Measuring transient intermodulation in audio amplifiers

The 'inverting-sawtooth' method for low t.i.d. measurements

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Over the last ten years transient intermodulation distortion (t.i.d.) has attracted considerable interest in audio engineering circles, as a glance at the bibliography shows. Among the many published papers on the subject a number deal with the measurement of t.i.d.

The best known method consists of feeding sine waves, superimposed onto square waves, into the amplifier under test. The output spectrum is then examined using a spectrum analyser and compared to the input. This method suffers from serious disadvantages: the accuracy is limited, the measurement is a rather delicate operation and an expensive spectrum analyser is essential.

Recently, a new approach has been described by S. Takahashi and S. Tanaka which is, in their own words, simple yet precise. This method, which we will refer to as the "inverting sawtooth" method, is also fast, cheap – it requires nothing more sophisticated than an oscilloscope – and sensitive – and it can be used down to t.i.d. values as low as 0.002% in high power amplifiers.

Transient intermodulation distortion is an unfortunate phenomenon associated with negative-feedback amplifiers. When a feedback amplifier receives an input signal which rises very steeply, i.e., it contains high-frequency components, the feedback can arrive too late so that the amplifier overloads and a burst of intermodulation distortion will be produced, as in Fig. 1.

Since transients occur frequently in music this is obviously a problem for the designers of audio amplifiers. Unfortunately, heavy negative feedback is frequently used to reduce the t.h.d. (total harmonic distortion) of an amplifier, which tends to aggravate the transient intermodulation (t.i.m.) situation.

Method of measurement

The 'inverting-sawtooth' method of measurement is based on the response of an amplifier to a sawtooth waveform. The amplifier has no difficulty following the slow ramp but it cannot follow the fast edge. The output will follow the upper line in Fig. 2, cutting off the shaded area and thus increasing the mean level. If this output signal is filtered to remove the sawtooth, a direct voltage remains which indicates the amount of t.i.m. distortion, although it is difficult to measure because it is indistinguishable from the d.c. offset of the amplifier. This problem is neatly avoided in the i.s.-t.i.m. method by periodically inverting the sawtooth waveform at a low audio frequency as shown in Fig. 3. In the case of the sawtooth in Fig. 2, the mean level was increased by the t.i.m. distortion; for a sawtooth in the other direction the opposite is true.

The result is an a.c. signal at the output whose peak-to-peak value is the t.i.m. voltage, which can be measured easily with an oscilloscope.

Practical measurements

The equipment needed for i.s.-t.i.m. measurement is shown in Fig.4. A 20kHz sawtooth generator, its output inverted every 256 cycles, is followed by a high pass filter which attenuates the 78 Hz switching component by more than 100dB. A suitable circuit, shown in Fig. 5, is a straightforward, 36 dB/oct. Butterworth filter, with a cutoff at around 1 kHz. The circuit contains a simple RC network to limit the maximum signal slope to a reasonable value. It can be switched to supply signals of varying severity to the test amplifier: for most purposes the $f_c = 30$kHz position gives realistic results but for "super-fi" amplifiers the 100 kHz position can be used. An intermediate position, not normally used, is provided.

After leaving the amplifier under test, the 20 kHz sawtooth must be filtered out so that the t.i.m.-induced voltage can be measured. The passive low-pass network in Fig. 6 gives the desired response.

Finally, the filtered output signal is displayed on an ordinary oscilloscope. If the peak-to-peak value of this signal and the peak-to-peak value of the inverting sawtooth are measured, the t.i.d. can be found very simply from:

$$\text{t.i.m.} = \frac{V_{\text{out}}}{V_{\text{sawtooth}}} \times 100$$

The two oscilloscope photographs show the waveforms actually observed. The top one shows part of the 20kHz inverting-
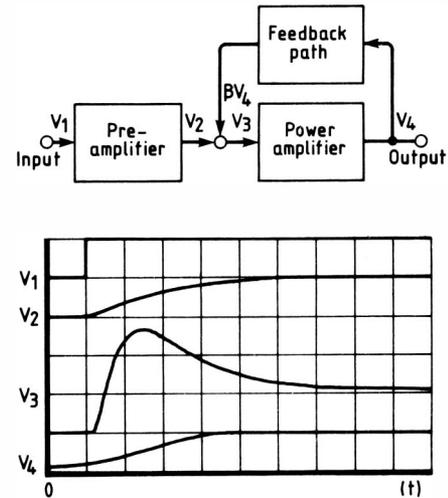
Fig. 1. Response of negative-feedback amplifier to very steep input slope
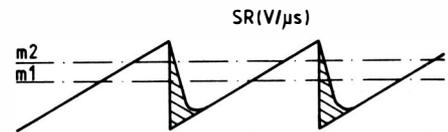
Fig. 2. Input sawtooth and response of amplifier
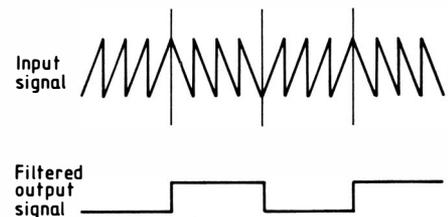
Fig. 3. Inverting sawtooth and filtered output of amplifier under test

