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SQUARE WAVE TESTING

by F. Langford-Smith and A. R. Chesterman

1. A square wave generator.

In selecting a circuit for development and test it was necessary to choose between squaring a sine wave and some form of multivibrator. Both methods are capable of giving good results, and the cathode-controlled multivibrator was selected for first trial as being comparatively simple and inexpensive, while capable of giving a very rapid rise time. The circuit first tried out (Fig. 1) is that described by G. W. Gray of the RCA Laboratories Division in Radiotronics, Vol. 17, No. 8 (August, 1952), page 133.

This article is well worth reading by those interested — the circuit gives square waves from 50 c/s to 1 Mc/s. In our case the desired range is from 50 c/s to 20 Kc/s. The rise time is stated to be 0.05 micro-second, which is completely invisible on an oscilloscope, being one five-hundredth of the width of the flat top at 20 Kc/s.

Some of our readers may question the omission of grid resistors in the multivibrator, and they are referred to the original article for a full description. The operation is fundamentally sound since the plate currents are limited.

Certain minor changes were made in Gray's original circuit diagram (his Fig. 4 on page 135):—

(1) A pair of 6AQ5 pentodes were used in place of cathode resistors on the multivibrator valves to provide constant current, and so to eliminate the slight tilt in the negative half. This is the modification suggested by Gray (his Fig. 3), except that we used a separate cathode bias resistor for each pentode, one being variable. Any lack of symmetry between the positive and negative pulses can be adjusted by altering the variable 5K Ω cathode resistor. Also, when the multivibrator valve is changed, the variable cathode resistor may have to be adjusted.

(2) The 12AT7 cathode follower load resistances were changed from 200 to 2200 ohms, to increase the output voltage. This increases the effects of capacitance on the high frequency performance, but at 20 Kc/s, on our A.W.A. Model A56031 oscilloscope, the difference was hard to see. At frequencies greater than 20 Kc/s it would be necessary to use the 200 ohm resistors.

(3) Radiotron Type 12AT7 was used as multivibrator in place of type 2C51 in the original circuit, for greater availability and essentially the same performance, thanks to the two 6AQ5 pentodes used as cathode load impedances. Thus the same valve type is used in the multivibrator and the cathode-follower output.

(4) The capacitances of the frequency controlling capacitors are derived from Fig. 2 of Gray's article, but the graph is so small that it can only be read with an accuracy of about 20%. There is additional inaccuracy from the tolerances in the capacitors themselves, so that quite a large total error may be expected. In our case the frequencies before correction were:—

0.004 μF	=	28.1 Kc/s.
0.005 μF	=	11.5 Kc/s.
0.02 μF	=	5.4 Kc/s.
0.1 μF	=	1140 c/s.
1.1 μF	=	57 c/s.

These may all be lowered to the correct frequency by adding a shunt capacitance of suitable value. Of course, for most applications, a slight error in frequency is not very important.

Operation.

The output voltage is about 18 volts peak-to-peak. The wave form at the higher frequencies was very good.

The 50 c/s waveform showed a top tilt of 2.5%, and bottom tilt 1%. In addition, there was a slight rounding at the far end of the flat top amounting to an additional 2.5% of the peak-to-peak voltage. This performance is entirely satisfactory for measuring overshoot and recovery time in amplifiers.

