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A technique for displaying the current and voltage output capability of amplifiers and relating this to the demands of loudspeakers

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A TECHNIQUE FOR DISPLAYING THE CURRENT AND VOLTAGE OUTPUT CAPABILITY OF AMPLIFIERS AND RELATING THIS TO THE DEMANDS OF LOUDSPEAKERS

by

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Abstract

The amplifier under test is fed with a large low-frequency sine-wave input on which 50μs pulses of alternate polarity are superimposed. These pulses drive the amplifier's protection circuits into current-limiting, and the set-up gives a CRT presentation having current-limit values vertically and instantaneous output voltage horizontally. The displays obtained, which sometimes disclose unsuspected amplifier design faults, are compared with those derived in tests made on loudspeakers using programme and other inputs.

0 INTRODUCTION

Transistor amplifiers usually incorporate protective circuit arrangements to limit the maximum peak instantaneous current that can be turned on, and the magnitude of this maximum current is often made to depend on the instantaneous output voltage.

The presence of such protective circuitry frequently makes an amplifier unable to produce as large a signal-voltage swing across a complex, i.e. partially-reactive, load impedance, as it can across a resistive load, and this sometimes leads to a disappointing performance when feeding typical loudspeakers.

The characteristics of amplifier protective circuits are not normally included in the specifications, and the purpose of the test set-up described below is to enable such characteristics to be readily examined.

It has been found that the use of the present testing method sometimes discloses misbehaviour of protective circuitry, in the form of over-protection, asymmetrical action, or RF parasitic oscillation, and it is therefore evident that application of the technique during the design and development of amplifiers would be advantageous in enabling such misbehaviour to be detected and eliminated.

1 THE BASIC TECHNIQUE

The essence of the scheme is shown in Fig. 1, and its functioning will now be described.

A low-frequency sine-wave input, typically at about 20Hz, is applied to the amplifier under test, its amplitude being sufficient to cause the amplifier output voltage to swing between voltage-clipping limits. Because the reactance of the 40μF capacitor on the output is about 200Ω at 20Hz, the amplifier does not give much low-frequency output current.

Superimposed on the low-frequency input is a waveform consisting of alternate positive and negative 50μs pulses with a fundamental frequency of normally about 500Hz. For the pulse component of the amplifier output, the 40μF capacitor has a very low effective impedance, so that large pulse output currents are produced. The amplitude of the input pulses is made sufficient to take the amplifier alternately into its positive and negative current-overload limits.

By feeding the amplifier output voltage to the X-plates of an oscilloscope, and the voltage across the 1Ω resistor (representing output current) to the Y-plates, the characteristic of the current-limiter circuit within the amplifier is displayed.

The waveform-generating circuit in Fig. 1 also produces bright-up pulses for feeding to the Z-modulation input of the oscilloscope. These pulses are preferably made of duration somewhat less than that of the 50μs pulses fed to the amplifier, the leading edge, of adjustable timing, being delayed by typically 20μs after the leading edge of the amplifier-input pulse. This ensures that the amplifier output current has time to reach its limiting value before the trace is brightened up.

If no bright-up facility is provided, a very messy and halated display is obtained, an example being shown in Fig. 2(a). With 50μs non-delayed-start bright-up pulses, the display of Fig. 2(b) is produced, delaying the start improving the picture to that shown at (c). A more striking improvement is sometimes obtained, depending on the nature of the amplifier being tested.

Though the figures of 20Hz, 500Hz and 50μs adopted in the above description will usually represent a suitable practical choice, they should not, of course, be regarded as inviolable. Also, as explained in section 3.1, there is a minor advantage in using a lower value of current-monitoring resistor than 1Ω.

2 THE WAVEFORM-GENERATING CIRCUIT

The circuit for producing the required pulse waveforms is shown in Fig. 3, this circuit being incorporated into the system as shown in Fig. 4.

The potentiometer in Fig. 4, in combination with the oscilloscope X-gain adjustment if provided, is set to give an appropriate scaling, such as 1 cm for 20V instantaneous output from the amplifier.

Some oscilloscopes give a spot movement to the left for a positive input to the X-terminal, and it may then be considered worthwhile to insert a phase-inverting op. amp. stage after the potentiometer, thus ensuring that the display is on the normal conventional sign basis. This was done for the photographs reproduced here, a Telequipment S54A oscilloscope being used.

Referring now to the detailed circuit of Fig. 3, input A is a 500Hz square-wave, supplied, in the author's set-up, by a Levell Type TG200DM oscillator. The amplitude is uncritical, and can be anywhere within the range 4V to 14V peak-to-peak. This square-wave is differentiated by C_1R_1, whose time-constant is approximately 1μs.

For most of the time, the non-inverting input of op. amp. 1 is at about +0.6V, owing to current flowing down R_2 into the diode D_1, and this causes the op. amp. output voltage to be at its positive overload-limit value of about +16V.

The occurrence of a positive-going pip at the inverting input then initiates a negative-going change at the op. amp. output, positive feedback via R_3 ensuring that the output reaches its negative overload limit of approximately -16V. The reason for including the resistor R_3 in the feedback path is that it prevents the magnitude of the negative voltage applied to the non-inverting input from exceeding the input-voltage rating of the op. amp..

The op. amp. output remains at this negative overload voltage until C_2/C_3 have had time to charge up sufficiently to bring the non-inverting input terminal back up to approximately earth level. Regenerative action again occurs, causing the circuit to revert rapidly to the state initially assumed. The value of $C_2 + C_3$ has been chosen to give a 50μs negative-pulse duration.

Op. amp. 2 operates similarly as a monostable circuit, but generates positive-going 50μs pulses, starting coincidently with the negative-going pips on the inverting input.

The positive and negative 50μs pulses are combined via R_6 and R_7 and controlled in amplitude by P_1. The purpose of the follower, op. amp. 3, is to ensure that the output impedance at socket B is constant, independently of the setting of P_1. Reference to Fig. 4 shows that variation in this impedance would affect the amplitude of the 20Hz component fed to the amplifier under test, such an interaction of controls being inconvenient and best avoided.

The remainder of the circuit is concerned with generating the bright-up pulses.

Op. amp. 4 inverts the 500Hz positive-going 50μs pulses from op. amp. 2, so that negative-going pulses occur on the points P and Q, interleaved in timing. These pulse waveforms are added by means of R_9 and R_{10}. If no other connection were made to the point S where these resistors join, the negative-going pulses there, at a PRF of 1000Hz, would have an excursion from approximately +16V down to earth.

R_{11}, however, shifts the mean level downwards and gives some attenuation in amplitude. With D_3 open-circuited, the pulse waveform at S would have an excursion from +3.1V to -6.8V, the effective source impedance being 3.1kΩ.

When D_3 is present, the upper level of the pulse waveform at S is held at approximately +1.2V by conduction of D_3 and D_4, the inverting input of op. amp. 5 being at about +0.6V. The output of this op. amp. is then at its negative overload limit of approximately -16V, no matter what the setting of P_2 may be.

Under these conditions, the top terminal of C_6 is at about -0.8V, as determined by R_{12} and R_{13}.

When a negative pulse occurs at S, D_3 and D_4 cease to conduct, and the voltage at the inverting input of the op. amp. becomes equal to that of C_6, i.e. approximately -0.8V. If the slider of P_2 is set at the top of the track, this -0.8V on the inverting input is sufficient to produce a pulse excursion from about +16V down to -16V at the op. amp. output, virtually coincident in timing, therefore, with the beginning of the 50μs pulse being fed to the amplifier under test.

If, now, the slider of P_2 is assumed to be set well down the track, the initial excursion to -0.8V at the inverting input is not enough to cause the op. amp. output to swing over. However, C_6 is meanwhile being charged negatively via R_{13}, and the voltage at the inverting input of the op. amp. soon descends to the voltage level at P_2 slider, whereupon the op. amp. produces a positive-going output transition, delayed in timing by an amount Δ dependent on the setting of P_2.

The supply voltages to the circuit have been made as high as the op. amp. ratings permit, to ensure comfortably adequate output amplitudes, particularly that for Z-modulation. Lower supply voltages may be used if found suitable, though a reduction to ±9V gives about a 10% shortening of the pulse duration, because the op. amp. slew rates are then less significant.

The two supplies should be kept well balanced for a predictable performance, because a 10% inequality in these voltages causes considerably more than 10% inequality in the durations of the positive and negative pulses.

Fig. 5 shows some actual waveforms produced by the Fig. 3 circuit. The input square-wave frequency was temporarily increased from 500Hz to 2kHz, to enable the pulses to be more clearly displayed. P_2 was set to delay the start of the Z-modulation pulses by about 20μs.

3 SOME FURTHER POINTS

3.1 The Display Mechanism in More Detail

The adoption of a lower value than 1Ω for the current-monitoring resistor has the advantage that certain small parts of the protection circuit characteristic are then not omitted from the display. The explanation of this point is as follows.

Suppose that the low-frequency input to the amplifier under test has swung its output voltage out to the point A in Fig. 6. The small low-frequency current component taken by the 40μF capacitor will be ignored for present purposes. A 50μs pulse then occurs, say of such polarity as to produce a positive-going change in the amplifier output voltage. If the amplifier is assumed, for the moment, to have a very rapid response, the working point jumps almost instantaneously to B, at which the current-limiter operates. The slope of the line AB corresponds to the 1Ω value of the current-monitoring resistor. After B has been reached, the output current continues to flow, charging the 40μF capacitor and causing the working-point to move to C by the time the 50μs pulse ends.