Fig. 4. Block diagram of measurement setup
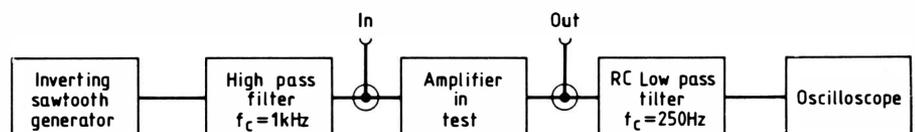
sawtooth waveform at the output of the generator, while the second shows a typical filtered output waveform (20kHz/256 = 78Hz) which is used to measure the peak-to-peak amplitude of the t.i.m.-induced voltage.

Inverting sawtooth generator

To generate the special inverting-sawtooth waveform we designed the simple circuit shown in Fig. 7. An ordinary sawtooth signal is generated by a relaxation oscillator consisting of the constant current generator Tr_1, a capacitor C1, inverting triggers IC_{1a} and IC_{1b} and an analogue switch, I_{2a}. Capacitor C_1 is charged by the constant current generator until the voltage across it reaches the upper threshold of the trigger IC_{1b}, which is about 6.5V. This closes the analogue switch and discharges C_1. Discharging continues until the voltage across C_1 falls to the lower threshold of the trigger, about 3V, when the analogue switch opens and C_1 charges again. The frequency of the resulting sawtooth waveform is adjusted to 20kHz by the trimmer in Tr_1 emitter.

The buffer, Tr_2, minimizes the loading on C_1 and attenuates the signal to avoid saturating the phase-splitter that follows. The phase-splitter, Tr_3, provides two out-of-phase sawtooth waveforms, the trimmer in the collector of Tr_3 adjusting the symmetry of these waveforms. Another trimmer in the inverted signal decoupling network, adjusts the relative offset of the two waveforms.

The analogue switches IC_{2b} and IC_{2c} select either the direct or the inverted sawtooth under control of the counter, which divides the discharge pulses from the relaxation oscillator by 256, so that the out-

Inverting sawtooth is shown in top trace; bottom picture is t.i.m.-induced voltage after filter.
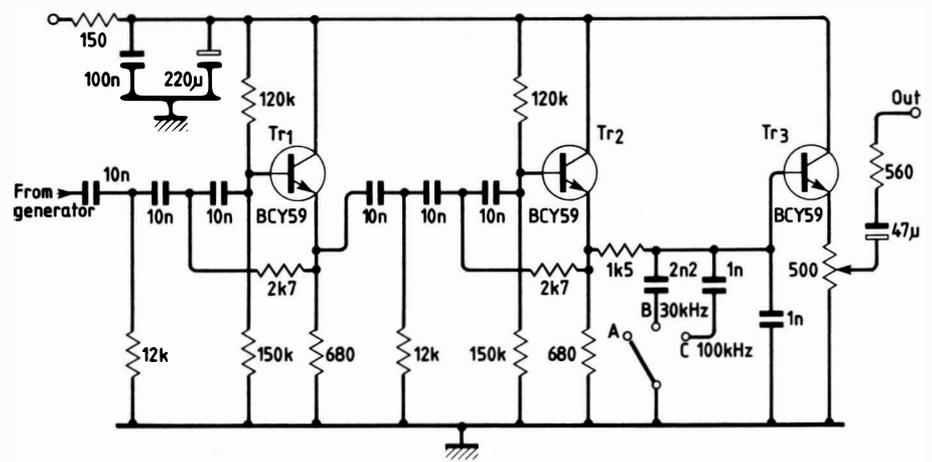
Fig. 5. High-pass filter and signal slope limiting network
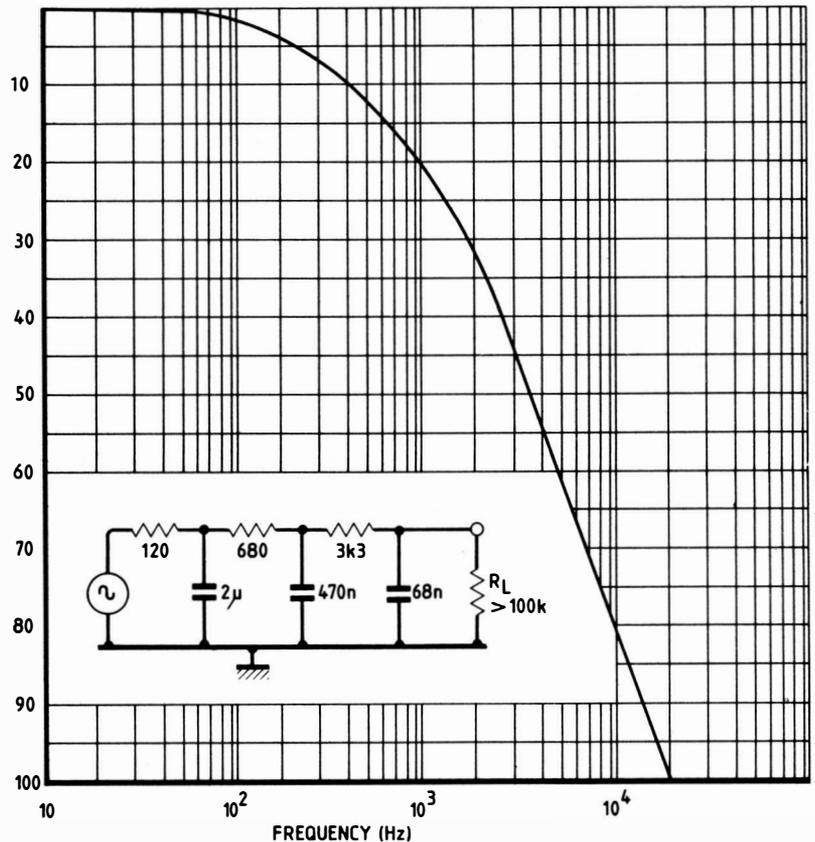
Fig. 6. Low-pass filter characteristic