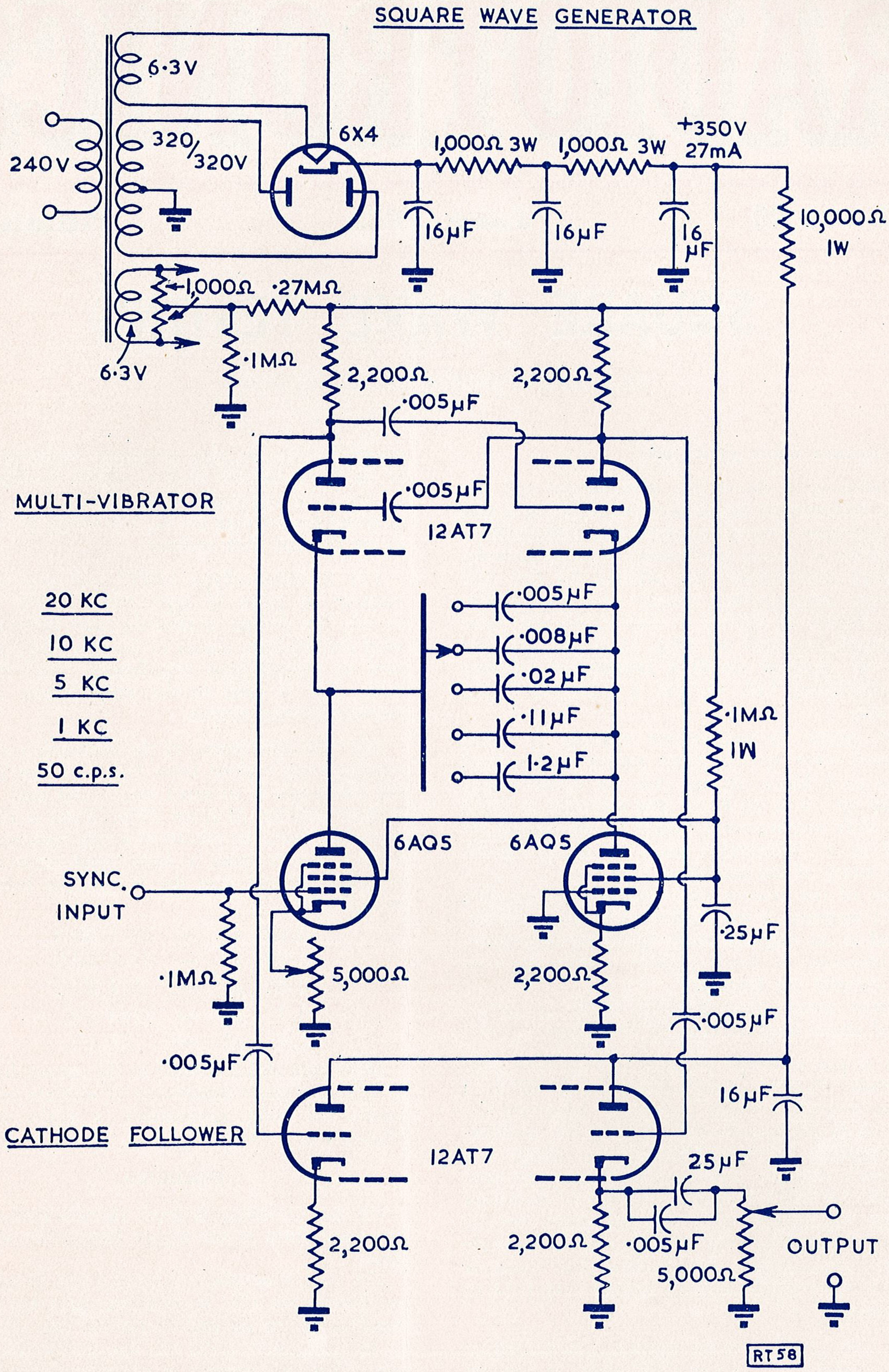

Fig. 1. Circuit diagram of square wave generator (RT58).

See note in March '56 issue, P.35, concerning the 5k Ω output control pot. A higher res. must not be used. Short connecting leads to the amplifier are necessary too.

The rise time is so rapid (0.05 μsec.) that it is a more severe test than is normally used for testing amplifiers. The 50 c/s top and bottom tilt simply mean that with extremely high fidelity amplifiers it is necessary to make a comparison between the input and output waveform.

All 12AT7 valves do not work equally well as multivibrators in this circuit, and it is suggested that the two 12AT7 valves be interchanged and left in the position giving the better 50 c/s waveform. Finally, the output may be taken from either cathode of the push-pull cathode followers, and here again the better side may be selected.

The performance at the higher frequencies is not much affected by interchanging valves.

We plan to develop a more elaborate square wave generator some time in the future for use in our own laboratory, but we honestly feel that for most applications this simple circuit is all that is necessary.

2. Oscilloscope for use with a square wave generator.

It is important to have an oscilloscope with a vertical amplifier with response up to at least 500 Kc/s, and without any overshoot in itself. The great majority of conventional oscilloscopes in use in Australia do not give any indication when overshoot commences, and are useless for such measurements owing to their limited high frequency response. A push-pull vertical amplifier is much preferable to a single-ended one, but there is no necessity for it to be a d.c. amplifier. There is a relationship between the rise time and the bandwidth of an amplifier, which may be expressed in the form (Ref. 1):

$$rB = 0.35 \text{ to } 0.45 \qquad (1)$$

(0.35 when the overshoot is lass than 5%)

where r = rise time in microseconds from 10% to 90% of the step function response of a low-pass amplifier without overshoot,

and B = bandwidth in Mc/s for response 3 db down.

For instance, in the case of an oscilloscope 3 db down at 0.5 Mc/s, with overshoot less than 5%, the rise time is

$$r = \frac{0.35}{0.5} = 0.7 \ \mu\text{sec.}$$

An old type oscilloscope, 3 db down at 50 Kc/s (= 0.05 Mc/s), with no overshoot, would have a rise time

$$r = \frac{0.35}{0.05} = 7 \ \mu\text{secs.}$$

Note that 7 microseconds is equal to 28% of the flat top at 20 Kc/s.

A rise time of 0.05 microsecond in an oscilloscope with overshoot greater than 5% requires a bandwidth of

$$\frac{0.45}{0.05} = 9 \text{ Mc/s.}$$

The oscilloscope used in our Radiotronics Laboratory is the A.W.A. Model A56031, which is fitted with a vertical amplifier designed and factory-adjusted to give good square wave response. The push-pull amplifiers are neutralized to control the overshoot and gain at high frequencies. When neutralized correctly they give no overshoot on square waves.

In order to check an oscilloscope, connect the square wave generator, working at 20 Kc/s, directly to the oscilloscope, using short leads. The generator should be allowed plenty of time to warm up and settle down to steady operating conditions. The effect of the volume control in the generator on the output waveform is practically nil, so it may be set to any desired position. On the other hand the stepped attenuators in most oscilloscope vertical amplifiers have quite noticeable effect on the waveform, and this effect should be minimized by setting it to the highest gain position (i.e., no attenuation).

In all cases it is wise to check the variations in square wave response with different stepped attenuator settings. If the variations are noticeable it is advisable to select the attenuator position giving minimum overshoot — probably zero attenuation in most cases — and to use the same position for all square wave tests wherever possible.

3. Measurements with square waves.

Square waves are usually treated as a special form of pulse (Fig. 2), which rises from a base line, the height of the pulse being the total peak-to-peak voltage (AB). This differs from the conventional sine wave approach, and all voltage measurements are peak-to-peak.

The most useful measurements to be made on feedback amplifiers are described below. These are normally carried out with a resistive load, or with both a resistive load and a loudspeaker load, since the loudspeaker may have a noticeable effect at certain frequencies. If a resistive load alone is used, it should be shunted by the value of capacitance which will give the maximum overshoot. This value may be determined by some form of decade capacitance box, going upwards in capacitance steps to the limit suggested in Table 1.

Table 1.

Shunt capacitance to give 60° phase angle at 20,000 c/s.

Impedance (ohms)	(Capacitance (μF)
2.0	7
3.0	5
4.0	3.5
6.0	2.5
8.0	2
10.0	1.5
12.0	1
15.0	1

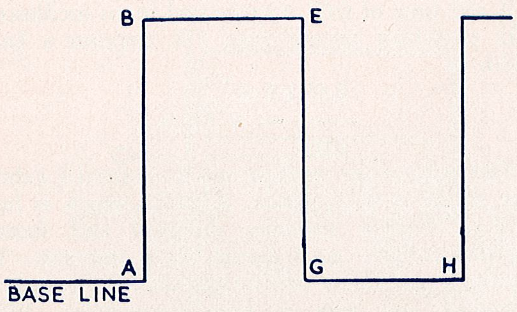

Fig. 2. A square wave shown as a rectangular pulse rising from a base line (RT59).

