A Technique for Displaying the Current and Voltage Output Capability of Amplifiers and Relating This to the Demands of Loudspeakers*

PETER J. BAXANDALL

Electro-Acoustical Consultant, 17 Victoria Road, Malvern, Worcestershire WR14 2TE, UK

The amplifier under test is fed with a large low-frequency sine-wave input on which 50-μs pulses of alternate polarity are superimposed. These pulses drive the amplifier's protection circuits into current limiting, and the setup 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 program 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, that is, partially reactive, load impedance as it can across a resistive load. 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 setup described is to enable such characteristics to be examined readily.

It has been found that the use of the present testing method sometimes discloses misbehavior of protective circuitry, in the form of overprotection, 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 misbehavior 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 20 Hz, 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 40-μF capacitor on the output is about 200 Ω at 20 Hz, 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 50-μs pulses with a fundamental frequency of normally about 500 Hz. For the pulse component of the amplifier output, the 40-μF 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 1-Ω 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 50-μs pulses fed to the amplifier, the leading edge, of adjustable timing, being delayed by typically 20 μs 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. Fig. 2(a) shows an example. With 50-μs non-delayed-start bright-up pulses, the display of Fig. 2(b) is produced. Delaying the start improves the picture to that shown in 2(c). A

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more striking improvement is sometimes obtained, depending on the nature of the amplifier being tested.

Though the figures of 20 Hz, 500 Hz, and 50 μs 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 Sec. 3.1, there is a minor advantage in using a lower value of current-monitoring resistor than 1 Ω.

2 THE WAVEFORM-GENERATING CIRCUIT

The circuit for producing the required pulse waveforms is shown in Fig. 3. It is 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 div for 20-V 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 operational amplifier (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 was used.

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

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

The occurrence of a positive-going pip at the inverting input then initiates a negative-going change at the op

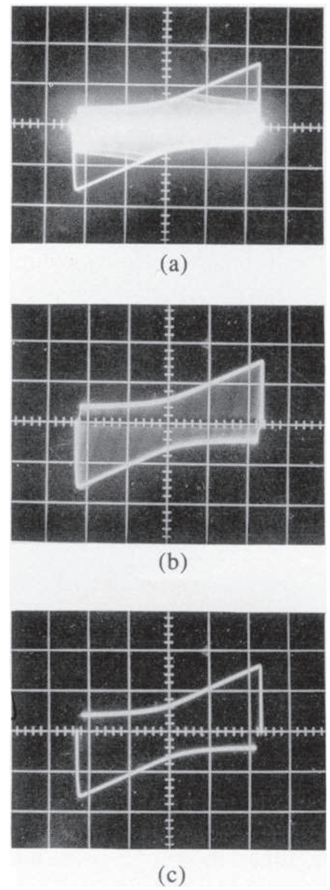

Fig. 2. *V–I* displays for early version of Quad 405 amplifier. (a) With no bright-up facility. (b) With 50-μs non-delayed-start bright-up pulses. (c) With start of bright-up pulses delayed by about 20 μs. Vertical scale—5 A/div; horizontal scale—20 V/div.

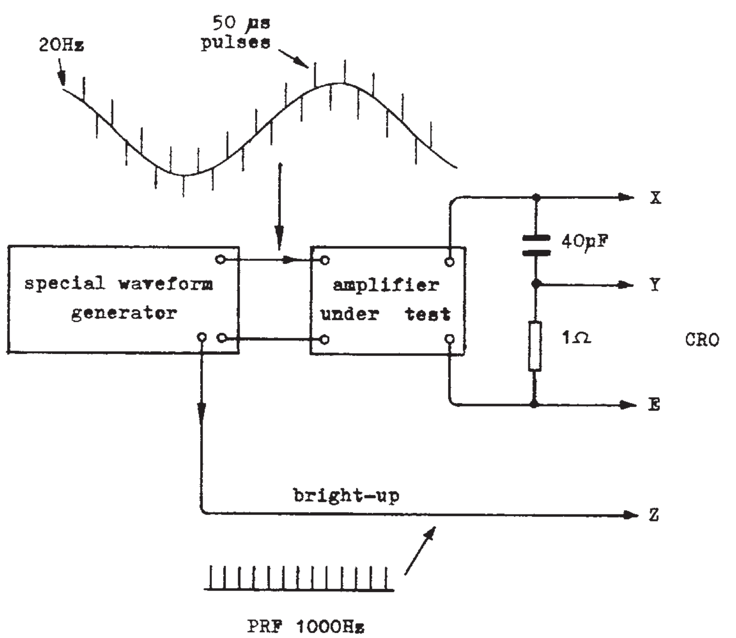

Fig. 1. Essence of the scheme.

amp output, positive feedback via R_3 ensuring that the output reaches its negative overload limit of approximately -16 V. The reason for including resistor R_3 in the feedback path is that it prevents the magnitude of the negative voltage applied to the noninverting 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 noninverting input terminal back up to approximately ground 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 50-μs negative-pulse duration.

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

The positive and negative 50-μs 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 variations in this impedance would affect the amplitude of the 20-Hz 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 500-Hz positive-going 50-μs pulses from op amp 2, so that negative-going pulses occur on 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 point S where these resistors join, the negative-going pulses there, at a repetition frequency of 1000 Hz, would have an excursion from approximately +16 V down to ground.

