Radiotronics

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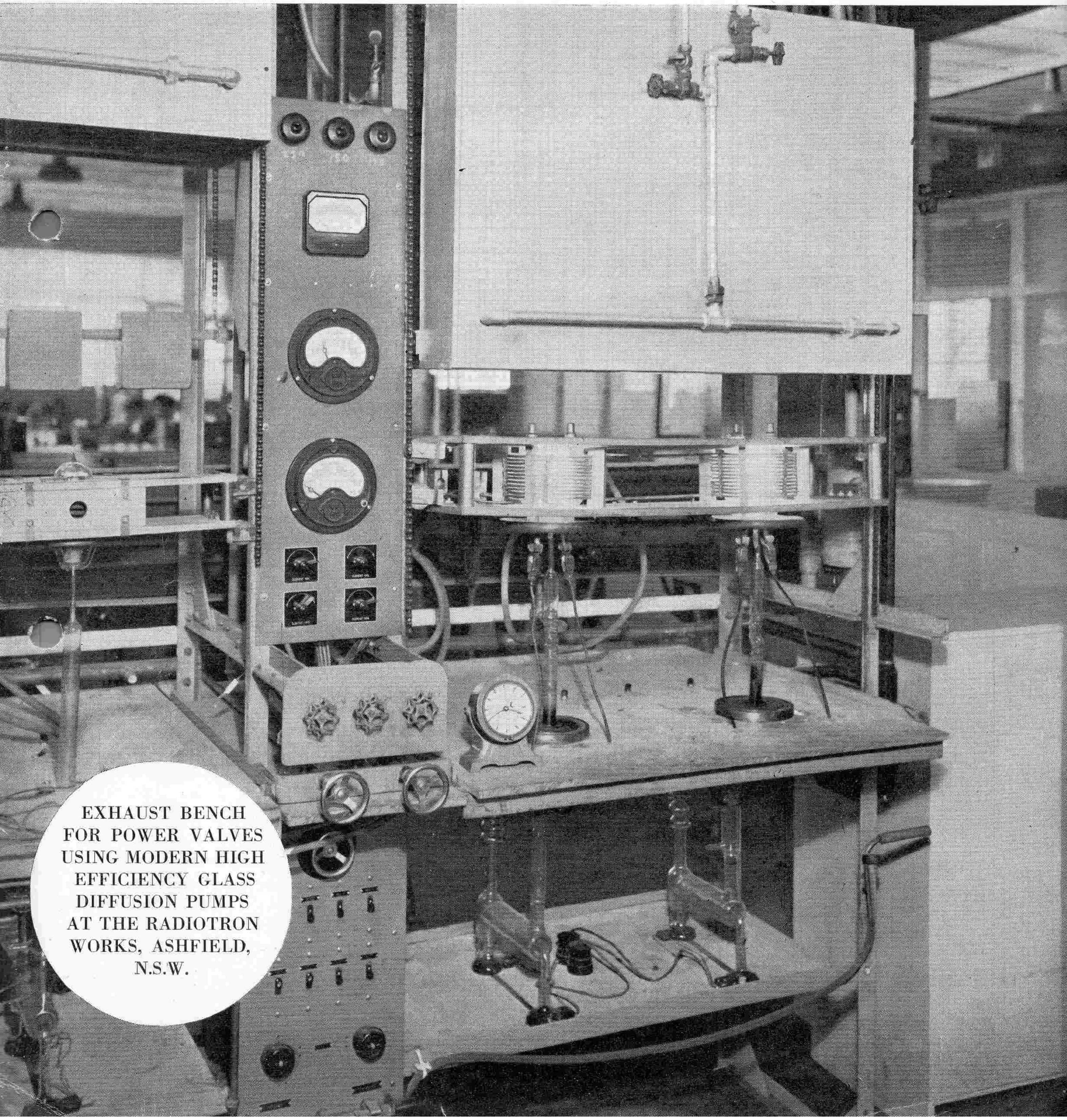

EXHAUST BENCH FOR POWER VALVES USING MODERN HIGH EFFICIENCY GLASS DIFFUSION PUMPS AT THE RADIOTRON WORKS, ASHFIELD, N.S.W.

RADIOTRONICS

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Technical Editor

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OUR COVER

The photograph reproduced on our cover shows an Exhaust Bench for Radiotron Power Valves, which exhausts the gases from power valves using modern high efficiency glass diffusion pumps.

In the centre of the photograph can plainly be seen two of the glass diffusion pumps in the process of exhausting two Radiotron type 833-A valves. At this stage the plates are being heated to over 1,000° centigrade by high frequency induction heating.

Radiotron 30 Watt Amplifier A513

By R. H. ASTON, A.M.I.R.E. (Aust.)

A very useful class of amplifier is one with an output of about thirty watts. This is quite adequate for almost any indoor work and, used with efficient speakers, is suitable for covering small areas out of doors.

Powers of this order can be economically obtained by the use of two type 807 valves in a class AB1 arrangement. There is much to recommend the 807 valve in this application. The output is well within its capabilities; the efficiency is high; the distortion is largely second harmonic which can be minimized by cancellation in a push-pull stage and the high power sensitivity can be utilized to provide feedback in an economical manner. Large stocks of type 807 are held, which are available at a special low price.

The logical driving source is the well tried phase inverter employing equal loads in the plate and cathode circuits. This arrangement is virtually distortionless and very much cheaper than a transformer capable of giving equivalent results.

The voltage amplifier stage employs a 6SJ7-GT with a high resistance plate load which results in a gain of more than 200 times, so that even with a gain reduction of 2.2 times due to feedback, the amplifier can be driven to full output with an input of less than 0.25 volt rms. This value of 0.25 volt is the order of the output usually available from the average magnetic pick-up and therefore forms a useful design basis for the input voltage of an amplifier.

Overall feedback is used to reduce distortion and output resistance. The arrangement is similar to that used in a 13 watt amplifier described in Radiotronics 112 where the pros and cons of this method were discussed at length.