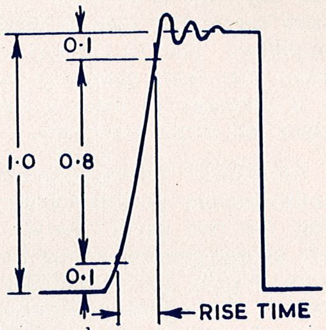

Fig. 4. Rise time for a pulse or square wave (RT61).

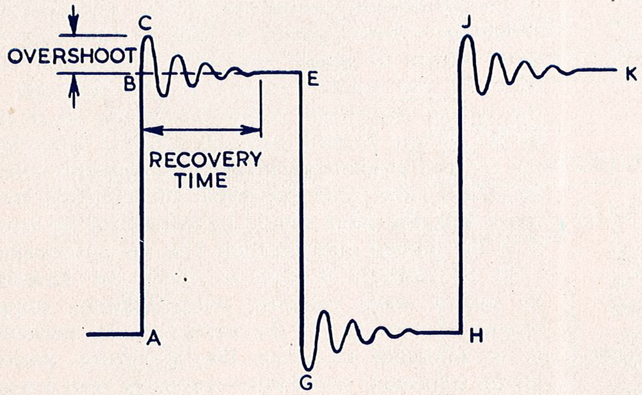

Fig. 3. Square wave response of an amplifier showing overshoot and damped transient oscillation (RT60).

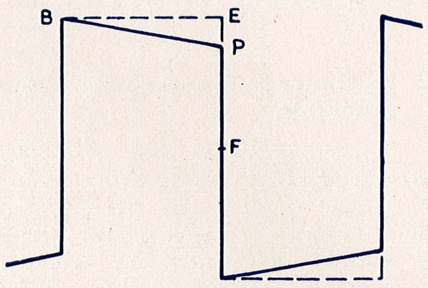

Fig. 5. Square wave response of an amplifier showing top and bottom tilt (RT62).

An amplifier should be sufficiently stable to be used with any likely form of loudspeaker crossover network, most of which present highly reactive loads at certain frequencies. A case is known where a Williamson type amplifier was stable on a single loudspeaker, but unstable on a dual system.

It is hoped to carry out tests in our laboratory with various amplifiers under all practicable loading conditions, and to cover the subject at some later date by another article in Radiotronics.

3.1. Overshoot.

The overshoot (BC in Fig. 3) is measured as the distance moved beyond the flat top, BE. It is expressed as a percentage of AB, that is the peak-to-peak voltage. It may be measured at any desired frequency. The overshoot is theoretically constant at all applied square wave frequencies, with an amplifier flat from zero frequency to infinity. However, the overshoot peak occupies only a very small portion of the half-cycle at low frequencies, and is therefore difficult to observe. Our usual practice is to use a square wave frequency of 5 or 10 Kc/s, where the overshoot may readily be observed. Usually, in amplifiers with restricted frequency response the greatest value of overshoot occurs with a square wave frequency of 5 Kc/s, while with very wide range amplifiers the greatest overshoot may occur with a square wave frequency of 10 Kc/s. The test frequency chosen should normally allow a reasonable amount of flat top to show. With care in design, it is possible to keep overshoot below 5%.

3.2. Transient recovery time.

This is the time in microseconds for the damped oscillation following the overshoot peak to die down — see Fig. 3. In well-designed amplifiers the recovery time should be a small fraction of the flat top at 5 Kc/s. Arbitrarily, we may take one quarter of the flat top at 5 Kc/s, equivalent to 25 microseconds, as the maximum permissible recovery time.

3.3. Rise time.

This is usually measured as the time in microseconds for the wave to rise from 10% to 90% of the peak-to-peak value — see Fig. 4. The relationship between rise time and bandwidth is given in eqn. (1).

3.4. Top tilt.

This is a measure of the low frequency characteristics of the amplifier. It should be carried out at the lowest available frequency — here 50 c/s. It is expressed by the ratio of the "sag" (EP in Fig. 5) to the peak-to-peak voltage, expressed as a percentage.

The measured top tilt at 50 c/s on the 4.5 watt single-ended amplifier described in Radiotronics (February, 1955), was 8.5% of which 2.5% was due to the square wave generator and the balance of 6% was due to the amplifier. This is considerably better performance than that of a typical radio receiver, but is not so good as can be achieved in a push-pull high fidelity amplifier.

3.5. Bottom tilt.

If the tilt of the lower half-cycle differs from that of the upper, both values should be stated.

References.

1. G. E. Valley and H. Wallman, "Vacuum Tube Amplifiers", M.I.T. Radiation Series, McGraw-Hill Book Company. Section 2.

Table 2.

Frequency (c/s)	Length of flat top (micro-secs.)
50	10,000
1,000	500
5,000	100
10,000	50
20,000	25

EFFECT OF DAMPING IN REDUCING LOUDSPEAKER DISTORTION

See comment on P.23 of Feb. 56 issue.

by F. Langford-Smith

Since the article on damping for loudspeakers appeared in the April issue, a most interesting article by Wilkins has appeared (Ref. 1). His Fig. 3 shows curves of power output for 5% total harmonic distortion with three degrees of damping. His tests were conducted, using a microphone, for the overall system and the distortion is almost entirely that of the loudspeaker. There was a slight peak at about 80 c/s, probably indicating bass resonance.

From his curves the following arresting information has been derived:

Frequency c/s	Improvement in power output for 5% THD: DF* changed from +3 to +30	DF* changed from +30 to —1.2
200	5 db	3 db
100	6	15
80	7	16
60	7	11
40	6	12
30	11	19
20	12	28

$$^{*}\text{DF} = \text{damping factor} = \frac{\text{voice coil d.c. resistance}}{\text{output resistance}}$$

Column 2 gives the improvement in power output when the damping factor is changed from +3 (roughly that of triodes without feedback) to +30 (a typical high fidelity amplifier). The average improvement is about 6 db down to 40 c/s, reaching 12 db at 20 c/s. This is startling enough.

