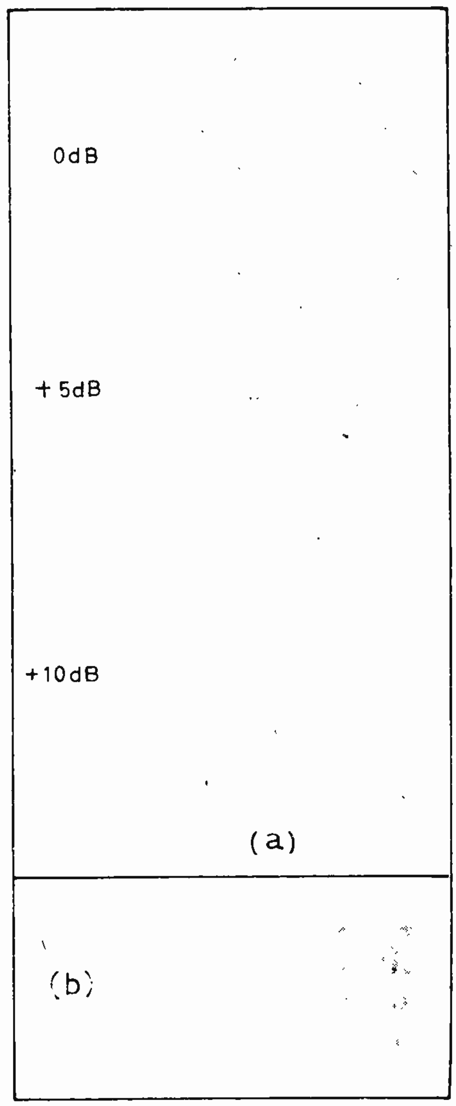
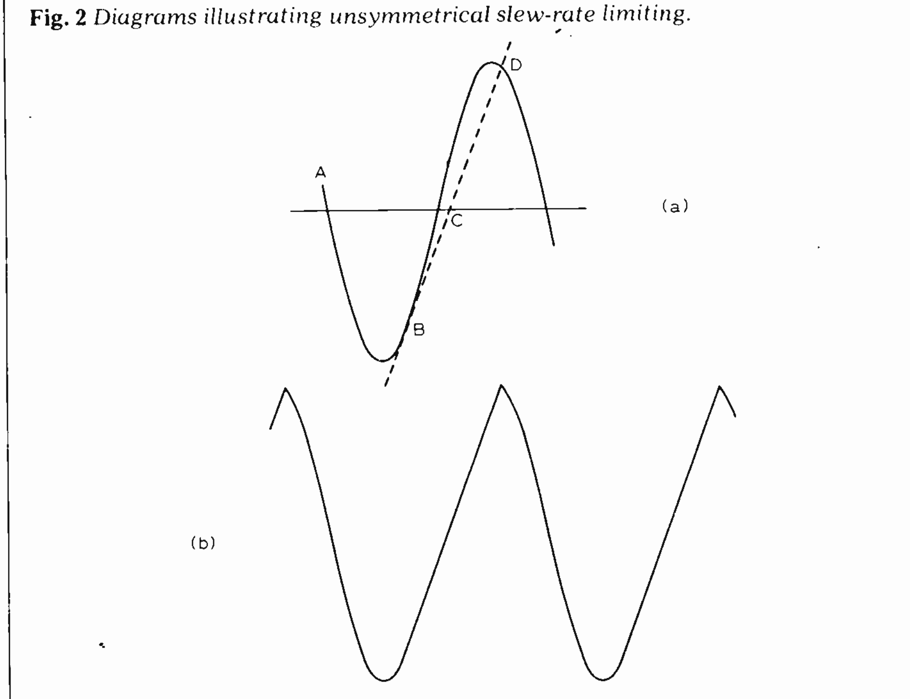