To repeat briefly: there should be close coupling between the two halves of the output transformer primary, and reasonable precautions taken to avoid phase rotation within the feedback loop. This is because the number of coupling circuits involved, if badly designed, would permit sufficient phase rotation to cause the feedback to become regenerative at the outside limits of the useful frequency range.
As some phase rotation is unavoidable and practical transformers do not have perfect coupling between primary halves, a small capacitor (C_1 in

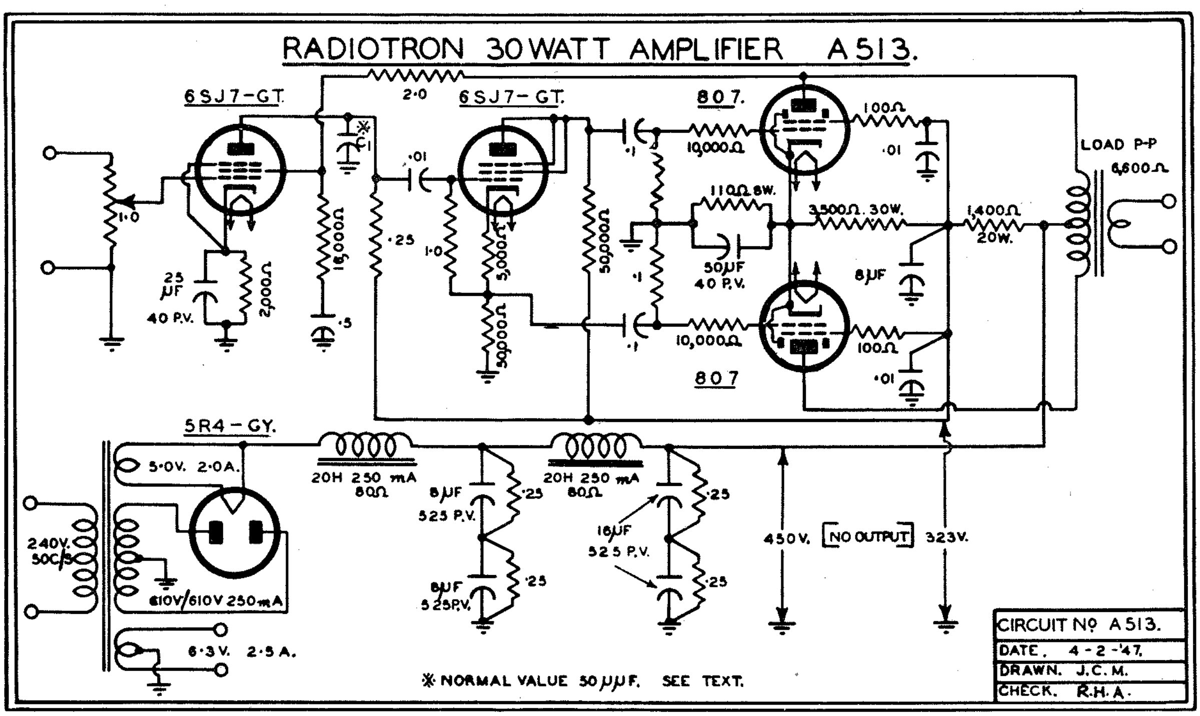


fig. 1) is used to bypass the higher frequencies at the plate of the 6SJ7-GT voltage amplifier. In our experimental model when C_1 was omitted a parasitic oscillation occurred as the amplifier was driven to full output, though it was quite stable for outputs up to 20 watts.

Including a 50 $\mu\mu$F capacitor C_1 was sufficient to make the amplifier completely stable. The attenuation of output at 10,000 c/s due to this shunting was only 1.5 db. If it is desired to reduce the high frequency response it should by done by increasing C_1 as necessary.

The presence of parasitic oscillations can best be detected with the aid of an oscillograph. If the input wave form is shown on the screen, parasitic oscillation can be observed as a widening of the trace line for part of the cycle—rather like a modulation envelope.

(1) PRACTICAL DETAILS

When the output capabilities of power valves are quoted for certain circuit conditions, this refers to the power developed across the prescribed load which is the load actually seen by the plate of the valve. With practical output transformers loaded on the secondary, the load seen by the valve may be quite different from the recommended value. Furthermore, the power available on the secondary may be as low as only 50 per cent. of the power input to the primary.

A fairly good transformer will have an efficiency of about 80 per cent.

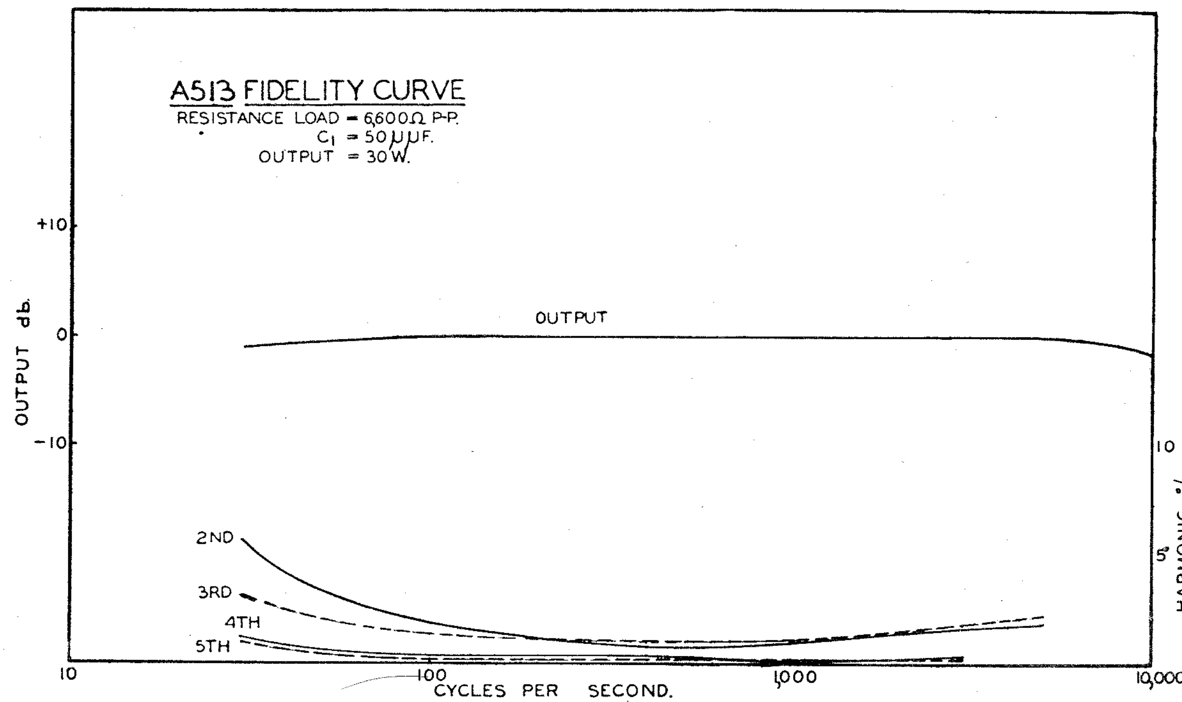
Therefore when measuring the output available from power valves it is necessary to use a resistance of the recommended value and of adequate wattage shunted across the primary of the output transformer. The secondary of the transformer is left open-circuited. The d.c plate current flows through the transformer primary in the normal way, but as the a.c. impedance of the transformer primary is assumed to be very high, practically the whole of the a.c. power is dissipated in the resistance. The voltage is measured across this resistance and the power calculated from $W = E^2/R$.

The overall frequency response is normally measured across a resistance placed in the secondary circuit of the output transformer. The power level for this test may be of the order of one tenth of the full power capability of the amplifier.