put sawtooth signal changes phase every 256 cycles. The inverting sawtooth is buffered by Tr_4.

The output of the counter also serves to synchronize the oscilloscope used in the measurements. Without this sync., it would be virtually impossible to observe the inverting sawtooth waveform.

Some results

TO see how the inverting sawtooth method works in practice, we have tested a variety of audio integrated circuits – standard operational amplifiers, monolithic power amplifiers and an RIAA preamplifier based on a new high-quality preamplifier i.c. For the t.i.d. measurements on the operational amplifier, a unity-gain buffer, shown in Fig. 8, was used to match the low

impedance filter to the op-amp. input. Figure. 9 shows the results obtained from an LS148 op-amp. with three different values of compensation capacitors.

These results show t.i.d. values higher than those obtained using other methods – a result of the greater sensitivity of the i.s.-t.i.m. technique. Extensive comparison of t.i.d. measurements using various methods have been published elsewhere and confirm the validity of the inverting-sawtooth method. Although it is possible to measure t.i.d. as low as 0.002% this only applies to high-power amplifiers when the t.i.m. voltage can be measured more easily.

Figure 10 shows typical t.i.d. values for a 15W monolithic amplifier, the TDA 2030, in the test circuit, Fig 11. As in the case of the operational amplifier the measurements were carried out at the three different settings of the signal slope
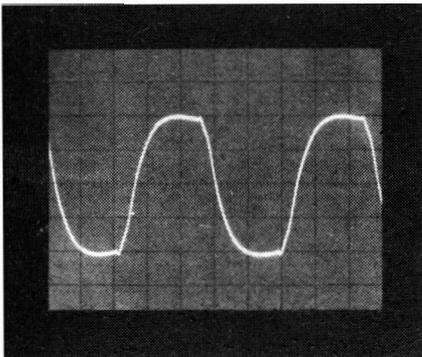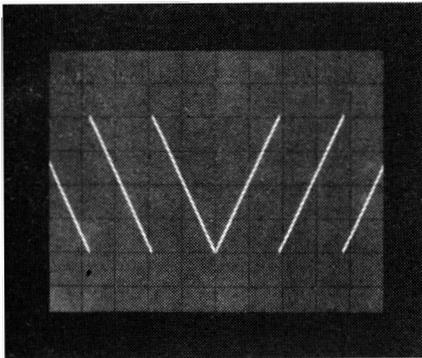
Fig. 7. Inverting sawtooth generator

Fig. 8. Unity-gain buffer

Fig. 9. T.i.d. measurements on LS148 op.-amp.

Fig. 10. T.i.d measurements on TDA2030 audio power amplifier

Fig. 11. Test circuit used for TDA2030/TDA2040

Fig. 12. R.I.A.A. preamp. based on TDA2310 (one channel)

third harmonic – is that due to the dependence of the resistance of the thermistors used to control the amplitude of the oscillation on the instantaneous value of the signal potential applied to them. This characteristic of oscillators with averaging control systems has been analysed by Robinson[5] who suggests that the distortion of such a system, which is shown to be mainly third harmonic, will be

$$\frac{x_3}{x_1} \simeq \frac{1}{8\eta} \cdot \frac{A_0-\eta}{\eta} \cdot \frac{1}{2\pi fT}$$

where $(A_0-\eta)/\eta$ is the fraction by which the low-level loop gain exceeds the gain required to initiate oscillation, and T is the time constant of the control system (thermistor or similar). In the case of a Wien-bridge oscillator, $\eta=3$.