In practice the rate of rise of amplifier output current is finite, so that the approach from A to the current-limiting condition is along a curve, the 40μF capacitor charging considerably during this approach. Thus current-limiting commences at a point appreciably to the right of B. However, if P_2 (Fig. 3) has been set to delay the start of the bright-up pulse by an appropriate amount, this relatively slow transition from A to the current-limiting state will not be displayed, bright-up occurring at a point between B and C.

At C the 50μs pulse ends, but the state of charge of the 40μF capacitor remains markedly different to what it would have been if the 50μs pulse had not occurred. The result is that substantial reverse current flows back into the amplifier output for a short time. However, since black-out occurs at C, this reverse current flow is not displayed.

Of more concern is what happens later, when the next 50μs pulse arrives, this time producing a negative-going change in the amplifier output voltage. The 20Hz input to the amplifier has meanwhile caused a slight change in output voltage, say to the point D. Again assuming a very rapid amplifier response, the occurrence of the pulse shifts the working point to E, DE having a slope of 1Ω. Alteration in the state of charge of the 40μF capacitor shifts the working point to F by the end of the pulse.

With a sufficiently large positive output voltage due to the 20Hz input, the point G at the corner of the complete V-I characteristic will be reached before the end of the 50μs pulse has occurred. Output current then collapses, since to maintain it would require the capacitor voltage, and hence the amplifier output voltage, to continue rising at a rate given by $dV/dt = I/C$. Voltage clipping in the amplifier prevents this further rise. The fall of current along the line GH is displayed, for black-out has not yet occurred.

Whereas, as just explained, the corner G of the V-I characteristic can be included in the display, the lower corner J cannot be. The nearest approach is obtained by starting from H, which would enable the point K to be displayed if the amplifier had a sufficiently rapid response. Since the slope of HK is 1Ω, it is clear that, under these conditions, reducing the value of the current-monitoring resistor to a much smaller figure would enable the point J to be much more nearly included in the display. In practice, however, the finite slew-rate of the amplifier would prevent this improvement from being fully realised.

There is a possible risk that some amplifiers may not retain adequate feedback stability when supplying a load consisting of a large capacitor in series with, for example, only 0.1Ω, and this favours the adoption of a higher value, such as 1Ω, for general use. A better compromise might, perhaps, be 0.5Ω.

3.2 Purity of the Current-Monitoring Resistor

In the earlier stages of this work, before the facility for delaying the start of the bright-up pulses was incorporated, it was found to be important to keep the amount of stray inductance in series with the current-monitoring resistor adequately small, especially if resistor values of less than 1Ω were used. Such series inductance can result in the display overshooting the proper boundary of the V-I characteristic when a fast-responding amplifier is being tested, though usually for less than 1μs. Quite a small amount of bright-up delay prevents this effect from being seen, rendering the use of ordinary wire-wound resistors perfectly satisfactory, however.

3.3 Rating of the Current-Monitoring Resistor

The amount of power dissipated in a 1Ω current-monitoring resistor due to the low-frequency output component from the amplifier is quite small provided the frequency is kept low enough. At 20Hz, with an amplifier operating on ±50V supplies, the dissipation will be less than 0.1W, even allowing for the fact that the 20Hz input voltage is made large enough to give voltage clipping and hence a non-sinusoidal 20Hz output waveform.

However, care should be taken during setting-up adjustments not to let the sine-wave oscillator operate inadvertently at a high frequency, since this could easily result in burning out the 1Ω resistor.

Assuming, by way of example, that the amplifier under test has a constant-current limiter characteristic limiting at 10A, the power dissipated in a 1Ω current-monitoring resistor during a 50μs pulse is 100W. If the square-wave frequency is 500Hz, there are 1000 pulses per second, i.e. one 50μs pulse every millisecond. The mean power dissipated in the 1Ω resistor is then 5W. A resistor, or series/parallel combination of resistors, rated at, say, 10W should therefore be comfortably adequate in most circumstances, permitting peak currents up to at least 14A.

3.4 Standardised Display Scalings

To facilitate easy comparison of the V-I limiter displays for various amplifiers, it is convenient to adopt standardised scalings for current and voltage, and it is suggested that an appropriate choice will normally be 5A/cm and 20V/cm.

This also aids comparison of amplifier and loudspeaker V-I displays, as described in section 8.

4 OPERATING PROCEDURE AND PHOTOGRAPHY

(a) To set up for the standardised X-scaling of 20V/cm, apply only the sine-wave input to the amplifier under test, at 20Hz or other fairly low frequency - beware making the frequency more than about 200Hz, as this may burn out the current-monitoring resistor and/or cause excessive dissipation in the amplifier output transistors. Put a voltmeter across the amplifier output and adjust the input for a reading of 28.3V rms. Then adjust P (Fig. 4) and/or the CRO X-gain for ±2cm deflection.

(b) Set the CRO Y-sensitivity to 5V/cm to obtain the standardised Y-scaling of 5A/cm - a 1Ω current-monitoring resistor is here assumed.

(c) With zero signal input, carefully set the spot to the centre of the graticule. Now apply 20Hz or other low-frequency sine-wave input at comfortably sufficient level to make the amplifier overload, i.e. give voltage clipping. The potentiometer P_1 in Fig. 3 must be at zero, so that there is no pulse input to the amplifier. Set the CRO brilliance control for a not-too-bright display as shown in Fig. 7.

In general the brightened-up small dots will be seen to be moving round the ellipse, unless the 20Hz and 500Hz are in precise integral relationship. For good photographs, it is best to set the frequencies so that the dots drift round quite slowly and then give a fairly long exposure, e.g. 5 seconds, for the subsequent V-I display photographs.

(d) Now turn P_1 (Fig. 3) up to maximum, or at least up to a sufficiently high setting to produce hard current-limiting as indicated by the display. There is something to be said for not keeping P_1 turned up for more than about 10 seconds, though it is probable that most amplifiers would actually withstand continuous stressing in this manner.

(e) Adjust the bright-up-delay control, P_2, for the best picture, and make the photographic exposure.

5 8Ω AND/OR 4Ω LINES

Lines representing 8Ω and/or 4Ω resistive loads may be introduced into the photographed displays if desired. This may be done by proceeding as follows, a separate exposure being made.

(a) Set P_1 to zero (no pulse input) and disconnect the CRO Z-modulation.

(b) Assuming the current-monitoring resistor to be of value 1Ω, replace the 40μF capacitor by an accurate 7Ω resistor of adequate power rating, for the 8Ω line, or by a 3Ω resistor for the 4Ω line, and turn up the sine-wave input for just long enough to take a photograph with, say, a 1-second exposure. (The 1Ω resistor will probably be dissipating much more than its normal rating during this burst of output.)

With the standardised X and Y sensitivities as described in section 3.4, the 8Ω line should be at "1cm up for 2cm across" slope, the 4Ω line being at 45° slope.

6 ADVANTAGES OF A CURRENT-MONITORING TRANSFORMER

There are advantages to be gained, in certain respects, by inserting a step-up transformer between the circuit whose current is to be monitored, and the monitoring resistor itself.

A 1Ω monitoring resistor used straightforwardly gives a sensitivity of 1V/A, inserts 1Ω in the monitored circuit, and dissipates 100W for a current of 10A. Using, for example, a 1 : 1000 transformer with a 1kΩ resistor across the secondary, this same sensitivity of 1V/A is obtained, but the effective resistance inserted in the monitored circuit is now, ideally, only 1 milliohm, and a current of 10A now dissipates only 100mW in the resistor.

The use of a transformer would therefore avoid the dangers, mentioned in sections 3.3 and 5, of accidentally burning out the current-monitoring resistor.

The point concerning amplifier stability, mentioned in the last paragraph of section 3.1, should be borne in mind when a transformer is employed, and this may necessitate the addition of a small amount of resistance in series with the primary when obtaining the V-I limiter displays of some amplifiers. The added resistor need not be of accurately known value, however, for it does not affect the display except in the minor respect explained in the penultimate paragraph of section 3.1.

The above-mentioned added resistor would not be required when displaying 8Ω and/or 4Ω lines, and the use of a transformer avoids the need for employing load resistor values of less than 8Ω and 4Ω as described in section 5.

The negligible effective series resistance introduced when a transformer is employed is also advantageous when determining the V-I displays for loudspeakers - see section 8.

The advantages of a transformer may be obtained, though not fully, merely by employing a very low value of current-monitoring resistor with appropriately-increased CRO Y-gain. It becomes increasingly difficult, however, as the resistance value is reduced, to avoid significant effects due to series inductance, and the connection of several resistors in parallel with very short leads is desirable. The measurement of the resistance value with good accuracy is also more difficult at very low values, and both these difficulties are avoided when a transformer is employed.

A transformer was in fact used for most of the limiter-characteristic, load-line, and loudspeaker displays shown in this paper, a 1Ω resistor being included in series with the primary for the limiter displays only, in the interests of good stability.