These measurements, made with resistance loads, form a basis on which amplifiers may be compared. Although the performance with speaker load will be rather different, any attempt to make measurements with a speaker load is complicated by the varying, and generally unknown, nature of the load.

The output transformer used for our model was a stock line, made by Ferguson, which was reasonably efficient and had sufficient inductance to maintain level output to 30 c/s and small enough leakage inductance to allow level output to well above 10,000 c/s, with a resistance load on the transformer secondary.

The bias arrangement for the power stage is a combination of fixed and self bias which results in



fairly uniform power consumption for varying outputs, and yet allows the use of a single power supply. The heavy bleed current through the screen voltage divider also passes through the cathode bias resistor thus making the bias voltage developed across it not wholly dependent on cathode current. The resistors comprising this voltage divider and bias resistor should be within five per cent. of the prescribed values. The power supply is then a simple one, although of fairly large physical dimensions. The total filter capacity shown in fig. 1 can be increased if a reduction in hum is considered necessary—as for low level listening in the home.

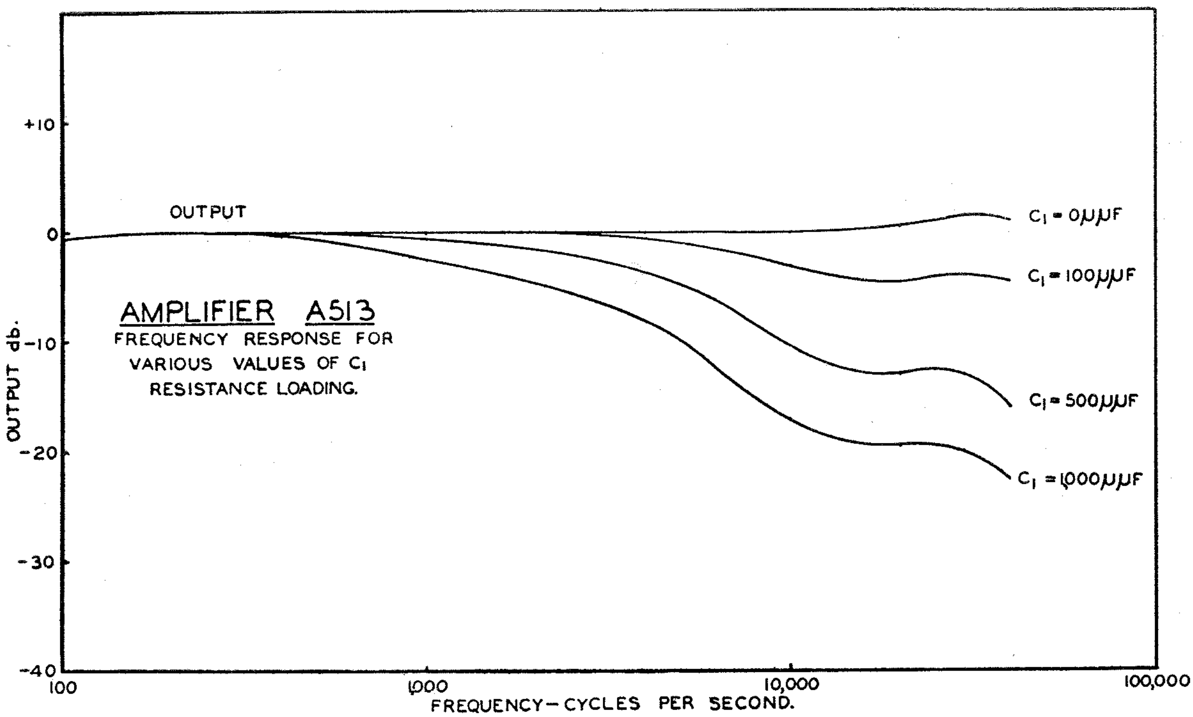
However, the hum level of 0.03 microwatt or 90 db below full output should be satisfactory for most applications.

this amplifier allow a good safety margin for line voltage fluctuation, and various valve tolerances.

The actual 807 valves used for the measurements were fairly typical; they were selected from a small batch as having reasonably similar plate currents. No attempt was made to achieve very close matching, so that the published results may be duplicated fairly easily.

(2) PHASE INVERTER

Some criticism has been directed at the type of phase inverter used; it has been claimed that the outputs available across equal plate and cathode loads are not equal. In order to investigate this point, a very careful check was made. As the loads consist of the d.c. loads in parallel with the combination of coupling condenser and grid leak in



With the precautions taken in this amplifier to ensure stable operation, no evidence of instability of the 807's was observed for any condition of output.

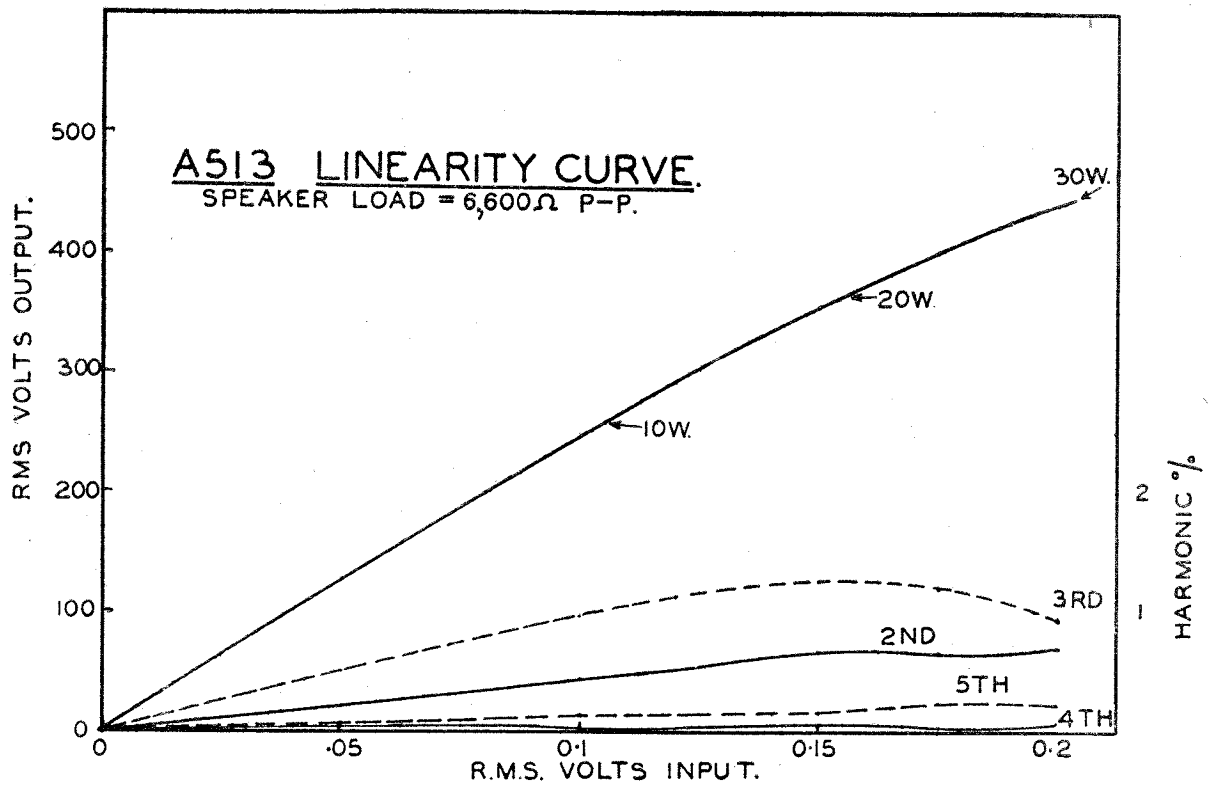
The only unstable condition observed was due to phase rotation in the feedback loop—which is corrected by including C_1 as previously discussed.