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

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

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

When a negative pulse occurs at S, then 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, that is, approximately -0.8 V. If the slider of P_2 is set at the top of the track, this -0.8 V on the inverting input is sufficient to produce a pulse excursion from about +16 V down to -16 V at the op amp output, virtually co-

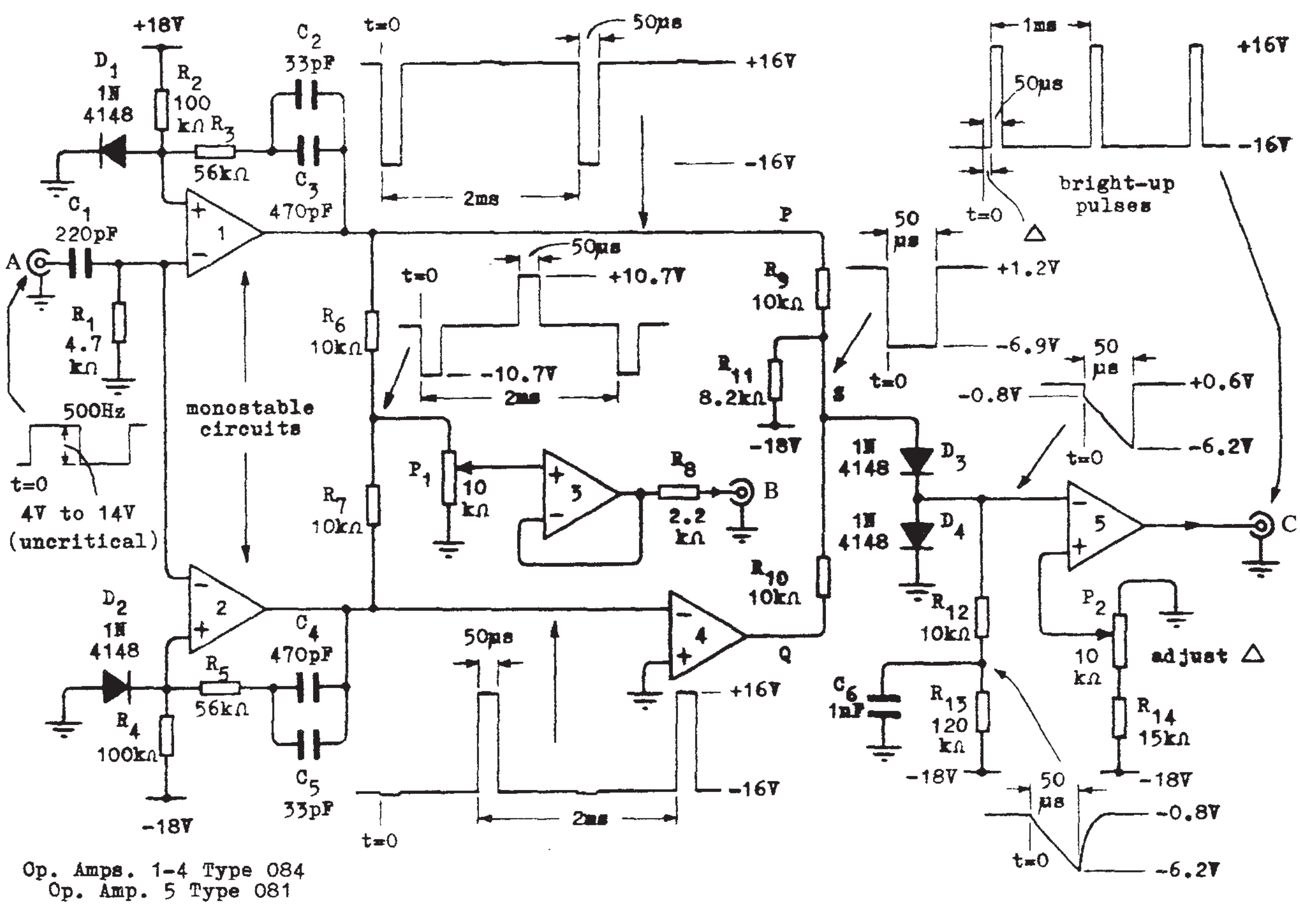

Fig. 3. Waveform-generating circuit.

incident in timing, therefore, with the beginning of the 50-μs 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 −0.8 V 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 the slider of P_2 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 ±9 V 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 circuit of Fig. 3. The input square-wave frequency was temporarily increased from 500 Hz to 2 kHz to enable the pulses to be more clearly displayed. P_2 was set to delay the start of the Z-modulation pulses by about 20 μs.

3 SOME FURTHER POINTS

3.1 The Display Mechanism in More Detail

The adoption of a lower value than 1 Ω 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 point A in Fig. 6. The small low-frequency current component taken by the 40-μF capacitor will be ignored for present purposes. A 50-μs 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 line AB corresponds to the 1-Ω value of the current-monitoring resistor. After B has been reached, the output current continues to flow, charging the 40-μF capacitor and causing the working point to move to C by the time the 50-μs 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 40-μF 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, and bright-up occurs at a point between B and C.