The transformer has a 1000-turn secondary of 0.12mm dia. enamelled wire (40 SWG, 36 AWG) wound in a single section on a core of 0.38mm (0.015 ins.) Mumetal (Permalloy C) laminations of maximum dimension 25.4mm (1 ins.) and stack 12.7mm (0.5 ins.). Centre-limb width 6.3mm (0.25 ins.) - No. 187 laminations. The primary is a single turn round the outside of the secondary.

A 100Ω ±1% metal film 0.125W resistor is connected across the secondary, giving a sensitivity of 100mV/A. A 2.2nF shunt capacitor is also added, to suppress a slight tendency to ring at about 1MHz, involving distributed winding capacitance and leakage inductance. The 100mV/A sensitivity is less than that given in the above example, but the use of a 100Ω rather than 1kΩ secondary resistor gives a much better very-low-frequency response, which is less than 3dB down at 1Hz, the phase error being less than 2^{o} at 20Hz. This phase error is sufficient, however, to give a noticeable opening-out of the 8Ω and/or 4Ω lines (see section 5) if the oscillator frequency is made as low as 20Hz, and about 70Hz was therefore used.

The above transformer had actually been made for other measurement purposes, and was therefore conveniently available. Had this not been the case, it must be confessed that straightforward current-monitoring resistors would probably have been used for all the measurements!

7 SOME EXAMPLES OF DISPLAYS OBTAINED

Fig. 8(a) relates to an early version of the Quad 405 amplifier, and is the same as in Fig. 2(c) but with the addition of 8Ω and 4Ω load lines. A point of interest is that the excursion with the 4Ω load goes outside the limiter characteristic. The explanation of this involves the fact that the unregulated positive and negative DC supply voltages fall considerably when the amplifier is feeding a 4Ω load at a high signal level. The nature of the voltage-dependent current limiter circuit is such that a fall in supply voltage increases the current value, for a given instantaneous amplifier output voltage, at which current limitation occurs. A decrease in mains voltage also increases the peak current that can be turned on into a 4Ω load.

With an 8Ω load, on the other hand, overload occurs because of voltage clipping rather than current limitation, a fall in DC supply voltage then causing a reduction in the peak output level achievable.

Later versions of the Quad 405, with serial numbers above 29,000, had resistors R_{27} and R_{29} in the limiter circuit increased from 8.2kΩ to 15kΩ, and Fig. 8(b) shows the display obtained with this modification incorporated.

Fig. 8(c) is for a Quad 303 amplifier, which has a stabilised DC supply and a simple form of non-voltage-dependent current limiter. The fact that this amplifier has a smaller output rating than the 405 is instantly conveyed by the smaller size of the display, emphasising the advantage of adopting a standardised scaling for the photographs.

In contrast, Fig. 8(d) relates to a pre-production version of the Quad 520 rack-mounting professional stereo amplifier, which has parallelled transistors of high rating in the output stage.

The four photographs in Fig. 9 illustrate some defective performance features, and were obtained with a rack-mounting amplifier made by another British firm.

Fig. 9(a) was photographed with the amplifier in its original state, as supplied. Investigation of the circuit disclosed that the current-limiter clamp transistor, on the positive side of the circuit, was called upon to clamp an indefinitely large current turned on by the previous transistor, this current being restricted only by the current-gain factor of the transistor. It was thus a matter of which transistor would win in a rather brutal contest! Attempts had evidently been made to rescue the situation, including fitting a sizeable heat sink to the amplifying transistor and a 470Ω resistor in series with its output - the latter, however, being singularly ineffective under the conditions applying in the top left-hand quadrant.

By inserting a 10Ω resistor in the emitter lead of the PNP stage whose output is clamped, and a series pair of small silicon diodes to limit the base voltage applied to this transistor, the clamping operation was made predictable and relatively gentle, giving the display shown at (b). Unenlightened design features such as the above are only too prevalent in much equipment whose appearance and advertising might lead one to expect the highest standards.

The (b) display is, no doubt, as intended by the designer, and it shows that the amplifier comfortably meets its specification of 70W into 8Ω and 100W into 4Ω, these being mean-power figures for continuous sine-wave operation with a resistive load. However, the voltage dependency of the current-limiter has been made so drastic that the full output-voltage level cannot be obtained without serious distortion, even under sine-wave conditions, when there is a substantial amount of reactance in series with the load resistance, particularly when the latter is well under 8Ω.

Analysis shows that if a straight line is drawn from the right-hand tip of, say, the 4Ω load-line, down to intersect the X-axis at an equal negative output voltage, and if any part of this line goes outside the limiter characteristic, then a sine-wave output voltage equal to that represented by the load-line will not be obtainable without distortion when certain values of series reactance are present. Conversely, if the above line lies everywhere below the limiter characteristic, the full output level will be obtainable with negligible distortion for any value of series reactance. Similar considerations apply, of course, to the lower half of the display.(1) See also section 8 and Fig. 17.

Fig. 9(c) shows that RF oscillation occurs when the limiter operates, if a 47nF capacitor in the positive side of the limiter circuit is removed - thus relieving the author's curiosity as to why this capacitor had been included! In another amplifier, of continental origin, however, oscillation of a similar kind occurred with the amplifier in an unmodified state.

For Fig. 9(d), the value of the bootstrapping capacitor joining the amplifier output to the junction of the two resistors that constitute the collector load of the pre-driver PNP common-emitter amplifying stage, was reduced from 47μF to approximately 8μF. The low-frequency input was at 20Hz, as for the other Fig. 9 photographs. Near the lower left-hand tip of the display, the slanting line just above the tip itself is traversed with the spot moving left to right, the previous right-to-left spot movement being via the actual tip. What happens is that the voltage across the bootstrapping capacitor, and hence the voltage across the upper of the two collector load resistors mentioned above, falls off while the amplifier output voltage is near its maximum negative excursion, ultimately becoming insufficient to provide the required base current to the lower driver transistor. The output transistor is then unable to turn on the full output current desired.

Increasing the low-frequency input from 20Hz to 50Hz or above eliminated the above effect from the display. With some amplifiers the effect may be observed, at least mildly, at 20Hz, even without reducing the value of the bootstrapping capacitor fitted.

It will be noticed that the right and left-hand boundaries of the Fig. 9 displays are much less steep than most of the previous ones. This is because the amplifier in question has 0.5Ω emitter resistors in the complementary output stage, whereas the Quad amplifiers of Figs. 2 and 8, except for the 303, have very much lower resistor values. This effect is ignored in section 3.1.

8 PROTECTION CIRCUITS WITH TIME-DEPENDENT BEHAVIOUR

Some protection circuits, though not those considered above, incorporate arrangements whereby the current limitation becomes more severe than normal if the operating point remains for more than a short time in "hazardous parts" of the V-I display area, or if the excursions to these places, though of short duration, are very frequent.

The Quad 405/2 amplifier is fitted with protection-circuit modules having the features just mentioned.*

The essence of the protection circuit on the positive side of the amplifier is conveyed by the highly-simplified diagram of Fig. 10. The device shown as a simple diode is actually a two-transistor circuit, rendered functional only when the voltage across TR_2 exceeds about 45V. Thus for the conditions existing in the top left-hand display quadrant, the circuit shown is the relevant one.

For large short-duration current pulses occurring in TR_2 only occasionally, C remains virtually uncharged, and the protective clamp transistor TR_1 is not brought on until about 1.5V drop

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* In some Quad diagrams the extreme left-hand of the four transistors in each module is inadvertently shown as PNP, whereas it is actually NPN.

occurs across the 0.18Ω resistor, corresponding to about 8A. The time-constant of 68Ω and 47μF is 3.2ms, so that for isolated current pulses approaching this duration or longer, the capacitor is charged to a significant voltage. When this has occurred, a smaller voltage drop across the 0.18Ω resistor is sufficient to bring on the clamp transistor. Therefore, for an isolated current pulse of long duration, the maximum current that the amplifier can turn on is more severely limited during the later parts of the pulse than at the beginning.

Though, as already mentioned, an isolated short pulse does not significantly charge C, a rapid succession of short pulses will do so, again causing the maximum current that can be turned on to be reduced.

This more subtle type of protection circuit takes more fully into account the transistor manufacturer's V and I rating limitations as expressed in the form of SOAR curves and as extended beyond these to allow for frequent operation with current pulses of large magnitude.(2),(3) The circuit does not, however, introduce the type of instantaneous negative output resistance inherent in more normal types of protection circuit, the latter sometimes producing rather unpleasant sounds, with some types of load, if overdriven. Only the minimum necessary protection is provided when short-duration musical transients occur.

The displays of Fig. 11(a) were obtained with a Quad 405/2, the low-frequency sine-wave input being at 70Hz rather than 20Hz to avoid a small amount of the effect described in section 7 in relation to Fig. 9(d). The PRF of the pulses was 100Hz for the dimmer display and 2kHz for the brighter one, the exposures being approximately equal.

The display of Fig. 11(b) was obtained with the 47μF capacitor in the negative-side protection circuit removed, the PRF being 500Hz. This demonstrates that the effective diode of Fig. 10, but in the negative-side protection circuit, becomes functional only when the amplifier output voltage is more positive than about -8V.