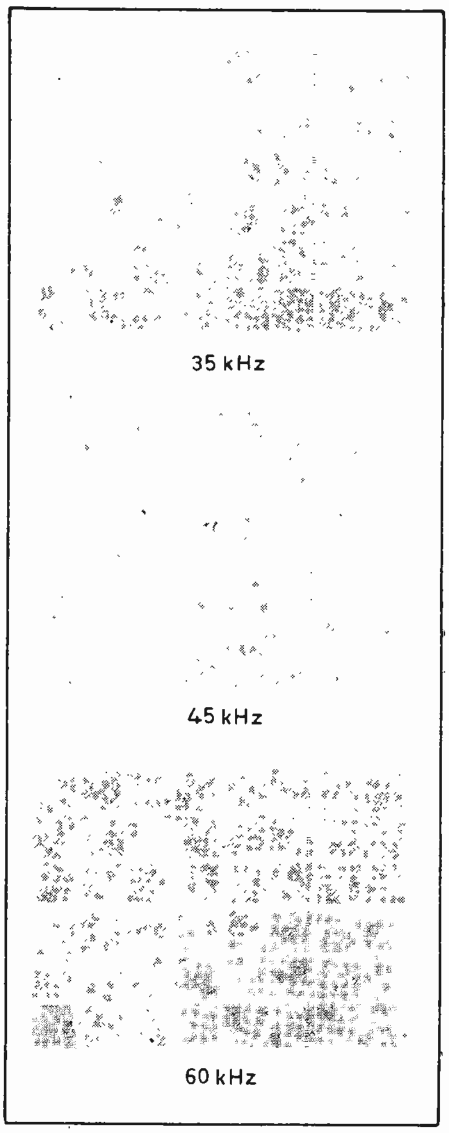
Fig. 3 *(a) Output voltage waveforms from amplifier exhibiting unsymmetrical slew-rate limiting, for three different levels of sine-wave input, all at the same frequency. (b) Output voltage waveform for square-wave input. The negative-going transitions are not slew-rate limited.*

Fig. 2 *Diagrams illustrating unsymmetrical slew-rate limiting.*



A great deal of attention has been given to this aspect of amplifier behaviour in recent years, and while it is certainly important to avoid significant distortion of this type, the notion that it is a fairly newly-discovered form of distortion is quite unjustified. It all boils down to the fact that, to avoid unwanted intermodulation effects, a good amplifier should be able propertly to track all normal programme waveforms, whether of a sustained-tone or a transient nature, without any internal circuits overloading in the process – surely an old and familiar notion? Indeed, I cannot do better than quote Jung, who says "there is nothing new, unique, or mysterious about slew-induced or transient intermoduation distortion"[2]. It may be added, however, that since some – but certainly not all – of the earlier transistor amplifiers suffered seriously from this type of distortion, the widespread attention that has been given to it is a good thing. But removal of significant s.i.d. is not a panacea – there are also other important causes of distortion.

Fig. 4 *Output voltage from integrated-circuit operational amplifier for equal-amplitude sine-wave inputs at three different frequencies, showing slew-rate limiting. Scales: 1V/cm, 5µs/cm.*



As considered above, the slew-rate-limit mechanism sets a fairly sharply defined threshold, beyond which there is a rapid onset of gross distortion that the overall feedback is powerless to control. Below this threshold output level, which is, of course frequency-dependent, the distortion will be negligible only if there is sufficient overall feedback. Whether there is enough feedback to give this result depends on the details of the particular design, but in some instances there may not be enough. Thus it is of interest to consider the distortion mechanisms that are operative in the milder situation where drastic overloading does not occur.

Referring to Fig. 1, suppose we decide to apply 6dB more overall feedback to the amplifier by reducing R_{fb}. This is likely to necessitate doubling the value of C, for equally satisfactory stability. Thus, while we succeed in doubling the feedback loop gain at low frequencies, where C has little effect, the loop gain at higher frequencies, where C is dominant, remains as before. At a given high frequency, and a given output voltage, Tr_1 will have to supply twice the current to the doubled value of C, and the percentage second-harmonic distortion generated in Tr_1 will go up by a factor of approximately 2*. Since the amount of feedback at the high frequency involved is the same as before, the amplifier output distortion (due to distortion in Tr_1) will also be doubled.

Because of the doubling of the C value, the critical frequency for slew-rate limitation, above which full output ceases to be obtainable without drastic overload, is halved – see equation (2).