Column 3 gives the improvement in power output when the damping factor is increased from +30 (typical high fidelity amplifier) to —1.2, a value reached with positive current feedback, as in the Bogen D030A Amplifier. The feedback circuit of this amplifier is given in Fig. 5, on page 41 of Radiotronics for April, 1955. The improvement averages about 15 db from 100 down to 40 c/s, 19 db at 30 c/s, and 28 db at 20 c/s.

There is no doubt that both amplifier and loudspeaker designers will be doing some hard thinking. One snag is that the control on the Bogan amplifier is not the sort of control to put into the hands of an untrained person, since oscillation may occur and damage the loudspeaker. Another problem is that a large amount of negative voltage feedback is required in addition to the variable positive current feedback, and stability problems would be acute. As a consequence it would probably be limited to the most expensive high fidelity amplifiers.

One query comes to mind on this question of very heavy damping — although damping reduces "hangover", will it not also increase the rise time of a transient, and, if so, is it altogether desirable for highest fidelity?

References.

1. C. A. Wilkins, "Reducing Loudspeaker Distortion", Radio & Television News (April, 1955), 48.

Part 2:

"ULTRA LINEAR" AMPLIFIERS

by F. Langford-Smith and A. R. Chesterman.

This article treats 6V6-GT operation and makes a comparison between UL operation and the divided plate and cathode loading method used in the QUAD II and other amplifiers.

1. Type 6V6-GT UL operation.

Fig. 1A shows the power output versus tapping point for plate and screen 285 volts, and bias —19 volts, for selected values of load resistance. Fig. 1B shows the total harmonic distortion.

Figs. 2 and 3 show the same information for bias values of —21 and —22.5 volts. From these it appears that the 5% tap is the best all-round compromise, giving minimum distortion for —21 volts bias and 8000 ohms load resistance.

Fig. 4 shows the power output and distortion against grid bias. These confirm the impression given by the other figures, indicating a bias of —21 volts as optimum for power output and giving reasonably low distortion.

The load resistance of 8000 ohms plate-to-plate was selected as optimum, giving an output of 10.4 watts at 0.72% THD, even though an output of 11.2 watts was obtainable with a load resistance of 10,000 ohms. The reason for the choice is partly to make it less sensitive to increases in load resistance such as always occurs with a loudspeaker load, and partly to make the transformer simpler and with fewer primary turns.

Interchange "B" curves, as blocks were transposed.

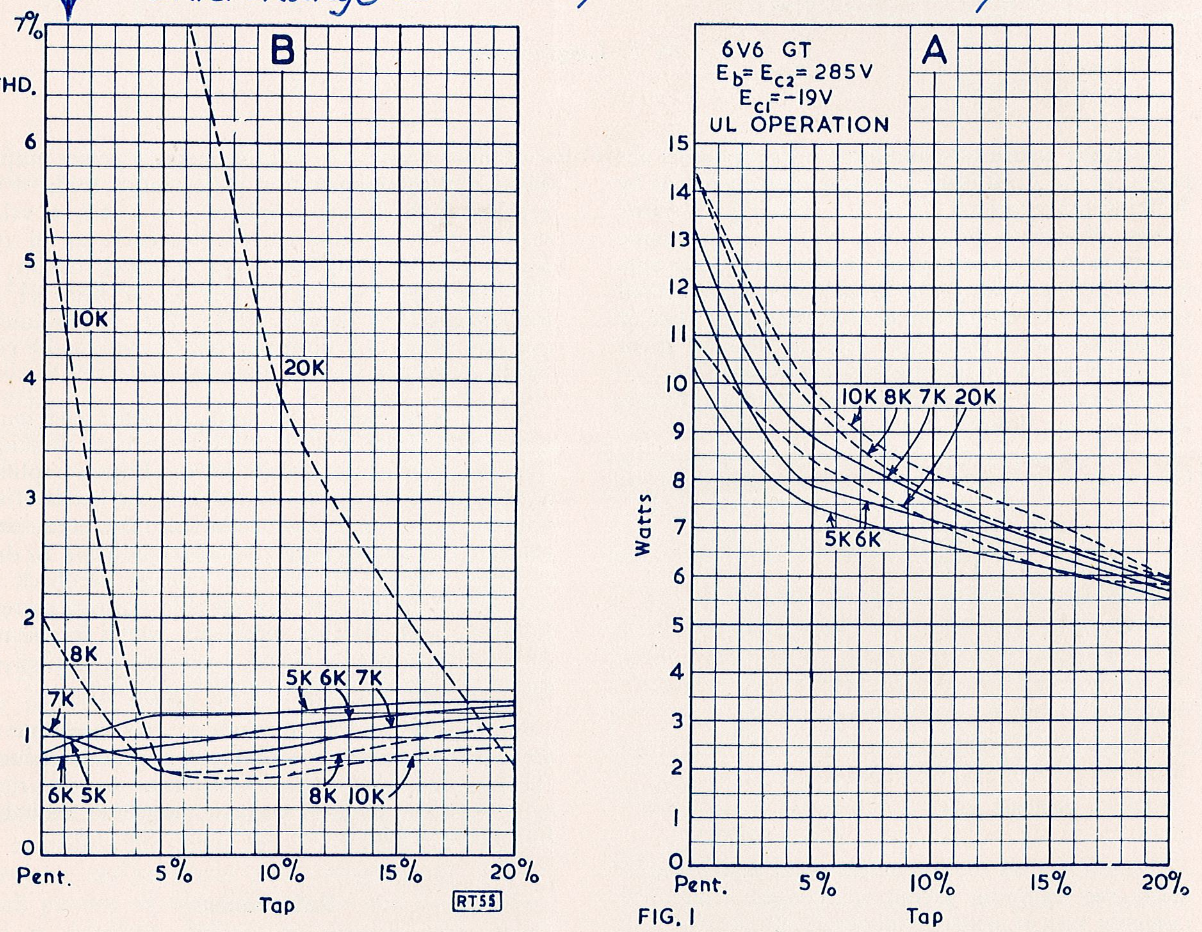

Fig. 1. 6V6-GT push-pull UL operation, $E_b = E_{c2} = 285V$, $E_{c1} = -19V$, peak grid voltage equals bias; (A) Power output versus tap; (B) Total harmonic distortion versus tap (RT54).