The screen dissipation of the 807's should be carefully checked at full output as it is quite possible to exceed the safe working region if there are quite small departures made from the recommended values of screen potentials and plate load. The screen dissipation is found from the product of screen current in amps. and the screen-to-cathode potential in volts. The conditions of operation for

series, it is necessary to carefully select all four of these components for symmetry. To eliminate the effect of grid current, the 807's were removed from the sockets. The effect of the loading of the valve voltmeter was checked and found to be negligible.

Then the respective output voltages were compared for several inputs ranging from zero to maximum, and found to be equal within 1.5 per cent. over the whole range of inputs and frequencies from 50 c/s to 10,000 c/s. However, one effect noticed is worth mentioning as it might account for some difficulty in obtaining apparent balance. The B.F.O. used has a small component of even harmonic output so that the positive and negative peaks were of unequal amplitude; as the valve voltmeter measures rectified half cycles, the readings of the two outputs were different because in one case the voltmeter was rectifying the larger peaks and in the other (with 180° phase difference) rectifying the smaller peaks.

The error was eliminated by reversing the phase of the input as the valve voltmeter was moved from one output to the other.



(3) SUMMARY OF TEST RESULTS

Power Stage and Power Supply (values are for both 807 valves)	No Output	30W Output
Voltage of secondary of power transformer	610 Vrms	608 Vrms
Voltage 5R4-GY filament to ground	488 volts	480 volts
B+ voltage output of power supply	450 volts	440 volts
Voltage plate to cathode 807's	412 volts	400 volts
Voltage screen to cathode 807's	300 volts	273 volts
Voltage cathode to earth	22.8 volts	25 volts
Total B+ current (including bleed)	219 mA	240 mA
Plate current 807's	128 mA	140 mA
Screen current 807's	5.6 mA	20.5 mA
Dissipation plates	52.8 watts	26.0 watts
Dissipation screens	1.68 watts	5.62watts
Load resistance plate to plate	6,600 ohms	6,600 ohms
Peak a-f grid-to-grid voltage	—	47 volts

Overall Measurements

Input voltage for 30W output	0.2 volt rms
Gains at 400 c/s (3W output):	
Voltage amplifier stage	210 times
Phase inverter stage (grid to grid)	1.54 times
Power stage	16.15 times
Overall with feedback	2,470 times
Overall without feedback	5,460 times
Gain reduction factor due to feed-back	2.21 times
Gain reduction due to feedback	3.44 db.
Output resistance (R_o) plate-to-plate	4,880 ohms
Damping factor R_L/R_o	1.355
Hum output (no signal)	.03μW (14.5 milli-volts across 6,600 ohms)

Radiotron 45 Watt Amplifier A514

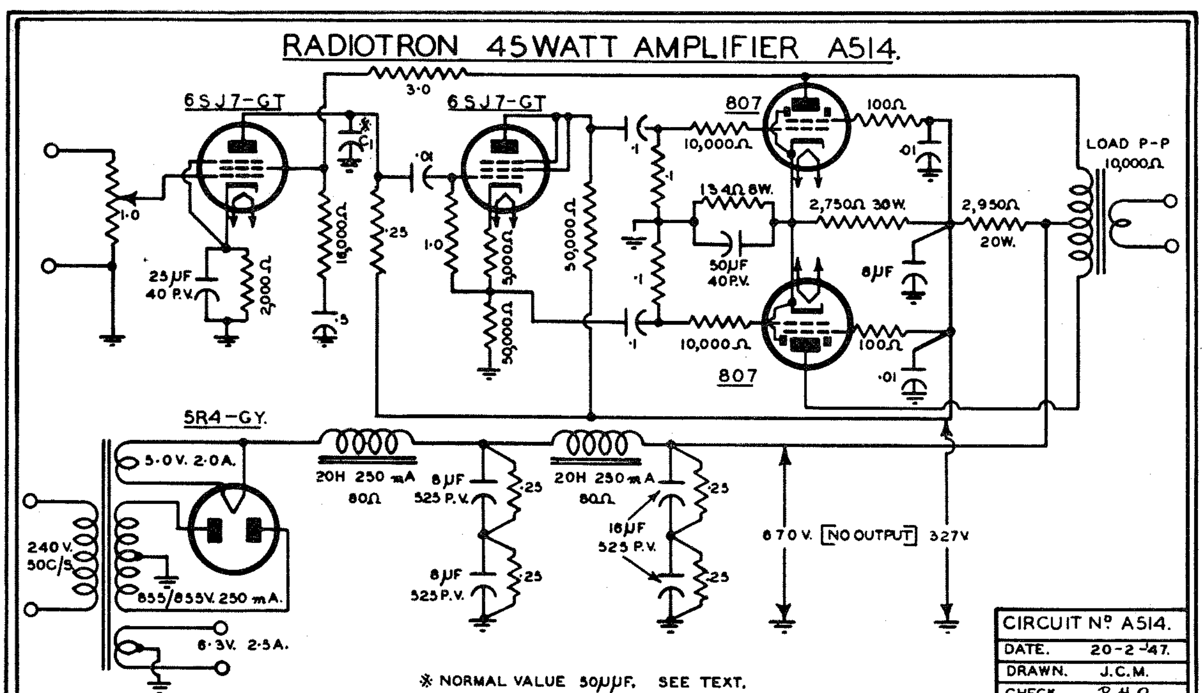
By R. H. ASTON, A.M.I.R.E. (Aust.)