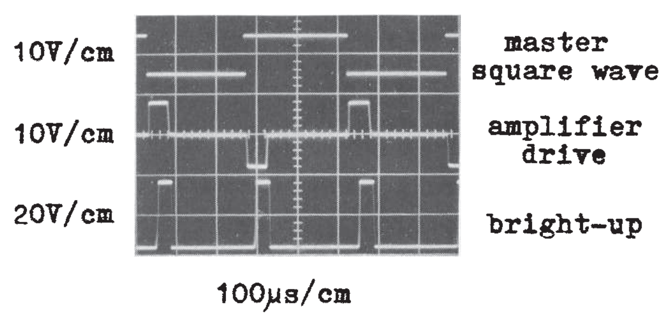

Fig. 5. Waveforms produced by circuit of Fig. 3. Square-wave frequency was increased to 2 kHz for greater clarity.

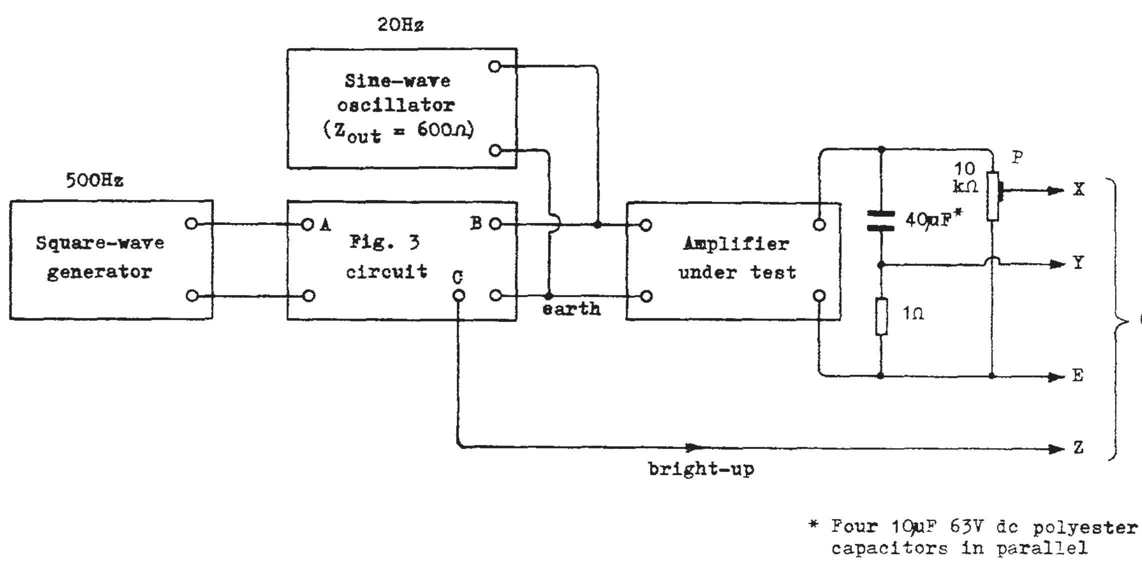

Fig. 4. Complete setup, incorporating circuit of Fig. 3.

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

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

With a sufficiently large positive output voltage due to the 20-Hz input, the point G at the corner of the complete *V–I* characteristic will be reached before the end of the 50-μs 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 line GH is displayed, for blackout 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. The nearest approach is obtained by starting from H, which would enable point K to be displayed if the amplifier had a sufficiently rapid response. Since the slope of HK is 1 Ω, it is clear that, under these conditions, reducing the value of the current-monitoring resistor to a much smaller figure would enable 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 realized.

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 0.1 Ω. This favors the adoption of a higher value, such as 1 Ω, for general use. A better compromise might, perhaps, be 0.5 Ω.

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

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

3.4 Standardized Display Scalings

To facilitate easy comparison of the *V–I* limiter displays for various amplifiers, it is convenient to adopt standardized scalings for current and voltage, and it is suggested that an appropriate choice will normally be 5 A and 20 V, respectively, per division.

This also aids the comparison of amplifier and loudspeaker *V–I* displays, as described in Sec. 9.

4 OPERATING PROCEDURE AND PHOTOGRAPHY

1) To set up for the standardized *X* scaling of 20 V/div, apply only the sine-wave input to the amplifier under test, at 20 Hz or other fairly low frequency. (Beware of making the frequency more than about 200 Hz, as this may burn out the current-monitoring resistor or cause excessive dissipation in the amplifier output transistors.) Put a voltmeter across the amplifier output

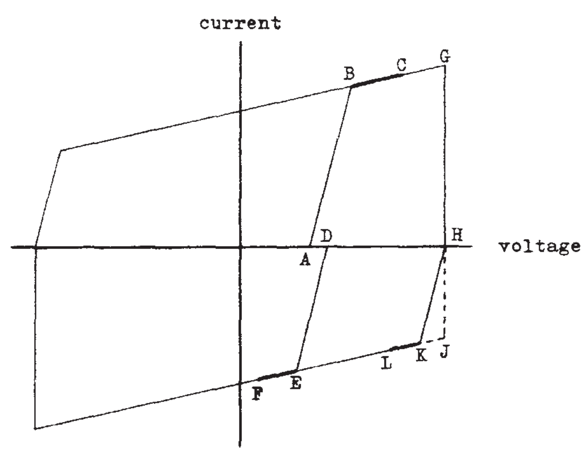

Fig. 6. Display details.

and adjust the input for a reading of 28.3 V rms. Then adjust P (Fig. 4) and/or the CRO X gain for $\pm$2-div deflection.

2) Set the CRO Y sensitivity to 5 V/div to obtain the standardized Y scaling of 5 A div. A 1-Ω current-monitoring resistor is here assumed.