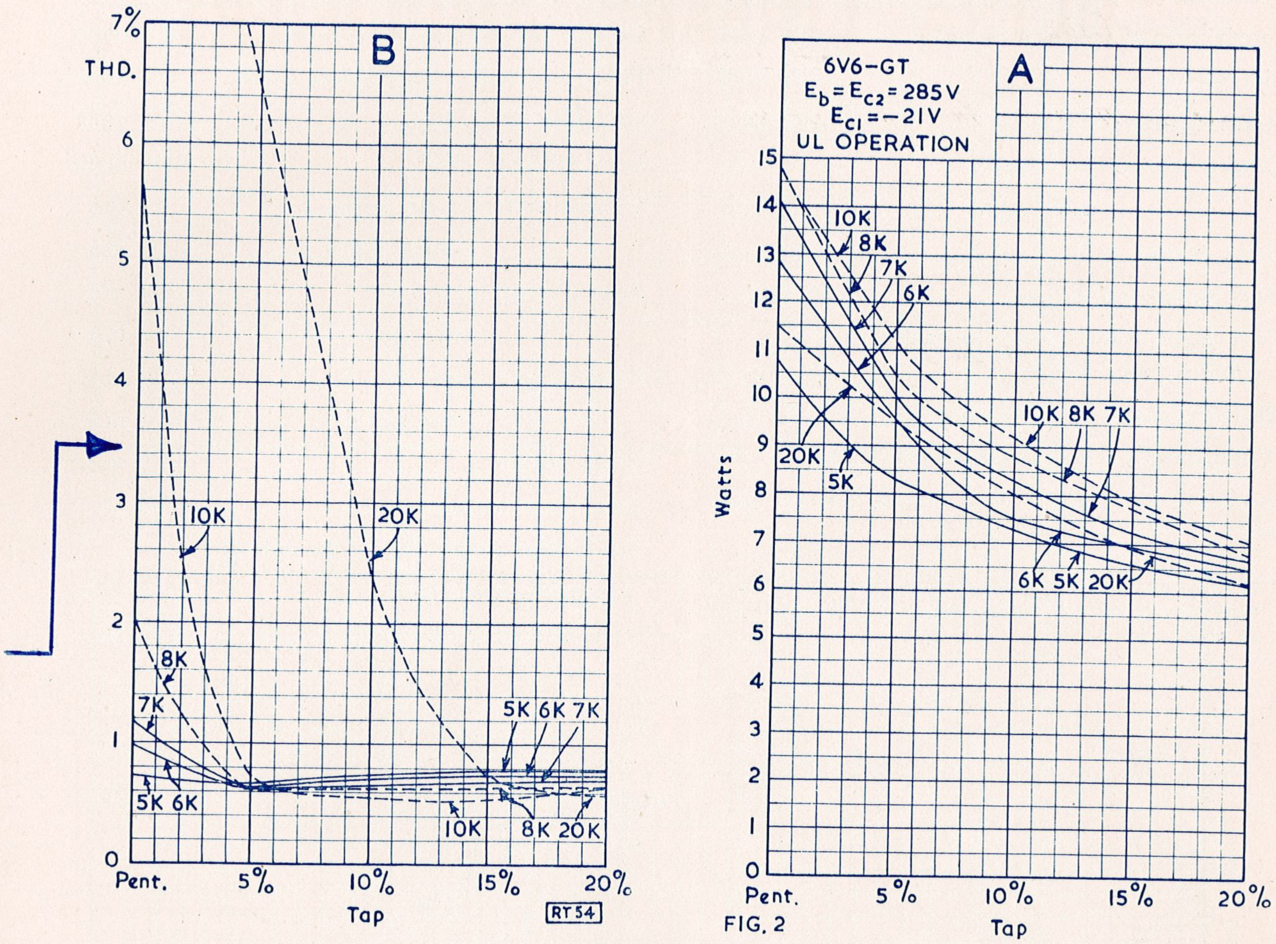

Fig. 2. 6V6-GT push-pull UL operation, $E_b = E_{c2} = 285V$, $E_{c1} = -21V$, peak grid voltage equals bias; (A) Power output versus tap; (B) Total harmonic distortion versus tap (RT55).

If high power output had been unimportant, a tapping point of 15% or 20% might have been selected to make the load resistance less critical for distortion. However, this distortion is less than that with pentode operation. The whole question of distortion with high impedance loads, such as occur with loudspeakers, will be treated in full detail in a later article in this series, and comparisons made between UL, triode and pentode operation.

The selected condition gives 71% of the output obtainable with the same valves as pentodes under the same conditions except that the bias is the published value (—19 volts).

The curves and other data are for operation in the conventional way with the peak signal voltage equal to the bias ("zero grid"). Of course, in any practical amplifier, it is not possible to drive to zero grid without grid current and consequent distortion, and the consequent reduction in effective power output applies to all types of operation. The output resistance under these conditions without external feedback is 15,400 ohms plate-to-plate. With the 20% tapping, this figure would become 9,600 ohms.