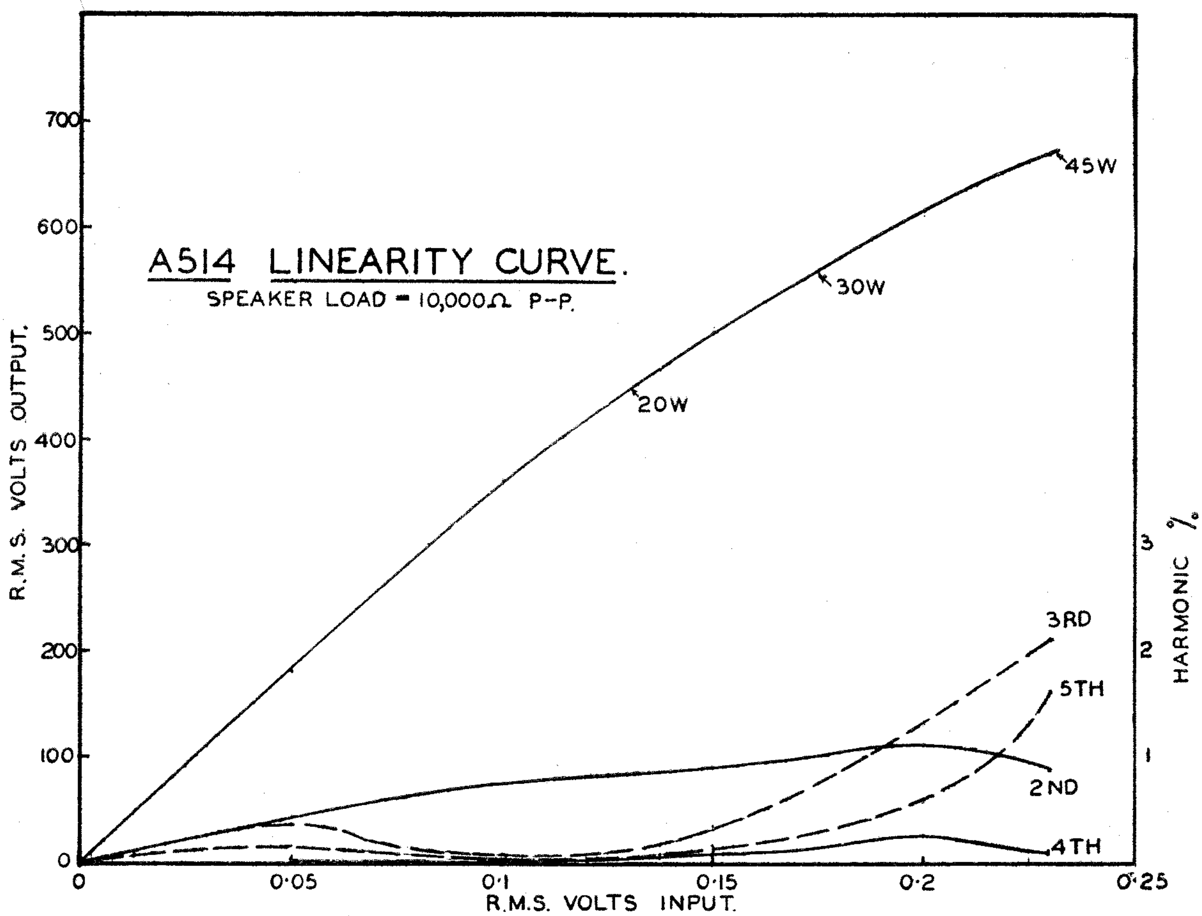
Only a few minor circuit changes including an increase in 807 plate voltage from 400 volts to 600 volts are necessary to convert the 30 watt amplifier A513 to one capable of an output of 45 watts. This new amplifier has been designated A514 and is included here with amplifier A513 as most of the performance characteristics are common to both. Only the modifications necessary to circuit A513 are described in this section.

The principal change necessary is in the power transformer which is now required to supply a secondary voltage of 845 volts rms at 250 milliamps. The output transformer should present a plate to plate load of 10,000 ohms and be capable of working efficiently with 45 watts of power input.

The voltage dropping resistor for the screen of the voltage amplifier 6SJ7-GT is now 3 megohms to provide approximately the same screen voltage with the increased supply voltage. The feedback factor remains practically unaltered because of the increased gain in the power stage.

The voltage divider resistance is increased because of the higher B+ supply, and the need to raise the bias to 27 volts for the no-output condition so that the power dissipated in the plates of the 807's is not excessive. This plate dissipation should be carefully measured for the no-output condition, because with 600 volts on the plate, it is very easy to exceed the safe maximum. From a measurement of plate volts and current, the dissipation can be calculated as W=EI. In the case where there is audio output from the amplifier, the audio power is subtracted from the input d.c. power supplied to the plate; so that although plate current is higher with audio output it is during stand-by conditions that the danger of excessive plate dissipation exists. However, with the higher plate voltage there is little danger of exceeding the safe screen dissipation which limits the power output available in the 400 volt plate condition.





SUMMARY OF TEST RESULTS

Power Stage and Power Supply (values are for both 807 valves)	No Output	45W Output
Voltage of secondary of power transformer	855 Vrms	845 Vrms
Voltage 5R4-GY filament to ground	700 volts	690 volts
B+ voltage output of power supply	670 volts	648 volts
Voltage plate to cathode 807's	620 volts	600 volts
Voltage screen to cathode 807's	300 volts	272 volts
Voltage cathode to earth 807's	27 volts	31.5 volts
Total B+ current (including bleed)	208 mA	241 mA
Plate current 807's	97 mA	130 mA
Screen current 807's	3.8 mA	15 mA
Dissipation plates	60 watts	33 watts
Dissipation screens	1.14 watts	4.08 watts
Load resistance plate to plate	10,000 ohms	10,000 ohms
Peak a-f grid-to-grid voltage	—	59.4 volts

Overall Measurements

Input voltage for 45W output	0.24 volt rms
Gains at 400 c/s (5W output):	
Voltage amplifier stage	210 times
Phase inverter stage	1.54 times
Power stage	21.15 times
Overall with feedback	3,500 times
Overall without feedback	7,460 times
Gain reduction factor due to feedback	2.13 times
Gain reduction due to feedback	6.56 db.
Output resistance (R_o) plate-to-plate	10,800 ohms
Damping factor R_L/R_o	0.93
Hum output (no signal)	0.012μW

The Design of a High Fidelity Amplifier

By F. LANGFORD-SMITH, B.Sc., B.E.

(1) THE POWER VALVE AND THE LOUDSPEAKER

In the design of any amplifier, it is highly desirable to commence with the loudspeaker and then to follow with the power stage and so backwards through the circuit. This is particularly so when dealing with a high fidelity amplifier in which special problems arise which must be solved before good results can be achieved.

With a triode power valve or, of course, push-pull triodes, the problem is comparatively simple and it is only necessary to select a loudspeaker having a nominal impedance equal to the normal resistive load required by the power valves. The increase in impedance, which occurs at high audio frequencies and at the speaker bass resonant frequency, merely causes some loss of output at these frequencies but this is not accompanied by any increase in distortion. In addition, the low plate resistance of the output valves provides a reasonable degree of damping on the loudspeaker.