The pulse duration for the Fig. 11 tests was 50μs, as with the other displays. Much longer pulses may be used, but to avoid excessive mean power dissipation, the PRF should then be much reduced. A satisfactory technique is to make the PRF nearly equal to that of the low-frequency sine-wave, so that the pulses are phased with respect to the X-deflection in a slowly-varying manner, gradually tracing out the relevant limiter characteristic.

Another aspect of time-dependent behaviour is that it is sometimes observed that the limiting values of output current, as displayed, gradually change, over a period of minutes, after switching on the amplifier or the test signal, normally reducing in magnitude. This effect, usually fairly small, involves warming up of the protection-circuit transistors by heat conducted from hot components in the amplifier, often mainly along PCB conductors.

9 THE RELATIONSHIP BETWEEN AMPLIFIER AND LOUDSPEAKER V-I DISPLAYS

To be of the greatest value, the V-I display for an amplifier should somehow be related to the characteristics of the loudspeaker to be driven, and it should here be borne in mind that loudspeaker data obtained on a sine-wave basis may not be directly relevant to the practical problem of feeding the loudspeaker with music waveforms.

On light load, an amplifier normally clips at a fairly closely defined peak instantaneous output voltage, and the question to which an answer is usually wanted is whether the amplifier can feed a particular loudspeaker with a wide variety of programme voltage waveforms peaking up to about this same instantaneous clipping voltage without running out of current-turning-on capability. Or, if the amplifier cannot produce this full instantaneous voltage level, then it is desirable to know up to what level it may be allowed to go before current-limitation sets in.

The answers to such questions may be obtained by carrying out tests on the loudspeaker using the set-up of Fig. 12.

To make the results easy to compare with the amplifier displays described earlier in this paper, it is convenient to set the system up so that a 45° line is produced on the CRO if the loudspeaker is replaced by a 4Ω resistor.

On the assumption that the loudspeaker may be regarded as a linear device, there is no need to do the test at a very high volume level - the X and Y gains may be increased appropriately so that displays of similar size to the amplifier displays may nevertheless still be obtained.

A variety of loud music passages should be used for this test, and they should be chosen to contain both high peak voltages and high peak rates-of-change of voltage; in other words, there should be plenty of high-frequency as well as low-frequency content. An indication of the degree to which a given music excerpt is suitable in this respect may be obtained using a special full-wave peak programme meter designed by the author.

This PPM has the following features:

(a) Charging time-constant about one hundredth of that for a normal PPM.

(b) Every time a programme transient occurs that is of higher peak instantaneous magnitude than that already being indicated on the meter, the needle jumps up to indicate this new level and remains stationary at the new reading for one second before starting to fall back, thus facilitating easy and accurate reading of the programme peak. If an even higher programme peak happens to occur during this one-second interval, the needle jumps up further to indicate this new peak level, and a new one-second "dwell time" is initiated.

(c) The circuit may be switched to read either the peak values of the programme voltage V, or to read the peak values of dV/dt. The relative sensitivities are made such that equal readings are given in the two switch positions for a sine-wave input at 5kHz.

(d) An output voltage proportional to meter readings is made available for operating a chart recorder or an oscilloscope.

The instrument as it now exists is linear rather than logarithmic, though it can be modified for logarithmic operation.

Fig. 13 shows the results obtained using this meter on two music excerpts each lasting about 40 seconds.

Fig. 13(a) relates to the beginning of Johann Strauss's Radetzky March from Deutsche Grammophon CD No. 410 027-2, and involves loud and vigorous playing by the Berlin Philharmonic, with cymbal clashes - it is the latter that give the high values of dV/dt shown.

Fig. 13(b) relates to the beginning of track 6 on Denon CD No. 38C37-7043, Debussy Preludes from Book 2 played by Jacques Rouvier on a Steinway. This is loud and dynamic piano playing, but it is interesting to see that the ratio of dV/dt to V is fairly small compared with that for the orchestral excerpt.

Other digitally-recorded material giving a high ratio of dV/dt to V includes applause, a brass group, a bell, and fast drum rim-shots. The ratio in each case may be expressed in terms of f_o, the frequency of a sine wave that has the same ratio of peak values of these quantities.(4) For items such as the last two, f_o values as high as 10kHz can occasionally be obtained, f_o for Fig. 13(a) being about 3.5kHz if derived on the basis of the ratio of the highest dV/dt value during the excerpt to the highest V value.

It is evident that it is music with cymbal clashes, and items such as those mentioned in the last paragraph above, that are likely to lead to problems with loudspeakers whose impedance drops to low values at high frequencies.

The photographs of Fig. 14 were obtained with three different loudspeakers, at a fairly low signal level, using the Fig. 12 set-up, a suitable three-second part of the Fig. 13(a) orchestral excerpt being chosen. The gain was set to obtain a maximum V value on the PPM corresponding to 2 cm horizontal spot deflection.

Because the signal, even during loud passages, actually spends most of its time at relatively small instantaneous voltage values, photography is somewhat troubled by a halation problem, and it is difficult to obtain a permanent record of the peak spot excursions very clearly - they can be seen better by direct vision, especially in a darkened room with a highish CRO brilliance setting.

Attempts were made to improve the result by adding a circuit to brighten the trace during large amplitude or high-velocity spot movements, but this was only partially successful.

What is ideally wanted is a circuit arrangement that leaves a permanent record on the CRT of the most extreme spot displacements that have as yet been reached during a given musical passage - a bright and clear boundary line that keeps being pushed outwards every time a larger spot excursion occurs.

A circuit for doing approximately this could be made, but would be fairly elaborate and expensive. One would have a number of gate circuits, all fed in parallel from a signal voltage represent- the loudspeaker current, the output of each gate feeding a peak rectifier with a very long decay time. Each gate would be opened only when the instantaneous loudspeaker voltage lay within a certain small voltage interval. The magnitudes of the outputs of the various peak rectifiers would be sampled and displayed as Y-deflections at horizontal positions representing the relevant amplifier output voltages. Negative-responding as well as positive-responding peak rectifiers would be required. A display would thus be built up on a stepped basis, the steps being small and close together if a sufficiently large number of peak rectifiers etc. were used.

A much easier technique for obtaining V-I displays for loudspeakers is to use the Fig. 12 set-up, but with a swept sine-wave input.

A low brilliance setting should be used, the camera shutter being open throughout the sweep, which should be logarithmic, occupying several seconds at least. A logarithmic sweep avoids an excessive sweep-rate at low frequencies, allowing virtually steady-state current values to be established, and it gives equal photographic exposure for each octave, which is appropriate.

The photographs of Fig. 15 were obtained in the above manner, for the same loudspeakers as were used for Fig. 14.

Comparison of Figs. 14 and 15 does not tend to lend support to the notion that the peak currents demanded by loudspeakers fed with programme input are liable to exceed those for sine-wave voltage input of the same peak magnitude.(5) Indeed, rather on the contrary, the sine-wave sweeps seem to give larger peak current values.

A weakness of the sweep technique, in one sense, is that the height of the display obtained is enhanced just as much by low loudspeaker impedance at 20kHz as it is by a low impedance at frequencies which are more significant from a programme point of view. This point shows up particularly in Fig. 15(c) for the Quad ESL63, whose impedance modulus at 15kHz is about 3.5Ω only. By limiting the sweep to 40Hz to 5kHz, instead of 20Hz to 20kHz, the display at (d) is obtained, and is clearly more relevant to likely current demands under programme conditions.

There is no doubt, on the other hand, that certain artificial test-signal waveforms may be produced that will cause a loudspeaker to draw larger peak currents than with a sine-wave input of the same peak value. This effect is not confined to devices such as loudspeakers, where a motional EMF is involved, but is a property of many passive networks containing reactive elements. The simplest example is that of Fig. 16, where it is seen that increasing the magnitude of the load impedance by adding a series capacitor, doubles the peak current.

Another test on the KEF Corelli loudspeaker employed above showed that, whereas with constant-voltage sine-wave drive it took maximum current at about 8kHz, the maximum current with square-wave drive occurred at a fundamental frequency of about 4kHz, and was of approximately 35% greater peak instantaneous value, the drive voltages being of the same peak value in both cases.

However, no music waveform observed has approximated even roughly to a square-wave of full amplitude, the tendency being for waveforms of very large peak amplitude to be of a spiky nature.

A combination of a hefty low-frequency component with a short-duration impulsive spike superimposed on it can, with suitable phasing of the two, give an unusually large peak current for a given total peak voltage. But the probability is that, at other times in the music, such components will be differently phased, in such a manner as to give a larger peak voltage and hence voltage clipping. The programme level then needs to be turned down if overloading is to be avoided, and the previously excessive current demands now no longer arise.

Thus, though the topic could obviously be further investigated at length, the author is inclined to think that if an amplifier can provide peak currents in accordance with the requirements indicated by sweep tests of the Fig. 15 type, it will also cope adequately with any normal programme material having the same peak instantaneous voltage.