Conditions for fixed bias (values are both valves):

Supply voltage (plate to cathode) ...	285 volts
Bias voltage	—21 volts
Load resistance (P-P)	8000 ohms
Cathode current: zero signal	58 mA
max. signal	85 mA
Power output	9.8 watts
Total harmonic distortion	0.8 %
Total input power (based on cathode current), max. signal	24.2 watts
Efficiency (based on cathode current).	40.5 %

(Compare this with the tetrode efficiency 46.5 %)

Conditions for cathode bias (values are for both valves):

Supply voltage	306 volts
Cathode bias resistor (each valve)* ..	430 ohms
Plate to cathode voltage (max. signal)	285 volts
Grid bias (max. signal)	20.2 volts
Load resistance (P-P)	8000 ohms
Cathode current: zero signal	83 mA
max. signal	94 mA
Plate current: zero signal	75.4 mA
max. signal	82.8 mA
Screen current: zero signal	7.6 mA
max. signal	11.2 mA
Power output	9.6 watts
Total harmonic distortion	0.98 %
Total input power (based on cathode current), max. signal	26.8 watts
Cathode current efficiency	36 %

* With each cathode resistor adequately bypassed. Alternatively a common cathode resistor of 220 ohms could be used.

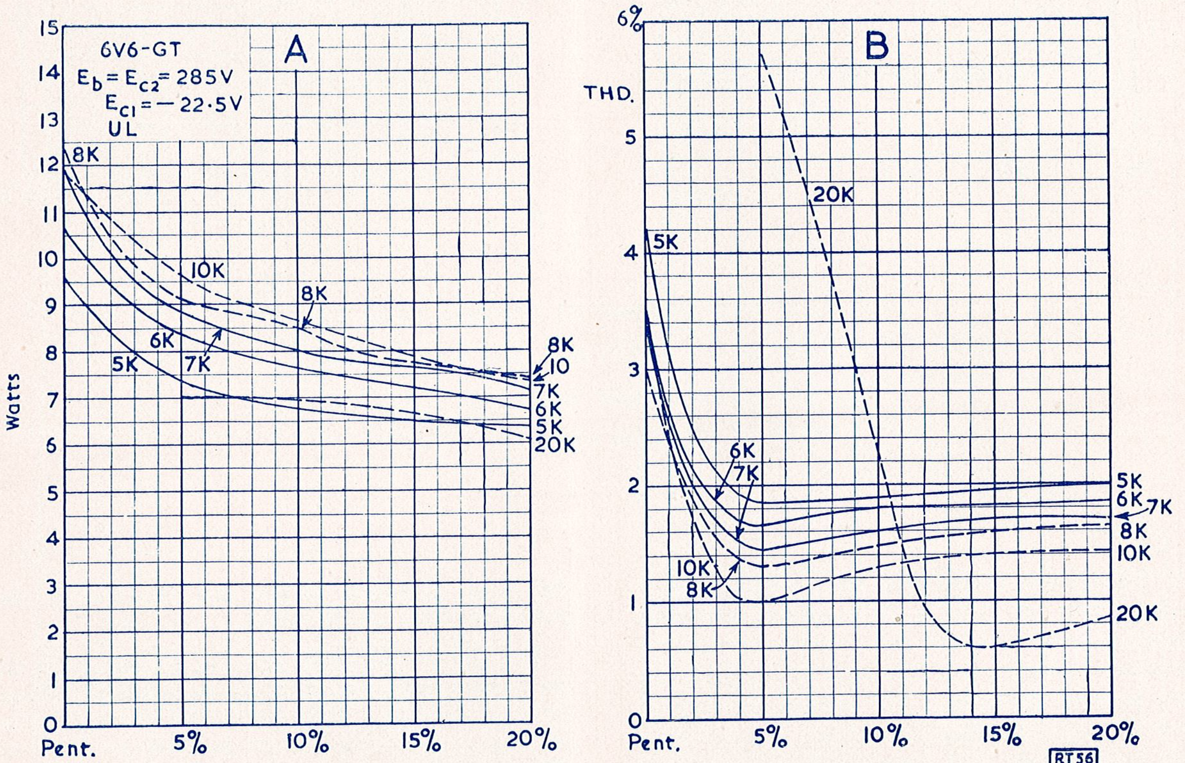

FIG. 3

Fig. 3. 6V6-GT push-pull UL operation, $E_b = E_{c2} = 285V$, $E_{c1} = -22.5V$, peak grid voltage equals bias; (A) Power output versus tap; (B) Total harmonic distortion versus tap (RT56).