3) With zero signal input, carefully set the spot to the center of the graticule. Now apply 20 Hz or other low-frequency sine-wave input at comfortably sufficient level to make the amplifier overload, that is, 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 around the ellipse, unless the 20 Hz and 500 Hz are in precise integral relationship. For good photographs it is best to set the frequencies so that the dots drift around quite slowly and then give a fairly long exposure, such as 5 s, for the subsequent V–I display photographs.

4) 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 s, though it is probable that most amplifiers would actually withstand continuous stressing in this manner.

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

5 8-Ω AND 4-Ω LINES

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

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

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

With the standardized X and Y sensitivities as described in Sec. 3.4, the 8-Ω line should be at "1 div up for 2 div across" slope, the 4-Ω 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 1-Ω monitoring resistor used straightforwardly gives a sensitivity of 1 V/A, inserts 1 Ω in the monitored circuit, and dissipates 100 W for a current of 10 A. Using, for example, a 1:1000 transformer with a 1-kΩ resistor across the secondary, this same sensitivity of 1 V/A is obtained, but the effective resistance inserted in the monitored circuit is now, ideally, only 1 mΩ, and a current of 10 A now dissipates only 100 mW in the resistor.

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

The point concerning amplifier stability, mentioned in the last paragraph of Sec. 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 Sec. 3.1.

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

The negligible effective series resistance introduced when a transformer is employed is also advantageous when determining the V–I displays for loudspeakers (see Sec. 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 1-Ω 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 0.12-mm-diameter enameled wire (40 SWG, 36 AWG) wound in a single section on a core of 0.38-mm (0.015-in)

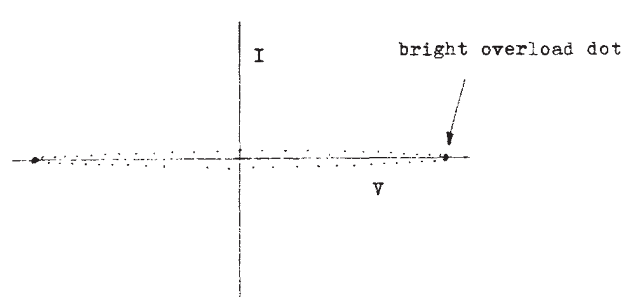

Fig. 7. Display obtained during setting-up adjustments.

Mumetal (Permalloy C) laminations of maximum dimension 25.4 mm (1 in) and stack 12.7 mm (0.5 in); center-limb width 6.3 mm (0.25 in), no. 187 laminations. The primary is a single turn around the outside of the secondary.

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

With an 8-Ω load, on the other hand, overload occurs because of voltage clipping rather than current limitations. A fall in dc supply voltage then causes a reduction in the achievable peak output level.

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 8.2 to 15 kΩ, and Fig. 8(b) shows the display obtained with this modification incorporated.

Fig. 8(c) is for a Quad 303 amplifier, which has a stabilized 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, emphasizing the advantage of adopting a standardized scaling for the photographs.

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

Fig. 9 illustrates some defective performance features, and was 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

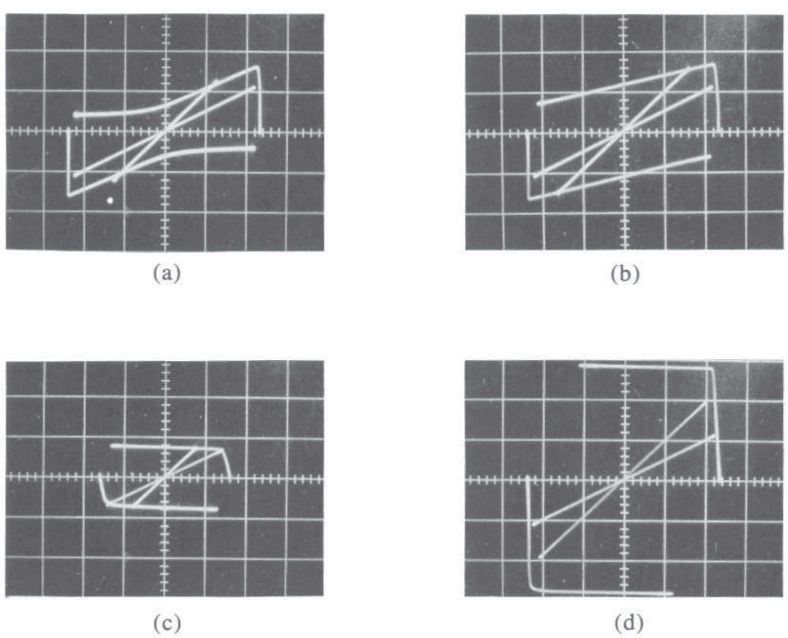

Fig. 8. V–I displays, including 8-Ω and 4-Ω load lines, for four amplifiers. (a) Early Quad 405. (b) Later Quad 405. (c) Quad 303. (d) Preproduction Quad 520. Vertical scale—5 A/div; horizontal scale—20 V/div.

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 sizable heat sink to the amplifying transistor and a 470-Ω 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 10-Ω 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 in Fig. 9(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 display in Fig. 9(b) is, no doubt, as intended by the designer, and it shows that the amplifier comfortably meets its specification of 70 W into 8 Ω and 100 W into 4 Ω, 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 8 Ω.