An alternative approach, which is attractive because it is so easy to apply, is based on a notion mentioned in section 7 and illustrated here in Fig. 17. No matter how complex may be the equivalent circuit representing the electrical impedance of a loudspeaker, this impedance is always purely resistive at, or very near to, the frequencies where it dips down to minimum values. At frequencies in the same broad region, either side of the minimum, the impedance remains fairly low but now has significant reactance in series with the same resistance value. In such regions at least, the impedance of a loudspeaker thus has the same nature as that of the simple R and X combination shown in Fig. 17(a), in which R remains constant and X varies with frequency. The V-I display for such a combination, with constant sine-wave input voltage of varying frequency, is therefore a series of ellipses, and it may be shown that these all fall within, and tangential to, the broken-line parallelogram of Fig. 17(b).

Hence, if the minimum impedance of a loudspeaker is known, as it usually is from the published impedance-modulus curve, a line representing this (resistive) impedance may be drawn on the V-I plot, a standardised scaling being adopted again, such that a 45° slope would represent 4Ω. The broken-line figure is then completed, starting at a point conveniently representing the peak voltage swing that the amplifier to be used is expected to be able to produce.

The actual V-I display for a loudspeaker, determined as for Fig. 15, will not normally occupy quite the full space inside the parallelogram obtained as in the previous paragraph, for full occupancy would require the series reactance to vary with frequency from zero up to an infinitely large value. Thus the parallelogram represents a worst case, so-to-speak, and if it fits within the protection-circuit display for a specific amplifier, when drawn to the same width, then that amplifier should comfortably be able to deliver its full output voltage to the loudspeaker without significant distortion. If the parallelogram will not fit within the amplifier display on this basis, then it should be scaled down until it just will - because of its simple shape, this is easily done. The reduced output level likely to be obtainable without significant distortion can then be seen.

Using the minimum-$|Z|$ values for the three loudspeakers tested above, and completing the parallelograms as described, gives the results shown in Fig. 18. These are not greatly different from the sweep results of Fig. 15, though the latter represent slightly easier amplifier-loading conditions.

The above parallelogram technique can occasionally give an unduly pessimistic prediction of the output level capability of a particular amplifier/loudspeaker combination. A good example of this would be a KEF 104/2 loudspeaker used with the amplifier to which Fig. 9(b) relates. This loudspeaker is unique in that, by the use of appropriate conjugate networks in the internal circuits, is has an almost purely resistive impedance of 4Ω throughout the audio spectrum. The amplifier can therefore produce about 32V peak across this loudspeaker, no matter what waveform may be involved, corresponding to a peak power of 256W. Since, with this loudspeaker, the current is zero when the voltage is zero, the rather severely restricted current capability of the amplifier at zero instantaneous output voltage is of no consequence. If, however, the above parallelogram technique were thoughtlessly applied in this case, a maximum peak output power of only about 64W, without significant distortion, would be predicted.

The rather awkward problems discussed above concerning the relationship between the peak instantaneous current demands of loudspeakers under sine-wave and programme conditions completely disappear, of course, ideally, when the conjugate-networks technique is used to give a purely resistive impedance.

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(2) Gates, T. W. and Ballard, M. F., "Safe Operating Area for Power Transistors", Mullard Technical Communications, Vol. 13 No. 122, pp. 42-65 (April 1974).

(3) Noble, P. G., "The Safe Operation of Power Transistors", Mullard Technical Communications, Vol. 14 No. 139, pp. 346-375 (July 1978).

(4) Baxandall, P. J., "Audio Power Amplifier Design - 1", Wireless World, Vol. 84 No. 1505, pp. 53-57 (Jan. 1978).

(5) Martikainen, I., Varla, A. and Otala, M., "Input Current Requirements of High-Quality Loudspeaker Systems", Preprint No. 1987 at AES 73rd. Convention, Eindhoven. (March 1983).

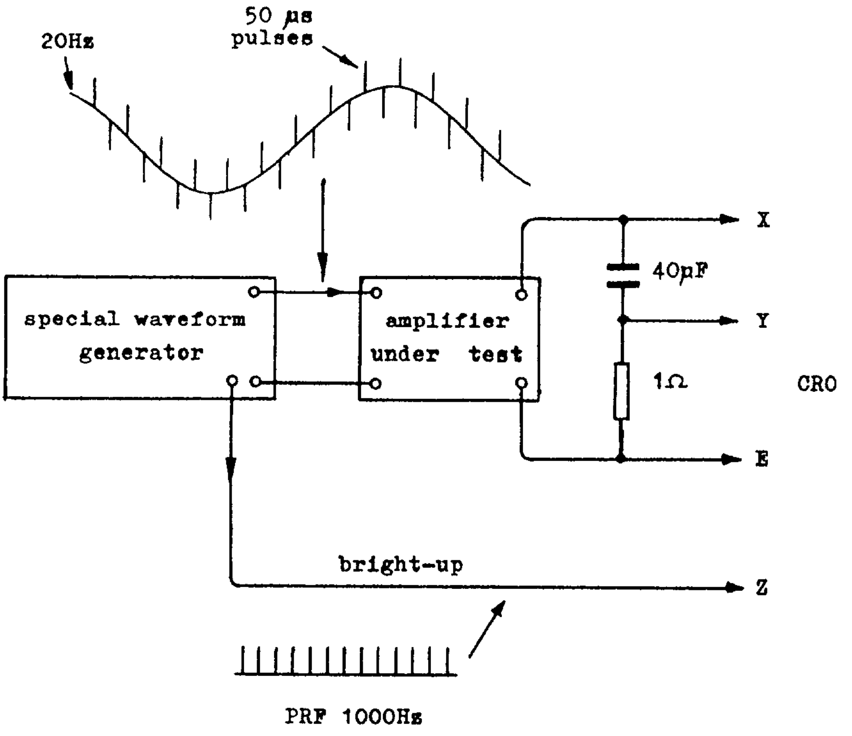


Fig. 1. The essence of the scheme.

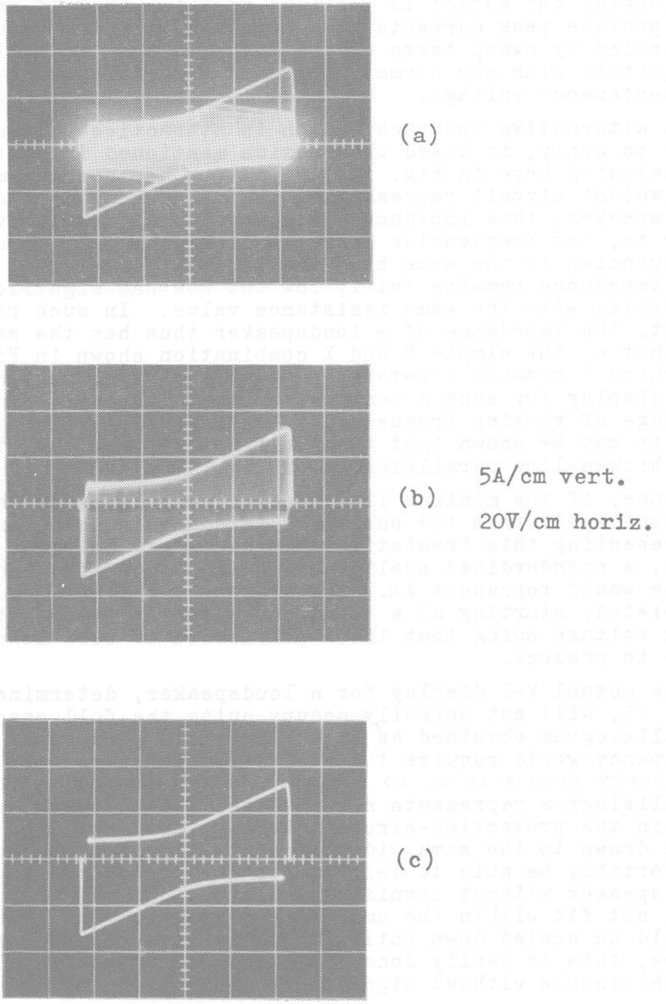


Fig. 2. V-I displays for early version of Quad 405 amplifier.

(a) With no bright-up facility.

(b) With 50μs non-delayed-start bright-up pulses.

(c) With start of bright-up pulses delayed by about 20μs.