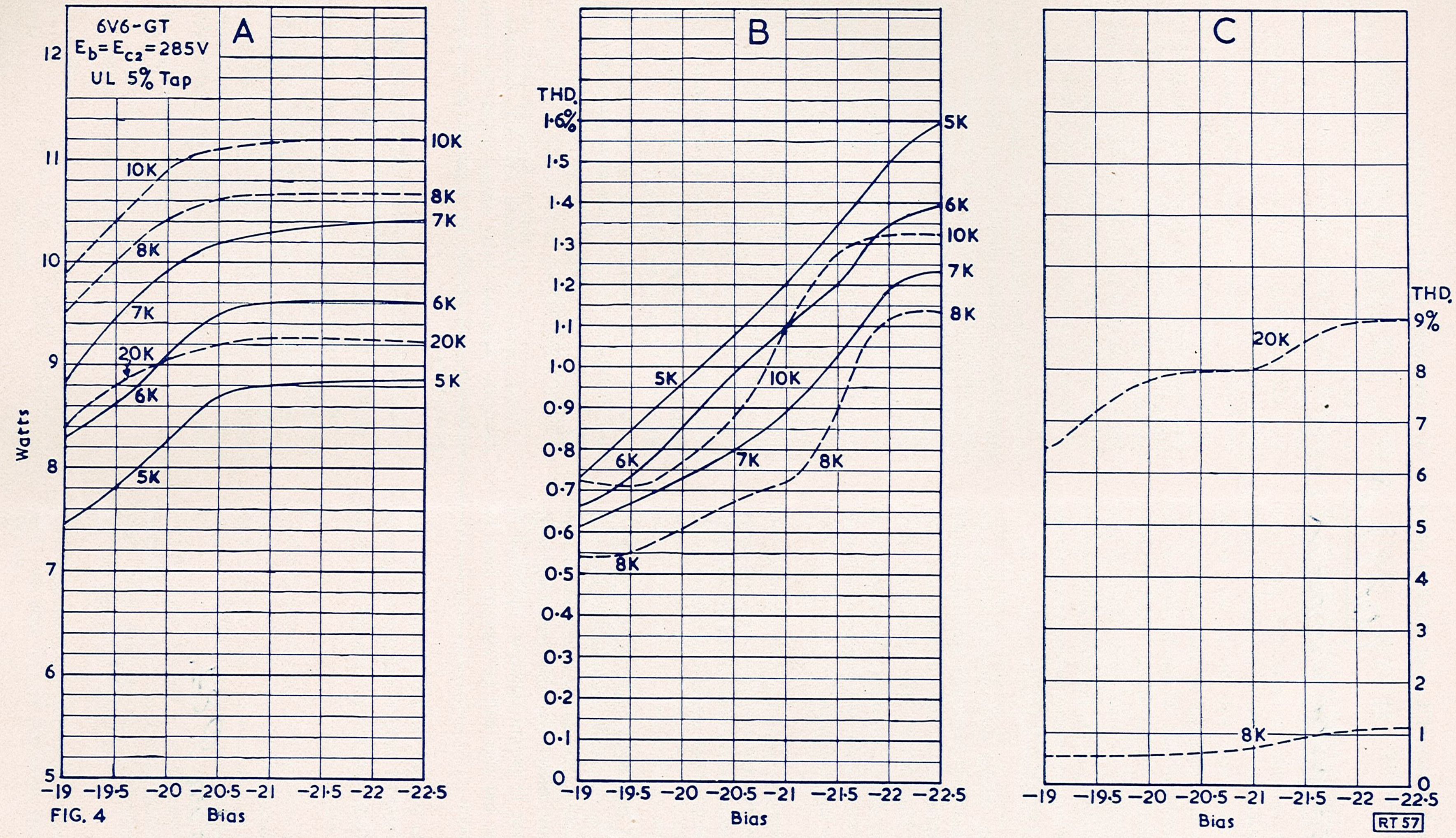

FIG. 4

Fig. 4. 6V6-GT push-pull UL operation, $E_b = E_{c2} = 285V$, tap 5%, peak grid voltage equals bias: (A) Power output versus bias; (B) Total harmonic distortion versus bias; (C) Total harmonic distortion versus bias for 20,000 ohm load, with 8,000 ohm load for comparison. Note that THD scales in B and C differ (RT57).

2. Comparison between UL and divided plate and cathode loading.

(Ref. 1 & 2)

Divided plate and cathode loading is used in the QUAD II and other amplifiers. The transformer half-primary is in two sections, part connected from cathode to earth and part from plate to B+, while the screen is bypassed to earth. It is obvious that there will be an a.c. voltage between screens and cathodes in the same way as the UL amplifier — in fact, it can be kept to the same value if desired. Thus divided loading operation has a performance in some ways similar to that of the UL operation. The principal difference is that divided loading has additional negative voltage feedback in the cathode-grid circuit.

For equivalent results with divided loading the cathode winding should have the same proportion of the total impedance as for UL operation. For example, with type KT66 this impedance ratio should be about 20%. In effect, the transformers will differ only in that the winding for the former is broken at the tapping point.

The choice between UL and divided loading will be influenced by the reduced gain with the latter. In most other respects the two methods are quite similar, and results for one apply very closely to the other.

The measured impedance ratio in the QUAD II amplifier is 3.5%, which is considerably less than the optimum (20%) found in our tests of type KT66. It seems likely that the preceding stage gain is insufficient to permit optimum operation of the output stage, but this will be checked later by direct measurement.

It is hoped to publish, at some future date, test results using a special output transformer permitting divided loads with a choice of several impedance ratios.

† Radiotronics, May, 1955.

Corrections from P. 84.

3. Comments on UL operation.

The UL amplifier is quite distinctive, neither a pentode nor a triode, but with its own marked characteristics. Its power output is definitely less than that of a pentode, as is clearly shown for type KT66 by Fig. 2 of Part 1 of this series†. This effect has often been obscured by the choice of conditions to give optimum performance for UL operation and then, without any other change except the connection of the screens on the transformer, measuring the power output for pentode operation. As shown in the earlier article, this would not give optimum pentode operation, so that the comparison is not a fair one. This remark applies particularly to the curve published by Hafler and Keroes.

Secondly, the reduction in distortion with UL operation is much greater than the reduction in gain. Taking Figs. 5A and B from Part 1 of this article, which are both for 5000 ohm loads, the ratio of gains from pentode to UL 5% operation is approximately 1.48 times (from the slopes of the tangents to the linearity characteristics). On the other hand the ratio of total harmonic distortion varies from approximately 1.4 times at low levels, to over 3.3 times at 18 watts, for the same power output in both cases. The same effect holds under all conditions which we have measured, for type 6V6-GT as well as KT66.

Thirdly, the optimum load resistance for UL operation, in all cases so far tested by us, is greater than that for pentode operation.

Fourthly, the shape of the plate characteristics of the valve is distinctly different from those of either triodes or pentodes.

These points all indicate that UL operation is distinctly different from either triode or pentode operation, and should be regarded as a separate phenomenon.

References.

1. W. N. Williamson and P. J. Walker, "Amplifiers and Superlatives", W.W. 58.9 (Sept., 1952), 357.
2. P. J. Walker (letter), "Ultra-linear Operation", W.W. 60.12 (Dec., 1954), 593.

PRE-AMPLIFIERS FOR USE WITH WILLIAMSON AMPLIFIERS

by F. Langford-Smith.

It is widely known that instability usually occurs when a high-gain pre-amplifier is connected to a Williamson amplifier, but the reason is not always fully understood. Williamson himself states that "a very carefully designed and necessarily expensive decoupling system is required if a high-gain pre-amplifier is to operate satisfactorily from the amplifier power supply. The cost of such decoupling is higher than that of a separate power supply unit producing, say, 350 volts at 20 mA, and therefore the use of a unit of this type is strongly recommended".