Quite frequently a long-tailed pair, or differential input stage, will be used in place of the single transistor Tr_1 shown in Fig. 1, and then, if well balanced, the dominant distortion introduced will be third-harmonic, the percentage distortion being proportional to the square of the output current [5]. (This is a characteristic of any device, e.g. a tape recorder, in which cube-law curvature is dominant.) Thus, with the low-frequency overall feedback increased by 6dB, and with C doubled as before, the third-harmonic distortion generated in the input stage will be up by a factor of 4 at high frequencies, as also will be the amplifier's output distortion due to this cause.

We thus have the situation that increasing the amount of low-frequency overall feedback, with corresponding adjustment of the stabilizing capacitor value, increases that part of the high-frequency output distortion which is due to smooth-curvature non-linearity distortion in the input stage. In many cases, below the true slew-rate-limitation overload point, this will be the main cause of distortion at high frequencies. However, with suitably modified circuit designs, to be described later, the input stage distortion may be fairly negligible.

It is interesting to consider how the above non-overloading type of distortion would be expected to vary with frequency. A long-tailed-pair input stage will first be assumed. Since, at high frequencies, the current supplied by the input stage is proportional to frequency, the percentage third-harmonic distortion generated within the stage is proportional to the square of the frequency. But because the overall-feedback loop gain is halved for each doubling of frequency, the distortion at the output of the amplifier, due to this mechanism, is proportional to the cube of the frequency. The percentage output distortion is thus proportional to $V_{out}^2f^3$, as established by Jung. The corresponding result for a single-ended input stage, as in Fig. 1, is that the percentage output distortion, now mainly second-harmonic, is proportional to $V_{out}f^2$. This is because in any device in which square-law curvature is dominant, the percentage distortion is directly proportional to the output current or voltage.

It will thus be seen that a characteristic feature of distortion of the type discussed above, which occurs before the onset of true slew-rate-limitation overload, is that it increases quite rapidly with frequency. Fig. 5 shows the ideal cube-law relationship deduced above for the balanced input stage case. With a single-ended input stage, though the rise in distortion with frequency is more gradual, the magnitude of the distortion is liable to be much greater [5].

Jung calls the input-stage-originated distortion that occurs before the onset of true slew-rate limitation "Category I slewing induced distortion", the gross distortion that occurs at higher levels being "Category II s.i.d." It is important not to let this terminology disguise the fact that Category I s.i.d. is, after all, just straightforward input-stage smooth-curvature non-linearity distortion, which may become significant at high frequencies because of the increased current demanded from the input stage and the reduced amount of overall feedback in action.

* The percentage second-harmonic distortion produced by an ideal voltage-driven transistor, having a characteristic $I=I_0 \exp qV_{be}/kT$, approximately $25\times(\hat{I}/I_{dc})$, where $\hat{I}$ is the peak value of the signal-current fluctuation and I_{dc} is the d.c. working current. Another convenient fact is that, at any working current, the percentage second-harmonic distortion is equal to the peak value, in millivolts, of the signal voltage applied between base and emitter[4,5].

Though, as shown in Fig. 5, the high-frequency distortion due to the input stage rises rapidly with the measuring frequency applied, it should not be imagined that the harmonics generated at any one measuring frequency are boosted according to their order, in any comparable manner. Consider first the effects that would occur with the overall feedback disconnected. Referring again to Fig. 1, the harmonics in the current fed by the input stage to the Tr_2 stage will be attenuated in this stage in proportion to their order, because of the integrating action of the capacitor C. Thus, with the feedback loop open, the harmonics in the amplifier output voltage, due to input stage distortion, would fall off in amplitude with increasing order at a rate 20dB/decade (6dB/octave) more rapid than that applying directly to their generation in the input stage. However, with the overall feedback loop closed, and because the amount of feedback at high frequencies falls off at 20dB/decade with increasing frequency – assuming C is the only cause of loop gain attenuation – the final output distortion spectrum will have the same relative amplitudes of fundamental and harmonics as for the input stage by itself. With a long-tailed-pair input stage, and assuming the circuit not to be operating too close to the slew-rate limit point, the dominant harmonic will be the third, the higher order harmonics decaying rapidly with increasing order. Thus the type of distortion generated is relatively innocuous compared with the worst forms of cross-over distortion. The important thing is simply to arrange the design so that the magnitude of the distortion does not become too high.