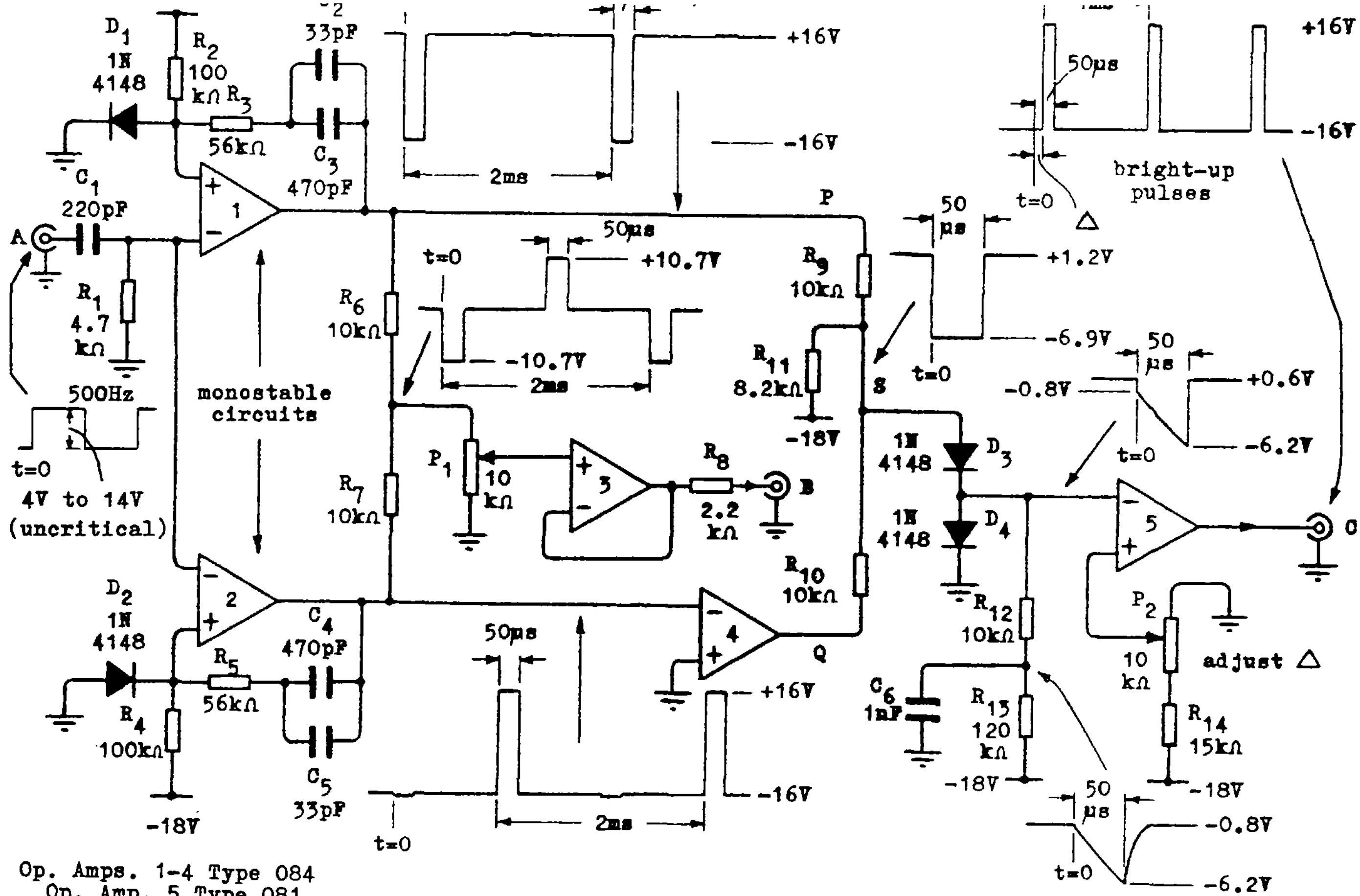
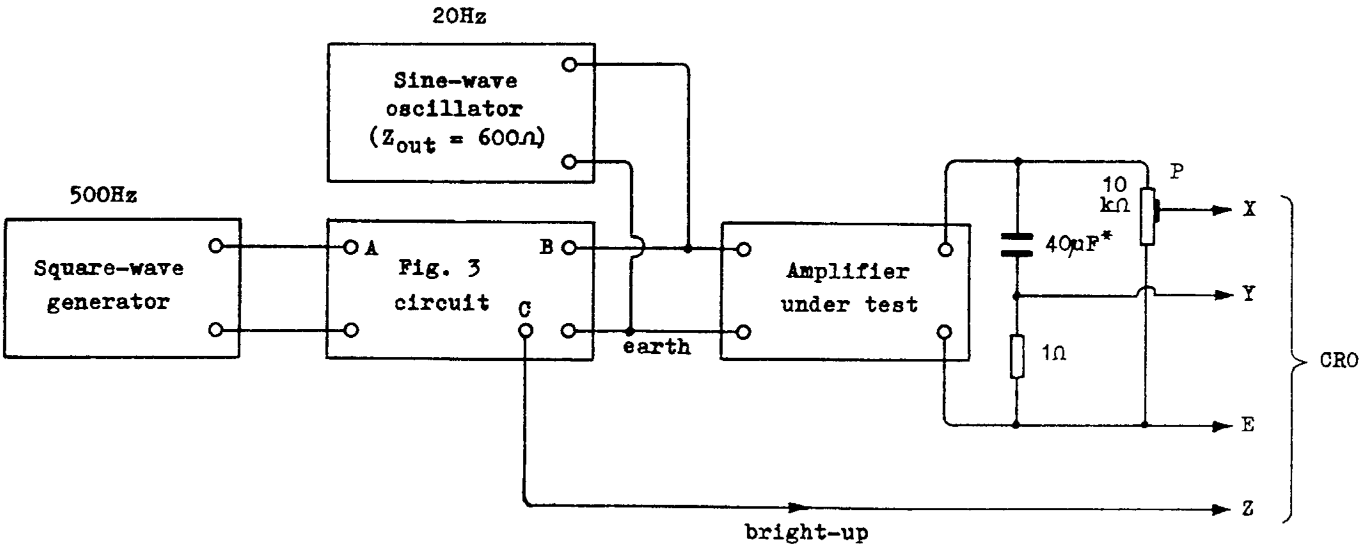


Fig. 3. The waveform-generating circuit.



* Four 10μF 63V dc polyester capacitors in parallel

Fig. 4. The complete set-up, incorporating the Fig. 3 circuit.

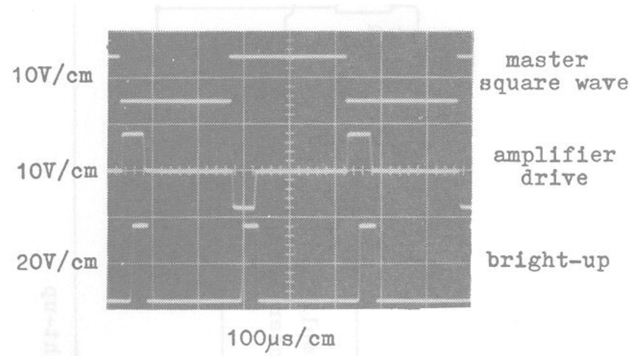


Fig. 5. Waveforms produced by the Fig. 3 circuit.
The square-wave frequency was increased
to 2kHz for greater clarity.

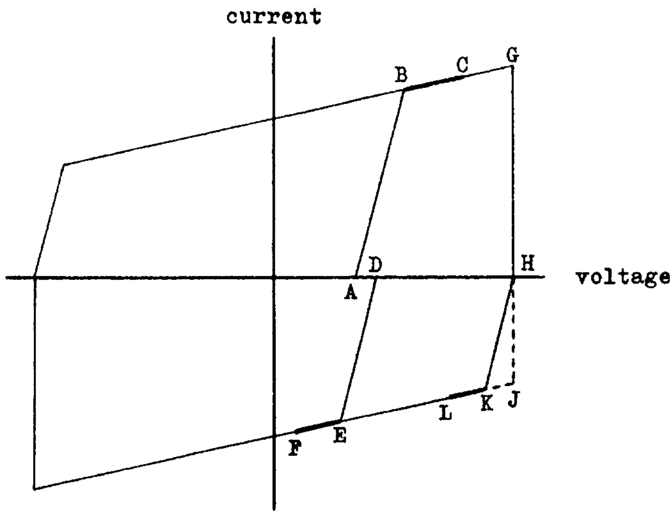


Fig. 6. Display details.

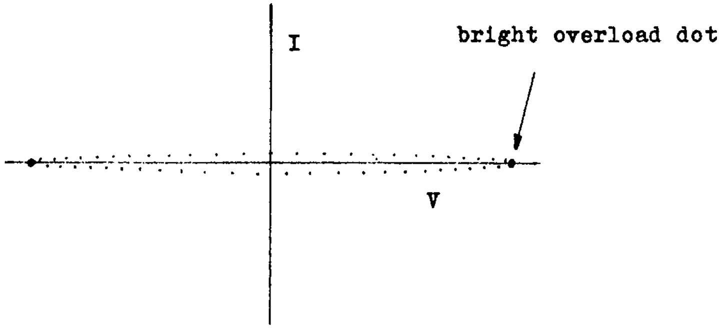
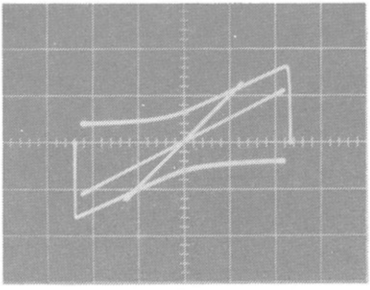
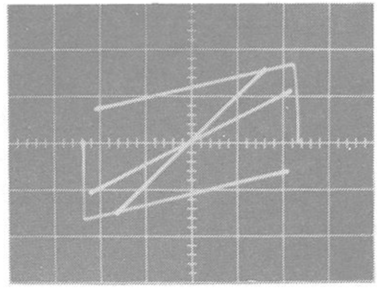


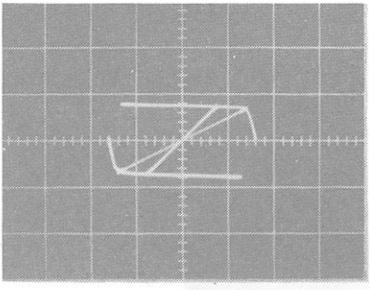
Fig. 7. Display obtained during setting-up adjustments.