Analysis shows that if a straight line is drawn from the right-hand tip of, say, the 4-Ω 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 Sec. 8 and Fig. 17.)

Fig. 9(c) shows that RF oscillation occurs when the limiter operates, if a 47-nF 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 predriver PNP common-emitter amplifying stage, was reduced from 47 μF to approximately 8 μF. The low-frequency input was at 20 Hz, as for the other oscillograms in Fig. 9. 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 20 to 50 Hz or above eliminated the above effect from the display. With some amplifiers the effect may be observed, at least mildly, at 20 Hz, even without reducing the value

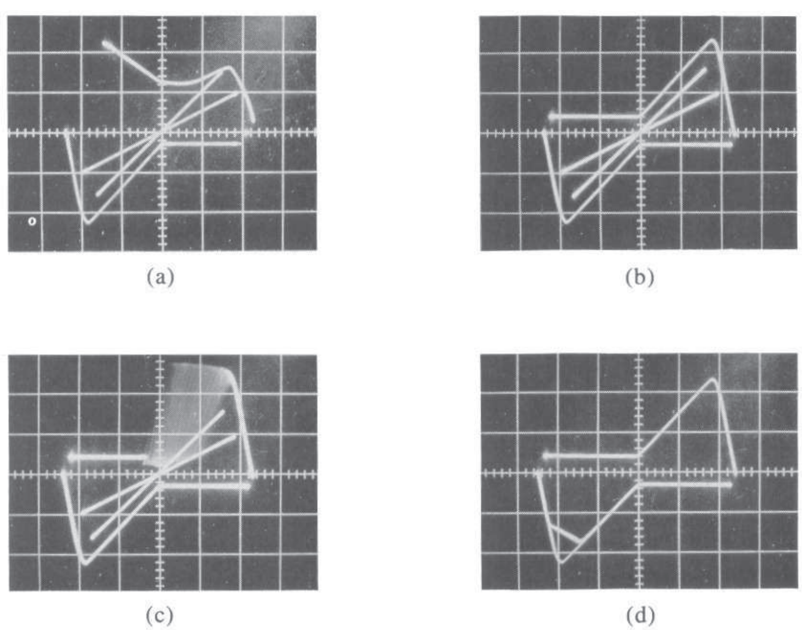

Fig. 9. V–I displays, with 8-Ω and 4-Ω load lines, exhibiting faulty behavior features. Vertical scale—5 A/div; horizontal scale—20 V/div.

of the bootstrapping capacitor fitted.

It will be noticed that the right- and left-hand boundaries of the displays in Fig. 9 are much less steep than most of the previous ones. This is because the amplifier in question has 0.5-Ω 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 Sec. 3.1.

8 PROTECTION CIRCUITS WITH TIME-DEPENDENT BEHAVIOR

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.[1]

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 45 V. 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 a 1.5-V drop occurs across the 0.18-Ω resistor, corresponding to about 8 A. The time constant of 68 Ω and 47 μF is 3.2 ms, 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 0.18-Ω 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 circuits, 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.

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

Fig. 11(b) was obtained with the 47-μF capacitor in the negative-side protection circuit removed, the pulse repetition frequency being 500 Hz. This demonstrates that the effective diode of Fig. 10, but in the negative-

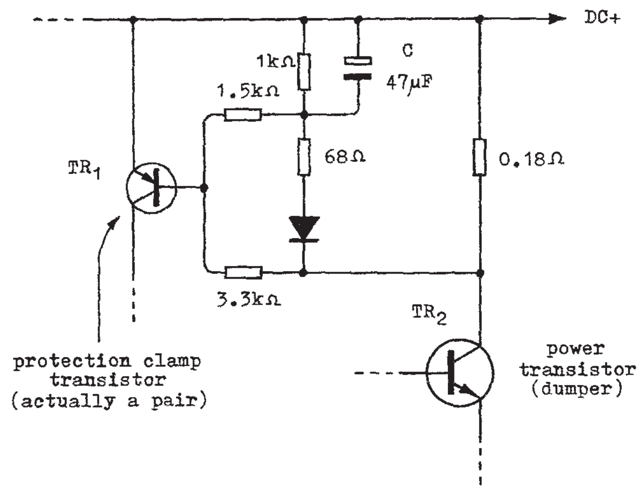

Fig. 10. Simplified circuit illustrating principle of Quad 405/2 time-dependent protection scheme.

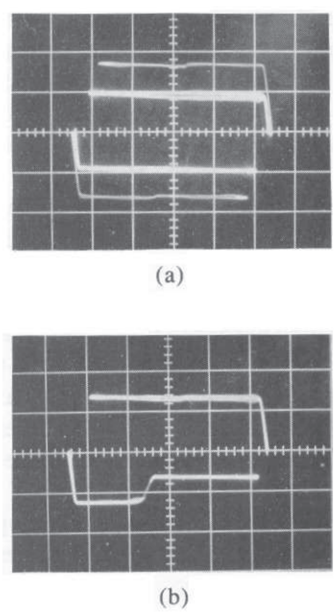

Fig. 11. V–I displays for Quad 405/2 amplifier. (a) 50-μs pulses at 100 Hz (dim) and 2-kHz (bright). (b) 50-μs pulses at 500 Hz, with 47 μF in negative-side protection circuit removed.