With pentode or beam power amplifier valves, the position is entirely different. All such valve types have a critical load resistance whether used as single valves under Class A conditions or in push-pull under Class AB1.

Any variation of the load from the optimum results in very severe distortion while the valves are operated with maximum grid excitation. This can only be overcome by reducing the grid excitation and so reducing the maximum power output which the valve is capable of handling.

All this is brought about by the extremely wide variations in impedance presented by the loudspeaker, so that the most logical approach is to see whether something cannot be done to restrict the variations in loudspeaker impedance with frequency.

Loudspeaker Characteristics

The impedance versus frequency characteristics of a typical loudspeaker have been widely publicised.* The impedance remains approximately constant from a frequency of about twice that of the bass resonant frequency to about 1000 c/s; at higher frequencies it rises steadily until it reaches a value about 6 or 8 times that at 400 c/s. At the bass resonant frequency, it rises to approximately the same level, but only over a narrow band of frequencies—the shape resembling that of a selectivity curve which, in fact, it is.

The problem is, therefore, how to bring down to an approximately common level, both the upper frequency impedance characteristic and the bass resonant peak.

Most of the energy in musical sounds is in the middle and low frequencies, and the percentage of the total power above 1000 c/s is comparatively small. It is only under extremely abnormal conditions that the power in these higher frequencies approaches the maximum. It is possible, therefore, either to neglect the impedance rise altogether (as may be done if negative voltage feed-back is used) or to shunt the loudspeaker with a resistor-capacitor network to provide almost constant impedance at higher frequencies. The problem of the bass resonant peak is much more difficult and it is not generally considered advisable to shunt this by means of a filter network, since the frequency of maximum impendance is critical and subject to drift. There would, therefore, be a distinct likelihood of the filter becoming mis-tuned from the bass resonance.

An alternative method, which is sometimes practicable, is to connect a very heavy shunt resistor across the loudspeaker circuit so as to be effective at all frequencies. This has the effect of reducing all rises in impedance, but they can only be made negligible by arranging the shunt resistor so as to carry the greater part of the power output. It is, therefore, very inefficient when judged on the basis of power input to the final stage and effective power output from the loudspeaker.

None of the methods so far outlined is a satisfactory solution if the power from the loudspeaker is required to approach that of the maximum power delivered by the valve. There is, fortunately, an alternative method which has proved extremely effective in practice and relatively simple to apply. This is variously known as the "bass reflex" or "vented baffle" loudspeaker, which may be applied to any ordinary dynamic speaker. It has very little effect on the high frequency performance of the loudspeaker, but when correctly adjusted has two very beneficial effects on the bass region.

Firstly, it imposes considerable acoustic damping on the loudspeaker over a fairly wide range of frequencies below about twice the bass resonant frequency; secondly, it has the effect of replacing the very sharp impedance/frequency characteristic with

* Radiotron Designer's handbook, 3rd edition, page 20, Figure 1.

one having two humps situated one above and the other below the bass resonant frequency. Each of these two humps is considerably lower in impedance than the bass resonant impedance of the loudspeaker on a flat baffle.

By the use of such a speaker baffle, it is thus possible to achieve a considerable improvement in restricting the variation of impedance with frequency over the range up to 500 c/s. In one practical case, it was found that an initial variation between maximum and minimum impedance over this frequency range was reduced from 10:1 on a flat baffle to the more satisfactory figure of 5.8:1 on the bass reflex baffle. Some further improvement is, however, still desirable.

Use of Two Bass Reflex Speakers

Since it is possible to achieve two peaks with a single bass reflex speaker, it seemed to be worthwhile exploring the possibility of using two such speakers with the frequencies of the resonant peaks staggered so that, to a large extent, the impedance peaks of one would fill in the valleys of the other. After considerable investigation and experiment this was found to be practicable and two such bass reflex speakers constructed. It is hoped to publish, at some later date, the details of the design of these speakers, but in this article it is necessary to restrict consideration to those features affecting the general performance.

Curve 1 of figure 1 shows the impedance versus frequency, in free air, of a typical 12″ loudsqeaker of reasonably good general performance.

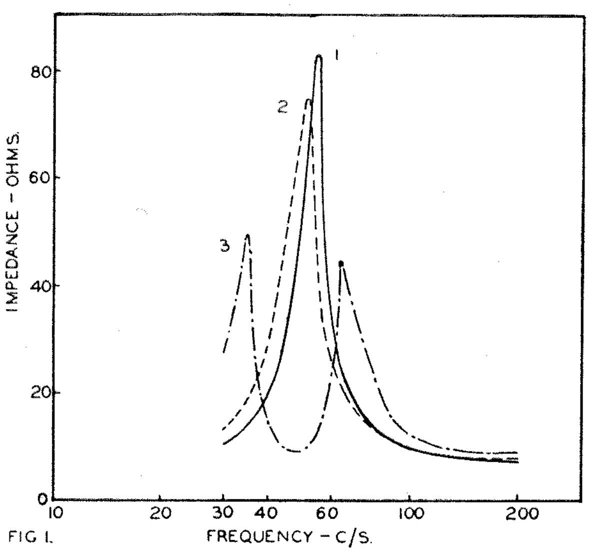


FIG I.

Fig 1. Curve 1 shows the impedance versus frequency characteristic of a 12″ loudspeaker in free air. Curve 2 shows the same on a flat baffle. Curve 3 shows the characteristic of the same loudspeaker in a bass reflex baffle.

Curve 2 on the same figure shows the impedance/frequency characteristic of the same loudspeaker operating on a flat baffle. It will be seen that the bass resonant frequency is very slightly reduced and that the rise of impedance is slightly reduced.

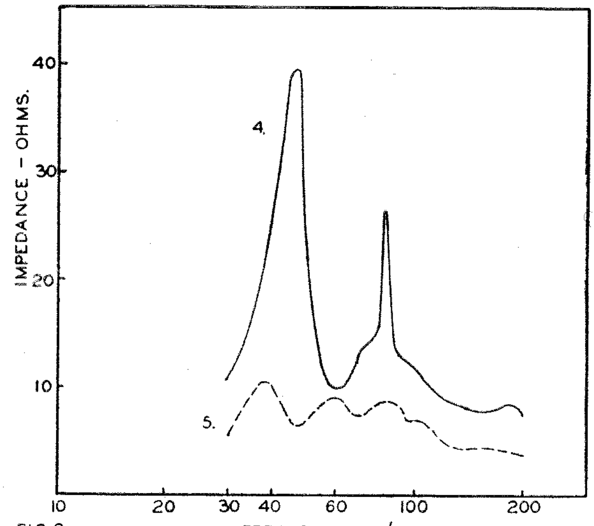


FIG. 2.

Fig. 2. Curve 4 shows the impedance versus frequency characteristic of a second 12″ loudspeaker, having a different bass resonant frequency, in a modified bass reflex baffle. Curve 5 shows the impedance of both bass reflex baffled speakers connected in parallel. The variation in impedance has been much reduced.