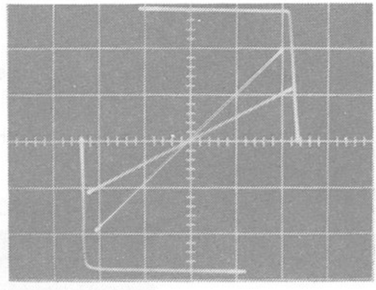
(a)



(b)



(c)



(d)

Fig. 8. V-I displays, including 8Ω and 4Ω load lines, for four amplifiers:

(a) Early Quad 405
(b) Later Quad 405
(c) Quad 303
(d) Pre-production Quad 520

Scales: 5A/cm vert., 20V/cm horiz.

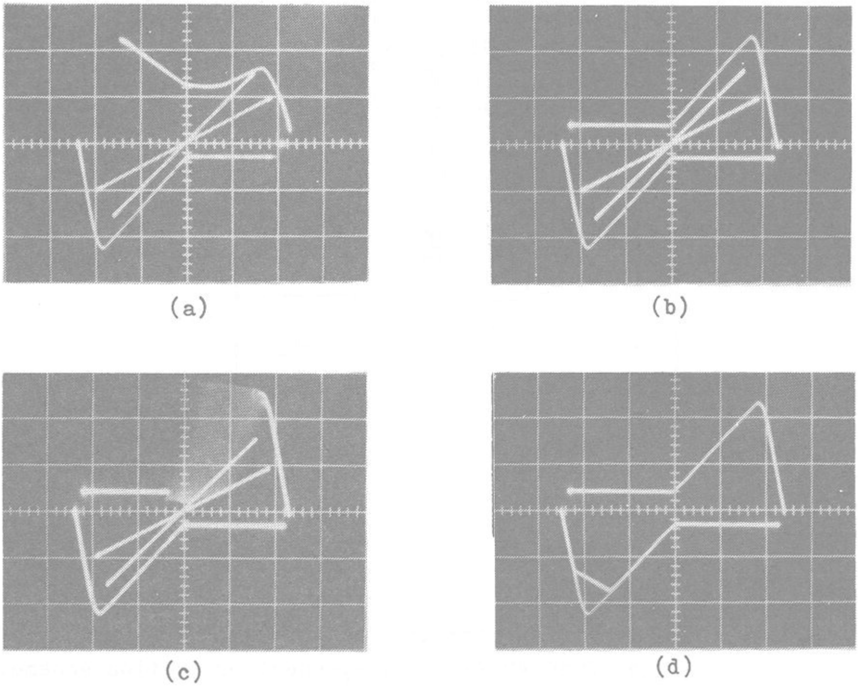


Fig. 9. V-I displays, with 8Ω and 4Ω load-lines, exhibiting faulty behaviour features.

Scales: 5A/cm vert., 20V/cm horiz.

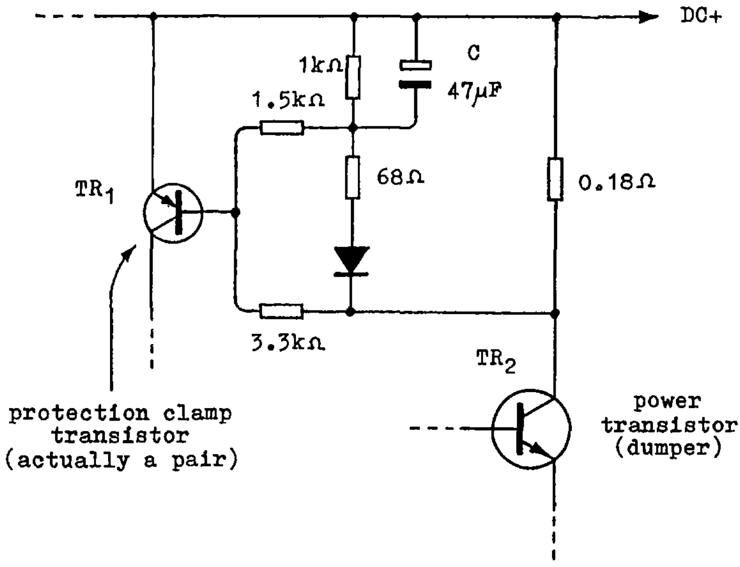


Fig. 10. Simplified circuit illustrating the principle of the Quad 405/2 time-dependent protection scheme.

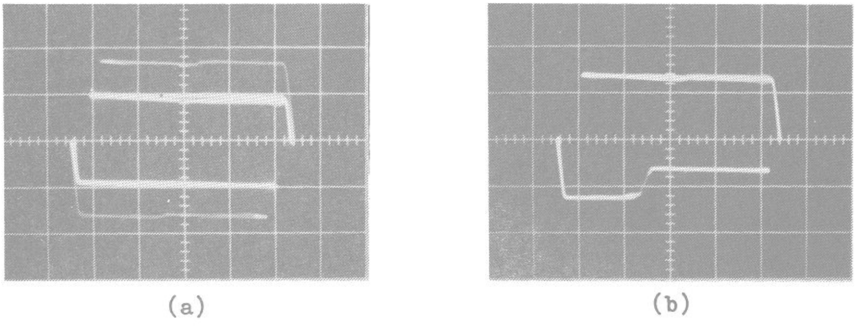


Fig. 11. V-I displays for the Quad 405/2 amplifier.

(a) 50μs pulses at 100Hz (dim) and 2kHz (bright).

(b) 50μs pulses at 500Hz, 47μF in negative-side protection circuit removed.

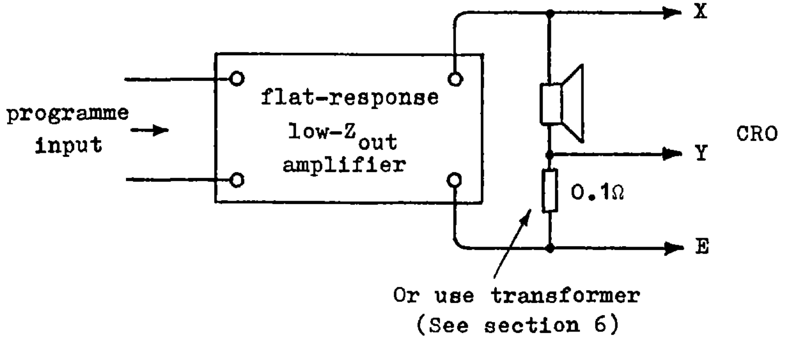


Fig. 12. Set-up for tests on loudspeakers.

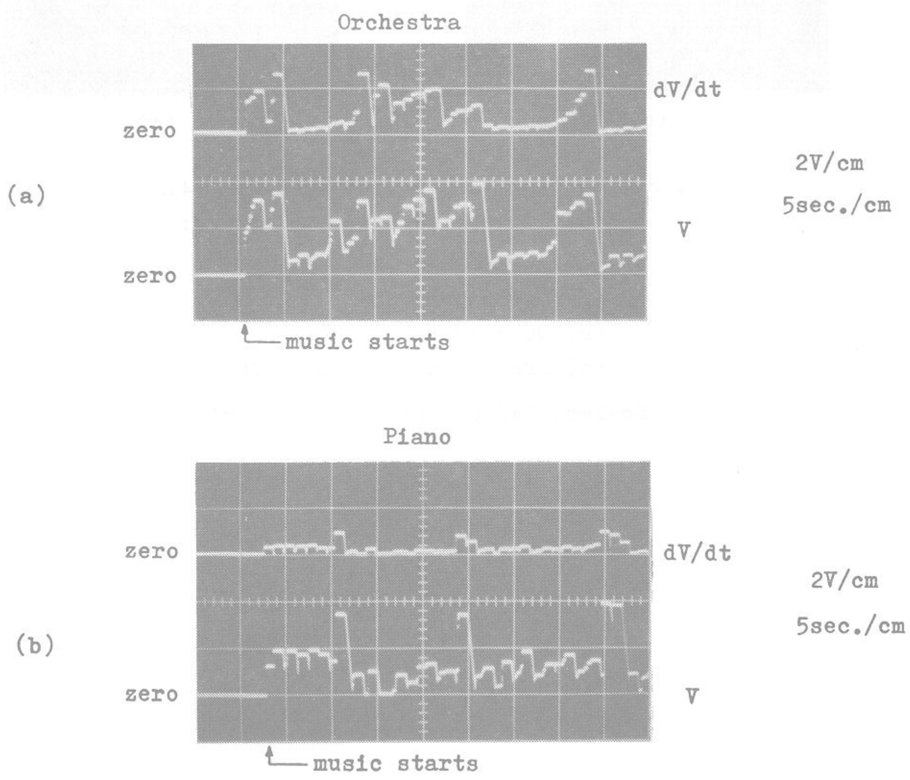
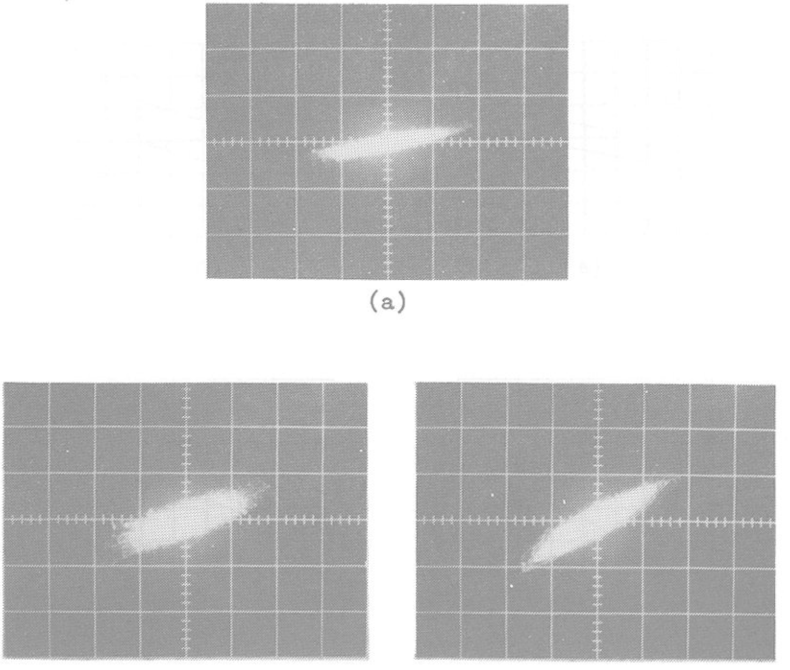