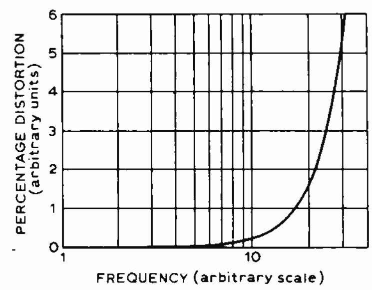


Fig. 5 *Theoretical variation of third-harmonic distortion with frequency for amplifier with long-tailed-pair input stage, when operating below the slew-rate limit.*

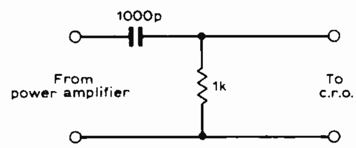


Fig. 6 *Simple differentiator circuit used in tests. The output is 1V for an input rate of change of 1V/μs.*

Slew-rates of programme waveforms

Gramophone records are frequently used as the programme source when subjective judgements of the performance of audio equipment are being made, so that it is of interest to know the order of slew-rate to be expected at the output of a high-grade RIAA equalized amplifier. This can easily be determined using a very simple differentiator circuit such as that shown in Fig. 6. This circuit is fed from the output of the power amplifier, and, with the values shown, gives an instantaneous output of 1 volt when the input slew-rate is 1V/μs. The objection may well be raised that the slew-rate limit may degrade the true slew-rate of the source, i.e. the pickup, but whether or not this is the case may be discovered by replacing the pickup by an oscillator and thus determining the slew-rate limit of the amplifier system. With good equipment, this will be found to be much higher than the slew-rate obtained with records.

The experimental procedure adopted was as follows. First a frequency test record was used to check that the system had a flat frequency response, within ±1dB, up to 12kHz. Then a suitable music record was selected, and the system gain was adjusted so that the input to the Fig. 6 circuit occasionally reached peak values of ±10V, but not more. The c.r.o. was then transferred to the differentiator output, the record replayed, and the maximum output voltage excursion from the differentiator during the replay was determined. The test was done with a wide variety of records, including one of the Sheffield direct-cut discs. The largest instantaneous outputs from the differentiator were caused by occasional dust clicks, and went up to over 0.40V, but on the music they never exceeded about 0.14V. The latter corresponds to a slew-rate of 0.14V/μs, which is the peak instantaneous slew-rate of a sine-wave with amplitude ±10V and frequency approximately 2.2kHz.

The implication of the above is that an amplifier with f_{crit} = 2.2kHz, i.e. capable of giving full output on sine-waves up to 2.2kHz, without suffering from slew-rate limitation, and with sufficient freedom from ordinary non-linearity distortion, will reproduce such records entirely satisfactorily. I can almost hear some readers saying "this is ridiculous – it's well established that amplifiers must be free from slew-rate limiting, at full output level, up to at least 20kHz"! But has this, or anything approaching it, in fact, been properly established? I do not think so. But because of such doubts, it is worth approaching the matter from a different angle, as follows.

The maximum instantaneous recorded velocities on records occur over the band extending from about 700Hz to, perhaps, 8kHz, and are normally in the region of 30cm/s[6]. Suppose the gain of an RIAA equalized replay system is adjusted so that a 1kHz sine-wave recording with 30cm/s peak instantaneous velocity gives an output voltage of 10V peak. Since for a sine-wave voltage with peak value $\hat{V}$, the peak rate of change of voltage is $\hat{V} \times 2\pi f$, the peak rate of change of voltage for a 1kHz sine-wave of peak value 10V is 0.063V/μs. It is probably fairly unusual for a peak velocity of 30cm/s to be recorded at a frequency as high as 8kHz, but if this did happen, then, ignoring for the moment the effect of the RIAA equalization, the output slew rate would be 8 × 0.063, i.e. 0.50V/μs. However, at 8kHz, the RIAA equalization introduces a loss of 11.7dB (×3.85) relative to the response at 1kHz, so the figure of 0.50V/μs is reduced to approximately 0.13V/μs. This, it will be seen, ties up surprisingly well with the experimentally determined figure, mentioned above, of 0.14V/μs.

The Fig. 6 differentiator was also used with a master tape recording of violin music with piano accompaniment, thought to be of unusually good fidelity. When adjusted to give a peak replay voltage of 10V as before, the peak instantaneous differentiator output voltage observed was 0.083V, so that the peak slew-rate was 0.083V/μs. A 10V peak sine-wave of 1.3kHz has this same slew rate.