The instability, which occurs with a common power supply and conventional decoupling, is at a very low frequency and is accentuated by the use of maximum bass boost in the pre-amplifier — usually about 20 db down to well below 10 c/s. The main amplifier itself has about 6 db boost at 2 c/s, falling to 0 db at 10 c/s, and this adds to the overall bass boost, to which must be added the bass boost inherent in the equalizing. Even the Williamson pre-amplifier with a steep attenuation characteristic below 20 c/s (—30 db at 10 c/s) cannot be used with a common power supply, and other pre-amplifiers with less effective rumble filters are even more likely to strike trouble.

The tendency of the Williamson amplifier towards instability at low frequencies may be made worse by not adhering closely to the values of components on the circuit diagram. For example, increasing the value of one or more of the grid coupling condensers is likely to have this result.

On the other hand, any tendency towards instability will be much reduced by a small reduction in the feedback — say from 20 to 14 db. The doubling of the measured distortion (being inaudible in any case) is of less consequence than the other by-products of too small a stability margin.

The ineffectiveness of conventional filters at very low frequencies is indicated by the fact that the reactance of a 20 μF condenser is 800 ohms at 10 c/s, and 4000 ohms at 2 c/s.

There are two methods which have been used successfully to allow a pre-amplifier to operate satisfactorily with a Williamson main amplifier. The most common method is to use a separate power supply, and this is used by Williamson. The other is to use one or two voltage regulator tubes to improve the regulation of the plate supply voltage. One Radiotron type 0C3 and one type 0D3 in series will handle an output of about 255 volts from a source voltage of 320 or more. Alternatively, a single type 0D3 will handle an output of about 150 volts.

MEASURING SECOND HARMONIC DISTORTION WITH MILLIAMMETER

It is well known that second harmonic distortion in a power valve causes the d.c. plate current to rise from no signal to maximum signal. It is not so well known that this may be used as an accurate measurement of the distortion, at least with Class A triodes where the higher order harmonics are small.

It may readily be shown* that

$$H2 = 71 \, \Delta \, I_b \sqrt{(R_L/P_o)}$$
$$\text{or } \Delta \, I_b = 0.014 \, H_2 \sqrt{(P_o/R_L)}$$

Where H2 = percentage second harmonic.
I_b = rise in plate current in amperes
R_L = load resistance in ohms
and P_o = power output in watts at max. signal.

For example, type 2A3 operates as a single-ended amplifier with R_L = 2500 ohms, P_o = 3.5 watts and H2 = 5%. Using eqn. (1) we find that ΔI_b = 0.0026 ampere = 2.6 mA.

The same method may be applied to push-pull amplifiers to give the second harmonic distortion in each valve, although this is effectively balanced out by the push-pull operation.

* Using eqns. (3) and (10) on page 549, and eqn. (11) on page 551 of Radiotron Designer's Handbook, 4th ed.

REDUCTION IN TOTAL HARMONIC DISTORTION BY FEEDBACK

By F. LANGFORD-SMITH.

One of our readers, who was fully aware that negative feedback reduced any individual harmonic by the gain reduction factor, was not sure what happened to the total harmonic distortion and asked us to investigate.

The total harmonic distortion percentage is given approximately by:

$$TDH = \sqrt{(H_2\%)^2 + (H_3\%)^2 + \ldots\ldots}$$

If each harmonic is reduced by feedback to one tenth of its original value, then the total harmonic distortion with feedback is:

$$THD^1 = \sqrt{(H_2\% \div 10)^2 + (H_3\% \div 10)^2 + \ldots}$$

$$= (1/10)\sqrt{(H_2\%)^2 + (H_3\%)^2 + \ldots}$$

$$= (1/10)\ THD.$$

This indicates that the total harmonic distortion is reduced in the same proportion that each harmonic is reduced. Two assumptions were made in this simple treatment which may not always be maintained:—

1. That the gain of the amplifier is constant up to the highest significant harmonic, and

2. That the gain of the amplifier without feedback is constant from low levels to maximum output, and throughout the cycle at maximum output. This is far from true with power pentodes.

CALCULATING FEEDBACK RESISTORS FOR DIFFERENT VOICE COIL IMPEDANCES

It is quite a simple matter for anyone with a slide rule to calculate the correct series feedback resistor to use with any voice coil, when the published value is not suitable. The formula is:

$$R_{f2} = R_{f1} \sqrt{\frac{Z_2}{Z_1}} \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots (1)$$

where Z_1 = published voice coil impedance
Z_2 = required voice coil impedance
R_{f1} = published series feedback resistor
R_{f2} = required series feedback resistor.

Equation (1) is a close approximation only when the feedback resistor is considerably larger than the a.c. impedance between the point to which the feedback is applied, and earth.

A capacitor C_f shunted across R_f follows the inverse relationship:

$$C_{f2} = C_{f1} \sqrt{\frac{Z_1}{Z_2}} \quad \ldots\ldots\ldots\ldots\ldots (2)$$

If a slide-rule is not available, Table 1 may be used to determine the ratio $\sqrt{(Z_2/Z_1)}$ for use in eqn. (1). Its inverse may be used in equation (2).

TABLE 1. $\sqrt{(Z_2/Z_1)}$

Z_1 \ Z_2 =	2.0	3.5	6.0	8.0	10	12.5	15
Z_1 = 2.0	1.0	1.32	1.73	2.0	2.23	2.5	2.73
3.5	0.76	1.0	1.71	1.51	2.85	3.56	4.3
6.0	0.58	0.76	1.0	1.15	1.29	1.44	1.58
8.0	0.5	0.66	0.86	1.0	1.12	1.25	1.37
10	0.45	0.59	0.77	0.89	1.0	1.12	1.22
12.5	0.40	0.53	0.69	0.80	0.89	1.0	1.09
15	0.36	0.48	0.63	0.73	0.82	0.91	1.0

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