Fig. 13. Results obtained using fast-response linear peak-programme-meter with 1 sec. dwell time.



(a) (b) (c)

Fig. 14. Loudspeaker V-I displays on loud orchestral music with cymbals.

(a) Rogers/BBC LS3/6 (15Ω nom.)
(b) KEF Corelli (8Ω nom.)
(c) Quad ESL63 (8Ω nom.)

Scales: 5A/cm vert., 20V/cm horiz.

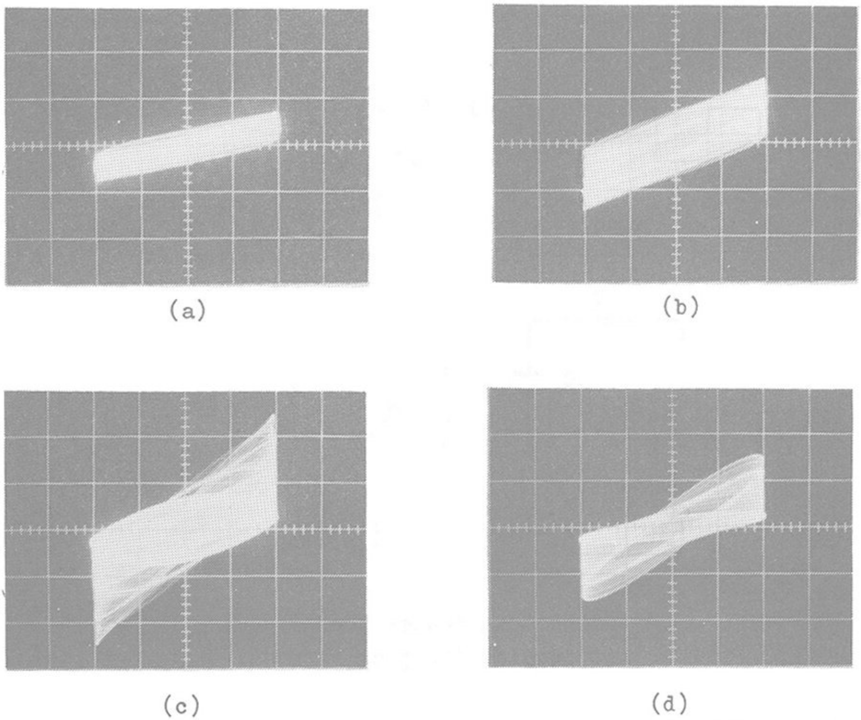


Fig. 15. Frequency-sweeps on three loudspeakers.

(a) Rogers/BBC LS3/6 (15Ω nom.), 20Hz - 20kHz.
(b) KEF Corelli (8Ω nom.), 20Hz - 20kHz.
(c) Quad ESL63 (8Ω nom.), 20Hz - 20kHz.
(d) Quad ESL63 (8Ω nom.), 40Hz - 5kHz.

Scales: 5A/cm vert., 20V/cm horiz.

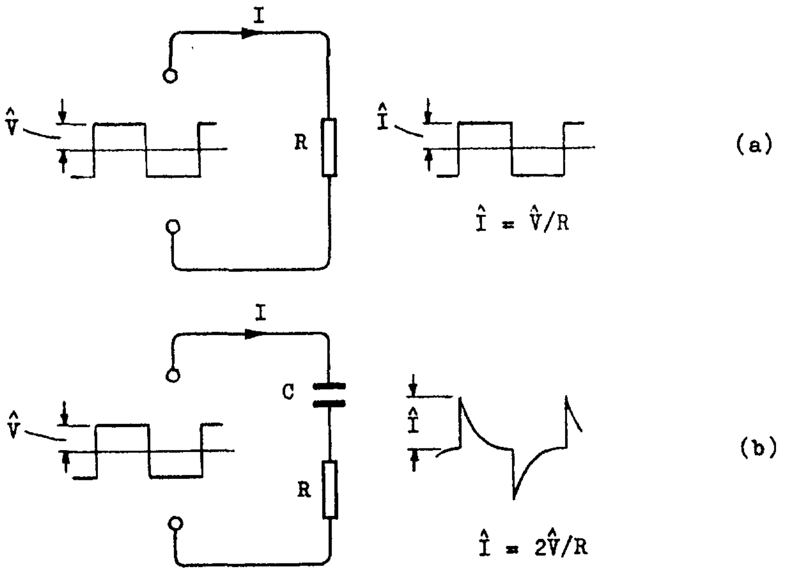


Fig. 16. Adding C increases the impedance but doubles the peak current.

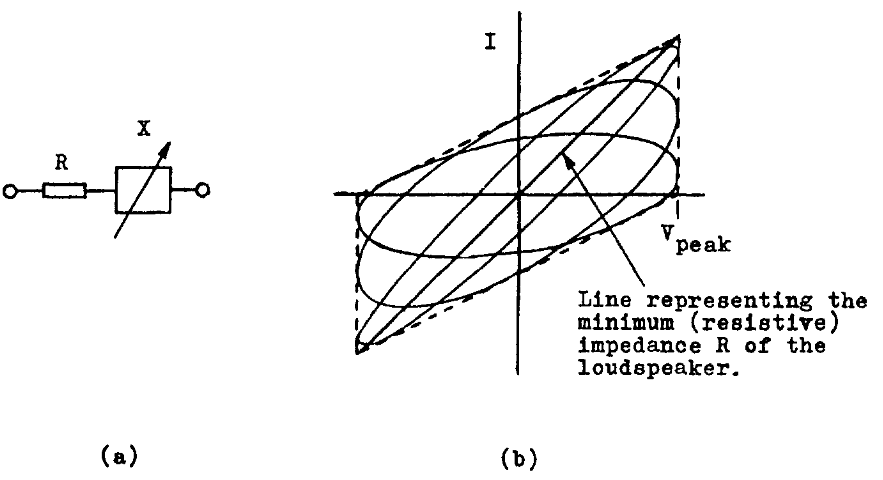


Fig. 17. Load ellipses and tangential parallelogram.

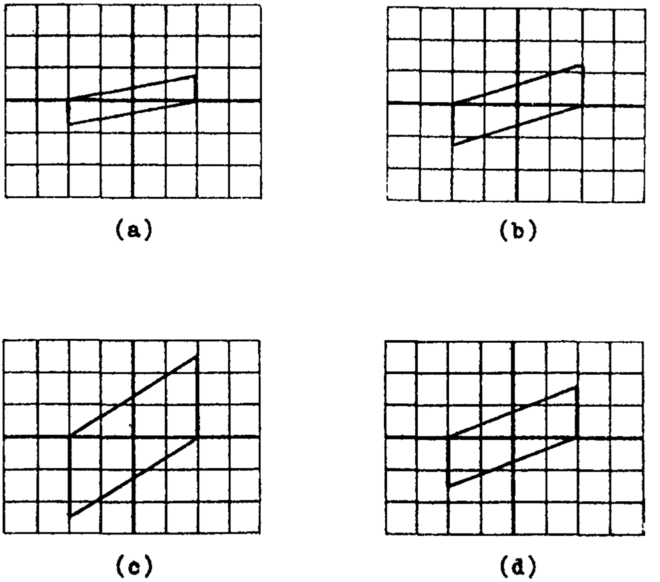


Fig. 18. V-I parallelograms deduced from minimum-impedance values.

(a) Rogers/BBC LS3/6, $|Z|_{min} = 10.7\Omega$, 20Hz - 20kHz.
(b) KEF Corelli, $|Z|_{min} = 6.4\Omega$, 20Hz - 20kHz.
(c) Quad ESL63, $|Z|_{min} = 3.2\Omega$, 20Hz - 20kHz.
(d) Quad ESL63, $|Z|_{min} = 5.2\Omega$, 40Hz - 5kHz.

Scales: 5A/cm vert., 20V/cm horiz.
40V peak signal assumed.