Curve 3 on the same figure shows the same loudspeaker in a bass reflex baffle, giving two impedance peaks of 5.8:1 and 5.3:1 as compared with the flat baffle.

Curve 4 on figure 2 shows the impedance/frequency characteristic of a second loudspeaker having a different bass resonant frequency and a modified bass reflex baffle.

Curve 5 on figure 2 shows the impedance/frequency characteristic of these two bass reflex baffled speakers connected in parallel. It will be seen that the variation has been reduced to a maximum of 2.8:1 while the variations over the greater part of the extreme bass region are comparatively small.

It can, therefore, been seen that a successful result has been achieved in reducing the impedance variation from 10:1 on the flat baffle to 2.8:1 on the two bass reflex units with staggered frequencies. This permits the design of an amplifier using beam power valves with considerably improved performance.

The Effect of Load Impedance

The effect of increasing load impedance with a beam power amplifier valve is illustrated in figure 3, in which the normal load resistance is 2,500 ohms. This normal loadline cuts the zero bias characteristic approximately at the knee of the curve. Increasing

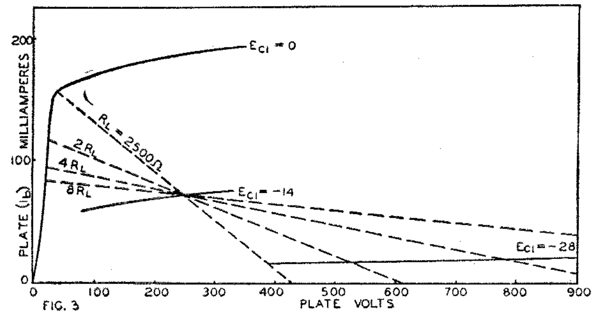


Fig. 3. Plate characteristics of type 807 or 6L6-G valve with normal load resistance (RL = 2500 ohms) and higher load resistances 2RL, 4RL, 8RL, without allowing for shifting loadlines caused by rectification.

values of load resistance of twice, 4 times and 8 times normal, are shown on the same figure. It will be seen how unsymmetrical and distorted the performance would be if full grid excitation is maintained. In practice, owing to the high degree of non-linearity, there is a rectification effect which results in the loadline shifting its position either up or down.

Figure 4 has been drawn to show the corrected loadlines, after allowing for the shift caused by rectification. Just as with figure 3, there is considerable distortion which is of a type particularly distasteful to listen to.

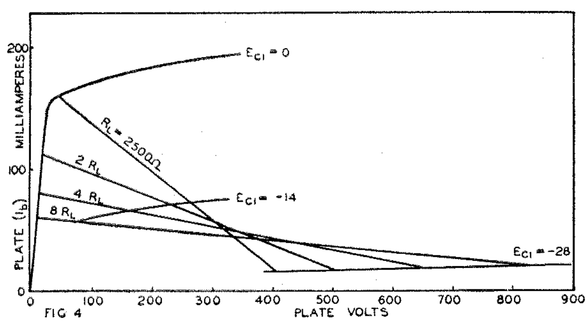


Fig. 4. is similar to Figure 3, except that allowance has been made for shifting loadlines due to rectification.

The wave form of the output current which occurs with full grid excitation with sine wave input, with a load res istance four times normal, is shown in figure 5. The flattening of one half of the cycle is very severe and results, not only in severe harmonic distortion, but also in the production of prominent inharmonic combination tones. Reproduction under these conditions is most distasteful to any listener.

There are two things which could be done to reduce this distortion. The first is to reduce the grid excitation until the distortion becomes sufficiently low at the highest impedance which the speaker develops at any frequency. The second is to reduce the load impedance at 400 c/s so that the loadline for the maximum impedance of the speaker does not cut so much below the knee of the curve.

The best compromise seems to be a combination of both these two methods, that is by somewhat reducing the grid excitation and simultaneously reducing the nominal load applied by the loudspeaker to the power valve.

Speaker Damping

One of the most important characteristics of loudspeakers is their damping. Horn speakers have a high degree of damping, down to a certain minimum frequency, and thus give very satisfactory performance. In most cases their size and cost preclude them

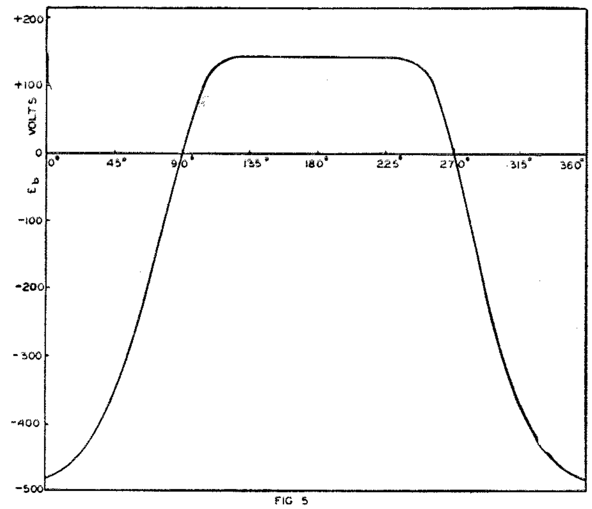


Fig. 5. shows the flattening of the peak of the output waveform which occurs with full grid excitation and sine- wave input, and a load resistance four times normal.

from being used in the home of the ordinary listener.

Ordinary dynamic speakers suffer from lack of sufficient acoustic damping and it is, therefore, desirable to introduce further damping by electrical

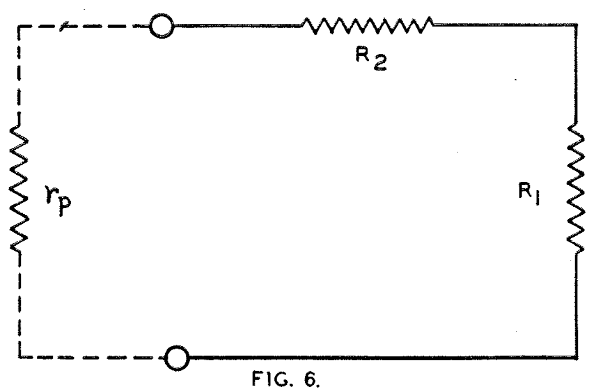


Fig. 6. The equivalent circuit of a loudspeaker to demonstrate damping by the plate resistance (r_p) of the valve. R_1 is the resistance of a "perfect reproducer" in series with the loss resistance R_2.

means. This is usually provided by the plate resistance of the output valve in parallel with any shunt resistor that may be used. It is sometimes thought that if the plate resistance could be reduced to zero, the damping on the loudspeaker would be very heavy, but this is not the case. As will be shown in a moment, even the use of cathode loading and the consequent reduction of effective plate resistance to a few hundred ohms does not achieve the desired effect.