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

side protection circuit, becomes functional only when the amplifier output voltage is more positive than about −8 V.

The pulse duration for the Fig. 11 tests was 50 μs, as with the other displays. Much longer pulses may be used, but to avoid excessive mean power dissipation, the pulse repetition frequency should then be much reduced. A satisfactory technique is to make the pulse repetition frequency 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 behavior 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 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 are liable to be only approximately 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 program 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 setup 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 4-Ω 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 program meter designed by the author.

This peak program meter has the following features.

1) Charging time constant about one hundredth of that for a normal peak program meter.

2) Every time a program 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 1 s before starting to fall back, thus facilitating easy and accurate reading of the program peak. If an even higher program peak happens to occur during this 1-s interval, the needle jumps up further to indicate this new peak level, and a new 1-s "dwell time" is initiated.

3) The circuit may be switched to read either the peak values of the program voltage V or 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 5 kHz.

4) 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 s. Fig. 13(a) relates to the beginning of Johann Strauss's Radetsky 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_0, 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_0 values as high as 10 kHz can occasionally be obtained, f_0 for

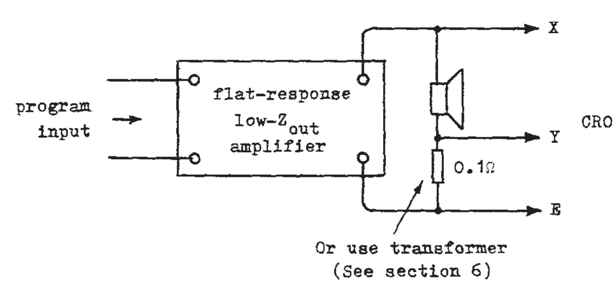

Fig. 12. Setup for loudspeaker tests.

Fig. 13(a) being about 3.5 kHz 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 preceding paragraph, that are likely to lead to problems with loudspeakers whose impedance drops to low values at high frequencies.

The displays of Fig. 14 were obtained with three different loudspeakers, at a fairly low signal level, using the Fig. 12 setup. A suitable 3-s part of the orchestral excerpt from Fig. 13(a) was chosen. The gain was set to obtain a maximum V value on the peak program meter corresponding to 2-div 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 outward every time a larger spot excursion occurs.

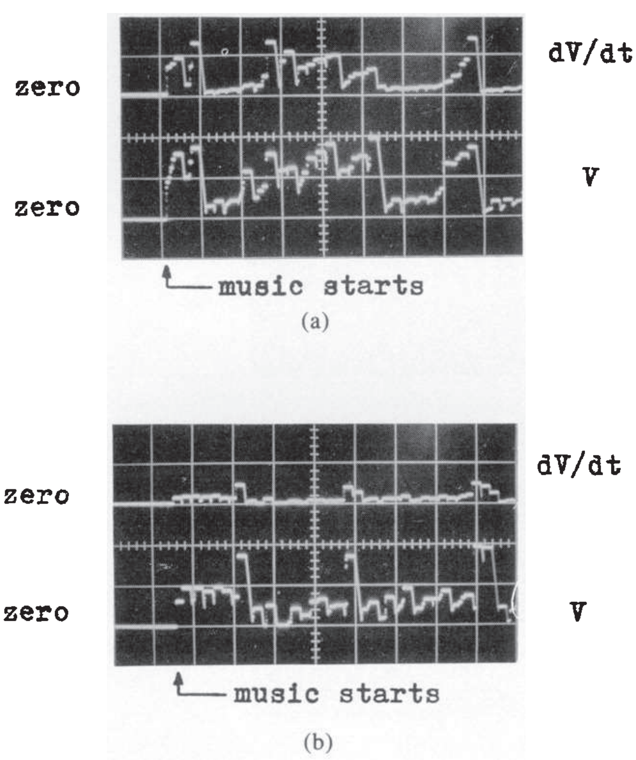

Fig. 13. Results obtained using fast-response linear peak program meter with 1-s dwell time. (a) Orchestra. (b) Piano. Vertical scale—2 V/div; horizontal scale—5 s/div.

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 representing 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 setup of Fig. 12, 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 at least several seconds. 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 oscillograms 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 program input are liable to exceed those for sine-wave voltage input of the same peak magnitude [5], [6]. 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 20 kHz as it is by a low impedance at frequencies which are more significant from a program point of view. This point shows up particularly in Fig. 15(c) for the Quad ESL63, whose impedance modulus at 15 kHz is about 3.5 Ω only. By limiting the sweep to 40 Hz to 5 kHz, instead of 20 Hz to 20 kHz, Fig. 15(d) is obtained. It is clearly more relevant to likely current demands under program 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 8 kHz, the maximum current with square-wave drive occurred at a fundamental frequency of about 4 kHz, 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 program level then needs to be turned down if overloading is to be avoided, and the previously