This equation indicates that if the feedback amplitude is very little above that required to sustain oscillation – which is implicit in the design – the residual distortion will be dependent on the time constant of the control mechanism. By the use of series and parallel resistors of appropriate values with the thermistor, this can be made to control the amplitude of the oscillation at a resistance value which is only a little less than its room-temperature value. Under these circumstances, the settling time of the amplitude is long – perhaps 3-4 seconds at 1kHz, but the t.h.d. will be very low. The penalty incurred in this type of adjustment, apart from the obvious inconvenience of a relatively long settling time following any disturbance, is that the two gangs of the potentiometer used to control the operating frequency of the oscillator must be reasonably well matched in resistance value across the adjustment range, and also, if switched capacitors are used to provide step changes in frequency sweep, the ratios of their values must remain the same.

However, this is merely a statement of the obvious, that it is a pointless exercise to try to design high-performance equipment using low-performance components. Nevertheless, within the limitations imposed by the use of a thermistor as the stabilizing element, the performance of a very simple oscillator, built around a dual operational amplifier (a Texas Instruments TL072), is very good, as is shown in Fig. 6. The total harmonic distortion from this arrangment, in which the resistors associated with the thermistor were adjusted to give a settling time of 5 seconds at 1 kHz, and an output voltage of 2 volts r.m.s., is lower than that obtainable from any other simple Wien-bridge oscillator (that is to say with the exception of systems with low-pass output filtering) known to the author. This distortion is almost exclusively third harmonic – decreasing with frequency – which implies that the source of this waveform distortion is the instantaneous change in gain of the system, during the excursion of each half sinusoid, due to the limited thermal inertia of the thermistor.

The very high performance obtainable from such a circuit encourages the consideration of alternative methods of amplitude control such as that employing a photo-conductive cell and the light-emitting diode combination shown in Fig. 7, in which the time constant and other dynamic characteristics of the control circuit can be optimized by a suitable combination of proportional, integral and differential (p.i.d.) adjustment to the gain of the control circuit (A_2). Needless to say, the photoresistive element should be chosen to have a very low voltage coefficient of resistance and an adequate response speed to avoid the introduction of a further significant time delay into the control loop.

Leaving aside the question of the means employed to control the amplitude of the output signal (which imposes limitations of an identical kind on any oscillator system, in terms of the settling time, and the influence of the control time constant on the harmonic distortion at any given frequency) the improvement in performance given by the circuit design shown in Fig. 5 over that obtainable from the more conventional arrangement shown in Fig. 1, suggests that it would be sensible to regard the improved circuit as a general replacement for the earlier system in all future designs.

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The impact of new technology at work

More than £90,000 is being made available over three years by two research councils for studies into the shop-floor impact of the introduction of new technology.

Three studies are planned: on telephone exchange modernisation; on the adoption of a computer-based freight information system in British Rail; and on the introduction of electronic news gathering (ENG) equipment in television.

The Science and Social Science Research Councils are sponsoring the work which will be carried out by the New Technology Research Group of Southampton University. The Group has been formed by engineers and social scientists committed to interdisciplinary research "on the introduction of new electronic and computer technologies at the level of the individual workplace".

The two main objectives of the work will be to explore the process of technological change and to develop interdisciplinary research methods for the problems that arise. The team will be investigating the nature of technological innovation and engineering decision-making in the economic and social context of business organisations; the bearing of organisational structures on the capacity of managers to generate methods and mechanisms for the introduction and control of new technology; the development of union strategies towards new technology; the consequences of technological change for the nature of work and occupations; and the effectiveness of industrial relations procedures in handling new technology issues.

Measuring transient intermodulation

limiting filter.

The authors are interested in measuring t.i.m. principally to test the effectiveness of anti-t.i.m. measures such as input filters, and to design low-t.i.m. monolithic amplifiers. The availability of a simple and accurate measuring system has already provided useful results, exemplified by the R.I.A.A. preamplifier shown in Fig. 12; a circuit designed around the TDA3210 stereo preamplifier i.c. The filter on the output is intended to minimize t.i.m. in the next stage. This circuit, in terms of traditional parameters, represents the current state-of-the-art in i.c. R.I.A.A. preamps in which the total harmonic distortion is 0.02% at 20kHz. The frequency response is 20Hz to 20kHz ±0.5dB and the dynamic range 100dB.

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