Similar tests done with programme from an f.m. tuner yielded generally equivalent results as far as the actual audio waveform was concerned, but with the complication that, on stereo transmissions, owing to imperfect filtering in the tuner, the (L-R) sidebands greatly increased the peak dv/dt value at the differentiator output, a figure of about 0.4V/μs being obtained with the audio level at ±10V as before. By using the 10kHz filter in the audio control unit, the f.m. multiplex waveform was almost eliminated, the peak slew-rate of the remaining audio waveform being about 0.15V/μs. It is clear that without the filter, the minimum acceptable slew-rate limit in the audio amplifier would be determined largely by the amount of f.m. multiplex waveform present in the tuner output, since unpleasant intermodulation effects can occur if the amplifier is unable properly to follow this waveform. The amount of multiplex waveform in the output of f.m. tuners varies a great deal from one make to another.

The above quite low slew-rates will seem less surprising when it is remembered that the success of the pre-emphasis and de-emphasis schemes universally used in both recording and f.m. broadcasting systems is dependent largely on the fact that the high-frequency components of all normal audio waveforms are of much smaller amplitude than the lower frequency components.

Necessary amplifier slew-rate limit

Provided an amplifier is not overloaded, and provided it has sufficient feedback to make the distortion when not slew-rate limiting adequately low, there is certainly no absolute necessity for the slew-rate limit of the amplifier to be any larger than the maximum rate of change, or slew-rate, of the waveforms handled by it. This point needs emphasising, for reading Jung's interesting articles can easily make one jump to the conclusion that there is a *fundamental* need for the amplifier slew-rate limit to exceed the maximum rate of change of the programme waveform by a large factor. That this cannot possibly be true may be seen by imagining, or actually making, an amplifier with the same broad configuration as in Fig. 1, but in which Tr_1 is replaced not by a simple long-tailed-pair, but by a more complex circuit having a large amount of internal feedback. Then the distortion of the part of the amplifier that precedes C will remain extremely low right up to the slew-rate-limit overload point. Such an amplifier will fail to satisfy Jung's "new slew-rate criterion" by a very large factor, and yet, provided the distortion in the output stage etc. is sufficiently low, it will give no subjectively detectable quality degradation on any normal programme material.

With an ordinary long-tailed-pair input stage, the distortion introduced by it will be mainly third-harmonic, with the higher-order harmonics well subdued, provided the amplifier slew-rate limit is made higher than the maximum slew-rate of the programme by a reasonable factor, say two or three times. The distortion will then be of much the same character as that introduced by a good tape recorder, but will be of appreciable magnitude only at high audio frequencies. Provided the distortion is held down to a reasonably low magnitude – well under that of a recording system, to be on the safe side – by sufficient overall feedback, it will not be subjectively detectable. □

References

1. Jung, W. G., Stephens, M. L. and Todd, C. C., "Slewing induced distortion in audio amplifiers", Feb. 1977 articles series preprint, *The Audio Amateur*, Box 176, Peterborough, New Hampshire 03458 (USA).
2. Jung, W. G. Stephens, M. L. and Todd, C. C., "Slewing induced distortion and its effect on audio amplifier performance – with correlated measurement/listening results," AES Preprint 1252, AES Convention May 1977.
3. Jung, W. G., "Slewing induced distortion," *Hi-Fi News*, Nov. 1977, pp.115-123.
4. Baxandall, P. J., "Low distortion amplifiers – Part 2," *J. British Sound Recording Association*, Nov. 1961, pp.246-256.
5. Taylor, E. F., "Distortion in low-noise amplifiers," *Wireless World*, August 1977, pp.28-32.
6. Kogen, J. H., "Gramophone-record reproduction: development, performance and potential of the stereophonic pickup," *Proc. IEE*, vol. 116, No. 8, August 1969, pp.1338-1344.

Correction

In the article "Audible amplifier distortion is not a mystery", in the November 1977 issue, the editor inadvertently omitted two resistors from the circuit diagram (p.65). These should be inserted one in each input lead of the operational amplifier near the right-hand side of the diagram.