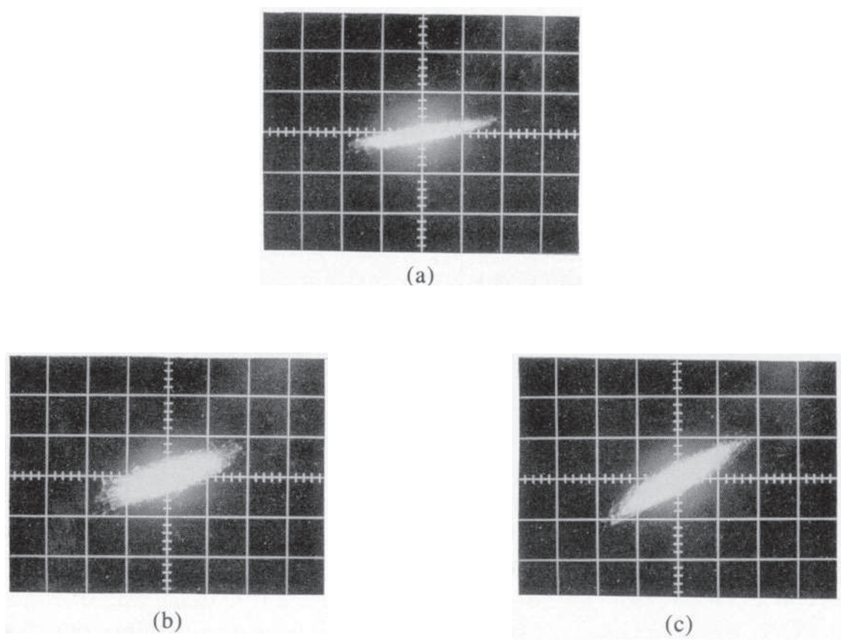

Fig. 14. Loudspeaker $V-I$ displays on loud orchestral music with cymbals. (a) Rogers/BBC LS3/6 (15 Ω nominal). (b) KEF Corelli (8 Ω nominal). (c) Quad ESL63 (8 Ω nominal). Vertical scale—5 A/div; horizontal scale—20 V/div.

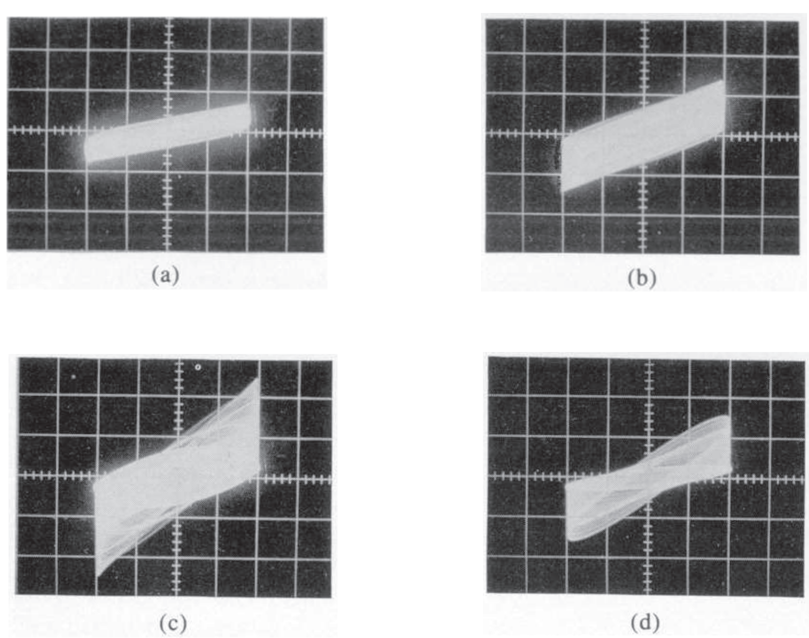

Fig. 15. Frequency sweeps on three loudspeakers. (a) Rogers/BBC LS3/6 (15 Ω nominal), 20 Hz–20 kHz. (b) KEF Corelli (8 Ω nominal), 20 Hz–20 kHz. (c) Quad ESL63 (8 Ω nominal), 20 Hz–20 kHz. (d) Quad ESL63 (8 Ω nominal), 40 Hz–5 kHz. Vertical scale—5 A/div; horizontal scale–20 V/div.

excessive current demands now no longer arise.

Thus although 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 program 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 Sec. 7 and illustrated here in Fig. 17. No matter how complex the equivalent circuit representing the electrical impedance of a loudspeaker may be, 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 output voltage of varying frequency, is therefore a series of ellipses, and it may be shown that these all fall within, and tangenital 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 standardized scaling being adopted again, such that a 45° slope would represent 4 Ω. 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 be able comfortably 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, 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 4 Ω throughout the audio spectrum. The amplifier can therefore produce about 32 V peak across this loudspeaker, no matter what waveform may be involved, corresponding to a peak power of 256 W. 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 64 W, without significant distortion, would be predicted.

The rather awkward problems discussed concerning the relationship between the peak instantaneous current

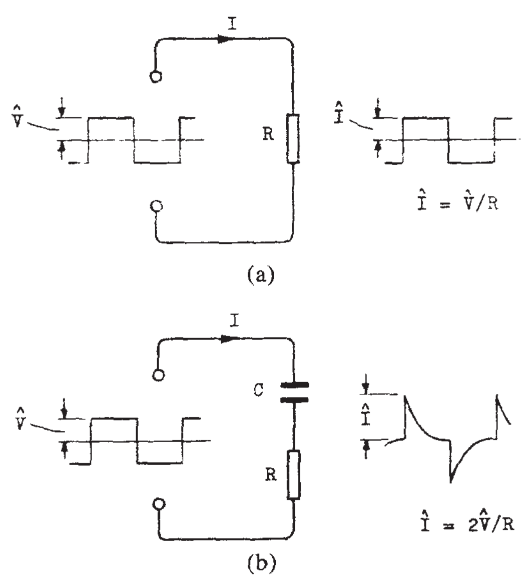

Fig. 16. Adding C increases the impedance but doubles the peak current.

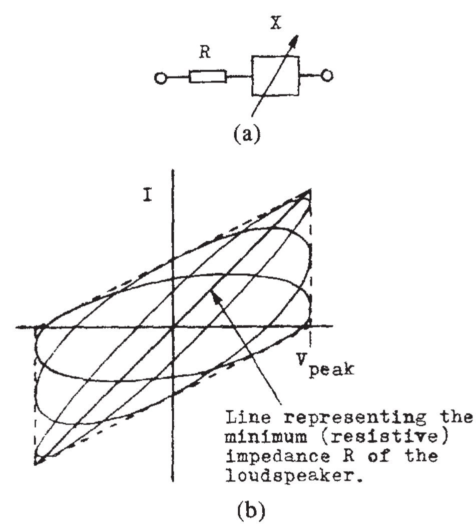

Fig. 17. Load ellipses and tangential parallelogram.

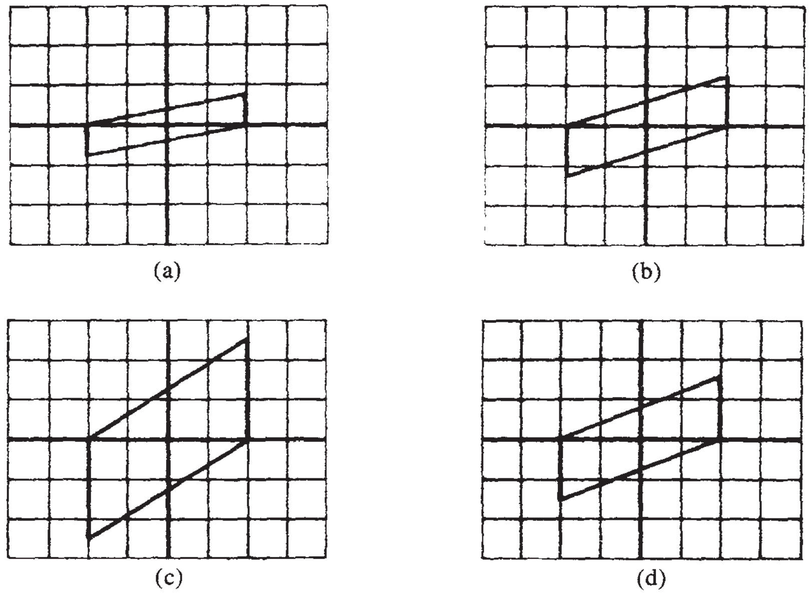

Fig. 18. *V–I* parallelograms deduced from minimum-impedance values. (a) Rogers/BBC LS3/6, $|Z|_{min}$ = 10.7 Ω, 20 Hz–20 kHz. (b) KEF Corelli, $|Z|_{min}$ = 6.4 Ω, 20 HZ–20 kHz. (c) Quad ESL63, $|Z|_{min}$ = 3.2 Ω, 20 Hz–20 kHz. (d) Quad ESL63, $|Z|_{min}$ = 5.2 Ω, 40 Hz–5 kHz. Vertical scale—5 A/div; horizontal scale—20 V/div. 40-V peak signal assumed.

demands of loudspeakers under sine-wave and program conditions completely disappear, of course, ideally, when the conjugate-network technique is used to give a purely resistive impedance.

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ADDENDUM

As a result of this work, a proposal for the addition of a new group of characteristics and measuring methods to IEC Publication 268-3 has been made. This addition is expected eventually to provide measuring methods for the characteristics of all types of protection circuits, including those controlling dc offset at the output terminals as well as those controlling potentially damaging combinations of output voltage and current. The first part of this addition, dealing with the latter subject as considered in this paper, is to be circulated shortly by IEC Technical Committee 84 to IEC national committees under the accelerated procedure.

THE AUTHOR

Peter J. Baxandall was born in 1921 and attended King's College School, Wimbledon, U.K., before reading electrical engineering at Cardiff Technical College and obtaining a B.Sc. (Eng.) degree in 1942.

After two years as a lecturer/instructor in the Fleet Air Arm radio training scheme at the college, he moved to Malvern in 1944 to work in a government radar establishment then known as TRE (now RSRE). After some initial work designing microwave test gear, he gravitated toward the circuitry field, under Professor F. C. Williams, who had been closely associated with A. D. Blumlein. Work included the design of magnetic modulators, low-noise amplifiers, oscillators, wide-dynamic-range phase-sensitive detectors, and active filters, largely for use in connection with research work being pursued in the physics department.

An interest in music and sound reproduction dating back to his schooldays resulted in the spare-time evolution of a negative-feedback tone-control circuit, now well known, in 1950.

In 1971, Mr. Baxandall left RSRE and became an independent electroacoustical consultant. Since that time his projects have included converter circuits for microphones, protective circuitry for loudspeakers, RF bridge circuits for capacitor microphones, amplifiers and power supplies for measuring microphones, a capacitive system for investigating loudspeaker diaphragm break-up modes, the design of oscillators for testing loudspeakers and microphones, and, more recently, the design of some special, very high precision test gear for the National Physical Laboratory, to enable them to check the accuracy and long-term stability of their setup for the reciprocity calibration of measuring microphones.