Communications tests with moving trains

TESTS WITH radio communication between signal boxes and trains were carried out in this country before the second world war, mainly on the London & North Eastern Railway. The equipment available at that time was relatively bulky, required considerable power and did not meet acceptable standards of reliability. Advances in mobile radio engineering have changed the situation, and since the middle 1960s most of the major European railways have investigated systems of radio communication with moving trains. Between 20,000 and 30,000km of route on the Continent are now equipped in this way or are awaiting delivery of systems in course of manufacture.

British Railways studied the subject in connection with the Channel Tunnel project, and although that is in abeyance it has been decided to proceed with a scheme on the recently electrified section of the Eastern Region between London, Welwyn and Hertford North. Some details of the project were given in a paper presented to the Institution of Railway Signal Engineers in London on 2 November by J. Boura and C. Kessel of the British Railways Board.

Each signalman in the King's Cross signalbox will control a group of uhf transmitting and receiving stations spaced at intervals at the line-side so as to cover the area for which he is responsible. A radio channel of four frequencies will be allotted to each area, comprising three transmit frequencies chosen to avoid mutual interference and one receive frequency. Transmit frequencies will be used in cyclic transposition along the line. The train receiver will incorporate a search and lock system by which it will lock on to the first satisfactory signal it receives from a lineside station. When the signal lever falls to 2μV it will search for another transmitter and lock again, but if an acceptable signal is not found a 'carrier fail' alarm will be displayed in the cab.

The use of synchronised transmitters within the groups was considered but would have required an accuracy of 30Hz in 450MHz and was not practicable in this situation. In passing from zone to zone a driver will reset his transmitter to the new receive frequency by pushbutton. Automatic returning could have been provided but would have increased the cost by about 60 per cent. Data transmission at 600 bauds will be used for establishing calls and the display of standard messages in the form of picturegrams. A speech circuit will also be provided.

At the signalbox the radio system will be linked with the existing computer-based train describer system which displays train identification numbers on the mimic diagram in their appropriate positions. A small computer in the signalbox radio installation will interrogate the describer computer to find the train identification number corresponding with the call signal received from a train and show both numbers in a queue type display of incoming calls on a VDU. The call signal will be unique to a particular set of vehicles, while the train running number changes according to the service the set is providing.

Contracts for the radio equipment have not yet been placed. When the paper was presented a somewhat similar system now being manufactured for the German Federal Railway by Telefunken was demonstrated. In the ensuing discussion there was some emphasis on the need to balance sophistication with reliability and cost. At present communication between trains and signalmen relies on signal post telephones, but these are specialised instruments manufactured in small numbers for the railways alone. The rapid expansion of the mobile radio and computer-linked data transmission businesses seems to hold hope of costs coming down in this area. □

SIXTY YEARS AGO

IN AN age when even resistors are of many types and integrated circuits continue to proliferate, the following piece, from our January 1918 issue, is seen to be prophetic. Prof. Pupin evidently did not understand that insufficient bafflement of the laity was to be obtained from plain speech.

"The scientist in question, Professor M. I. Pupin, said that if there must be a new name for each new detector — a new name for everything that comes up in the course of the development of the electrical art — pretty soon the science of electro-technics will be a mass of new names, and the learning of the names will be much more difficult than the learning of the facts connected with the art.

Today the following words are in common use by radio engineers, as the names of devices in appearance similar to and in principle based upon the original audion: Oscillation valve, regenerative audion, kenotron, pliotron, electron, relay, thermionic relay, thermotron, audiotron, amplitron, detecto-amplifier, Moorhead tube, oscillion, ultra-audion, dynatron, oscilaudion and pliodynatron.

After reading the foregoing, is it any wonder that Doctor Pupin was perturbed over the advent into the electrical art of new and mongrel names? When Doctor de Forest coined the word 'audion' he pulled the bung from a barrel which contained a vast and venerable assortment of Greek and Latin derivatives, and it is evident that these have been industriously raked up and picked over to supply bewildering additions to our already involved scientific vocabulary. Here in England we are not so fond of inventing new names, although scientists have not settled down to any one title for these particular devices. In the Services, where large numbers of these instruments are in use, we believe it is the custom to refer to them simply as 'valves', fancy names being debarred altogether." □