A dynamic loudspeaker is inherently an inefficient device; an efficiency of approximately 4% is all that is normally achieved over the greater part of the frequency range. This remark does not apply to horn type loudspeakers which may have efficiencies above 40%.

As with any other component having losses, the equivalent circuit diagram may be drawn (Figure 6) in the form of a "perfect" reproducer (R_1) in series with a loss resistance (R_2), the total impedance presented to the output valve being, therefore, $R_1 + R_2$. The damping effect of the power valve is provided in the figure by the resistor r_p. It will be seen that even if r_p is reduced to zero the damping on the "perfect" reproducer is still R_2, the loss resistance. For an efficiency of 4%, R_2 would be 96 ohms as against $R_1 = 4$ ohms on an arbitrary basis. Thus the best "damping factor"† which can be achieved is 4/96. At the bass resonant frequency, however, R_1 becomes considerably greater and may (in the same arbitrary case) rise to 7×4 or 28 ohms, thus giving a damping factor, at this frequency, of 28/72. Even this is extremely limited, so that the only conclusion which can be drawn is that the maximum effective damping of a dynamic loudspeaker by the plate resistance of a valve is extremely limited.

Acoustic damping on the other hand, acts directly on the "perfect reproducer" so that it is all usefully applied. We, therefore, need to search for some method of applying acoustic damping, particularly at the bass frequencies, so as to achieve satisfactory performance.

Fortunately, this result is provided by the bass reflex baffle, and is even better with two such baffles at staggered frequencies. This has the effect of reducing the forward and backward movement of the cone and thereby reducing the speaker non-linear distortion which occurs when the voice-coil moves out of the uniform flux area. In other words, for a given speaker distortion, this kind of baffle increases the effective power which a given speaker will handle under wide-range conditions. Moreover, the damping of the speaker is more nearly dead-beat at bass frequencies, so that the speaker does not continue emitting sound when the source of such sound ceases. It therefore gives a better reproduction of transient sounds.

Sufficient has been said to indicate the advantages of a bass reflex baffle over a flat baffle, and particularly over the type of baffle formed by the average radio receiver cabinet. Its size does not require to be excessive, a total enclosed area of about eight cubic feet or even less being sufficient. The enclosed back has advantages, not only in excluding dirt, but also preventing unwanted reflections from the wall at the back of the set.

If a single bass reflex speaker is used in an ordinary room, one very satisfactory position is in the corner so that the two walls and floor form a sort of horn for concentrating the sound in the direction where it is required. The shape of the bass reflex enclosure is immaterial, and it may be made in the form of an angle so as to fit right in the corner of a room. Alternatively, it could be mounted in the corner between the ceiling and two walls, with the speaker pointing downwards to approximately the centre of the floor. This has the advantage of reducing the floor space occupied by the cabinet.

If two bass reflex speakers are used, their relative positions will have an effect on the impedance characteristics of the two in parallel. Tests carried out in our laboratory have indicated that two speakers placed closely together and connected in parallel give a more satisfactory impedance characteristic than when placed 15 feet apart. In any actual case, it is desirable for impedance tests to be taken with the speakers in position and correctly phased, and the final location should be determined so as to give the flattest impedance characteristic.

If the impedance characteristic permits the two speakers to be placed some distance apart, the writer prefers this to close spacing, since it provides some semblance to a third dimension. It might be possible to have one speaker in each of two adjacent corners of a room or even at opposite ends of a room. This spacing of the speakers is not recommended unless a check of the impedance characteristic has been made and shown to be satisfactory.

In later articles in this series, we hope to describe the design of a high fidelity amplifier to operate with these loudspeakers and give extremely good results. We wish to emphasize that this amplifier will not give anything approaching equivalent results with an ordinary dynamic type of speaker on a flat baffle or in a conventional radio receiver cabinet.

† Radiotron Designer's handbook, pages 15 and 22.

R.C.A. Application Note No. 116

PROPERTIES OF UNTUNED R-F AMPLIFIER STAGES

The design of an untuned (or resistance-coupled) r-f stage is facilitated by first reducing the usual tube and circuit relations to a group of characteristic curves which correlate transconductance, circuit values, frequency range, and stage gain. Such curves are directly useful as a guide in the preliminary steps of designing conventional r-f amplifiers, but are not intended for use in any special cases where the usual trial-and-error method of development may not be feasible.

Figures 1 and 2 show simplified r-f circuits for two conventional untuned r-f systems. In each case, the diagram shows two condensers of equal capacitance. One represents the output capacitance of the r-f amplifier tube plus wiring capacitance. The other represents the input capacitance of the following tube and the wiring capacitance. Making these two capacitances of equal value is a simplifying assumption which is permissible within the scope of this Note.

SIMPLIFIED CIRCUIT OF UNTUNED R-F STAGE WITHOUT PEAKING COIL

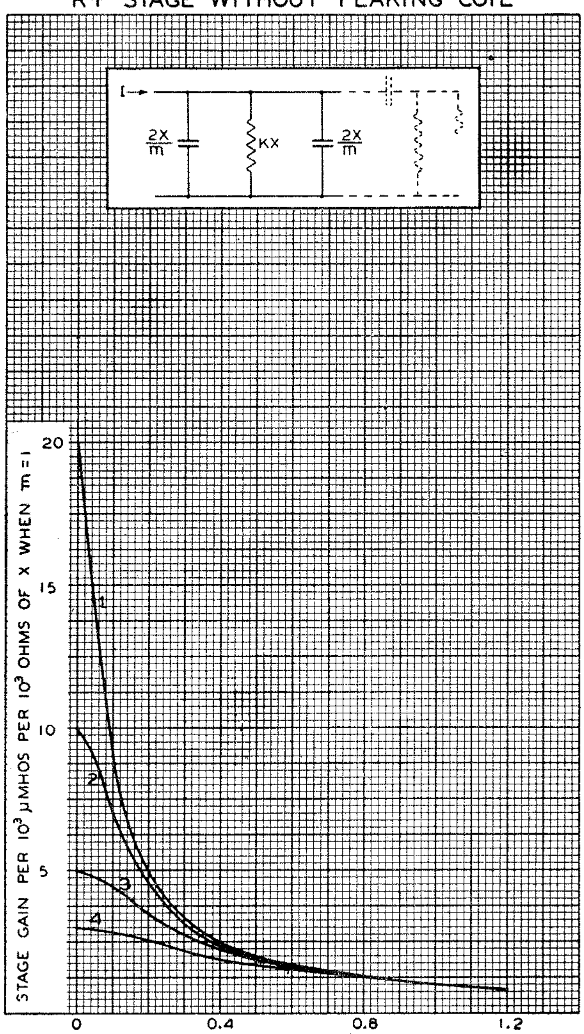


FIG. 1 m

SIMPLIFIED CIRCUIT OF UNTUNED R-F STAGE WITH PEAKING COIL

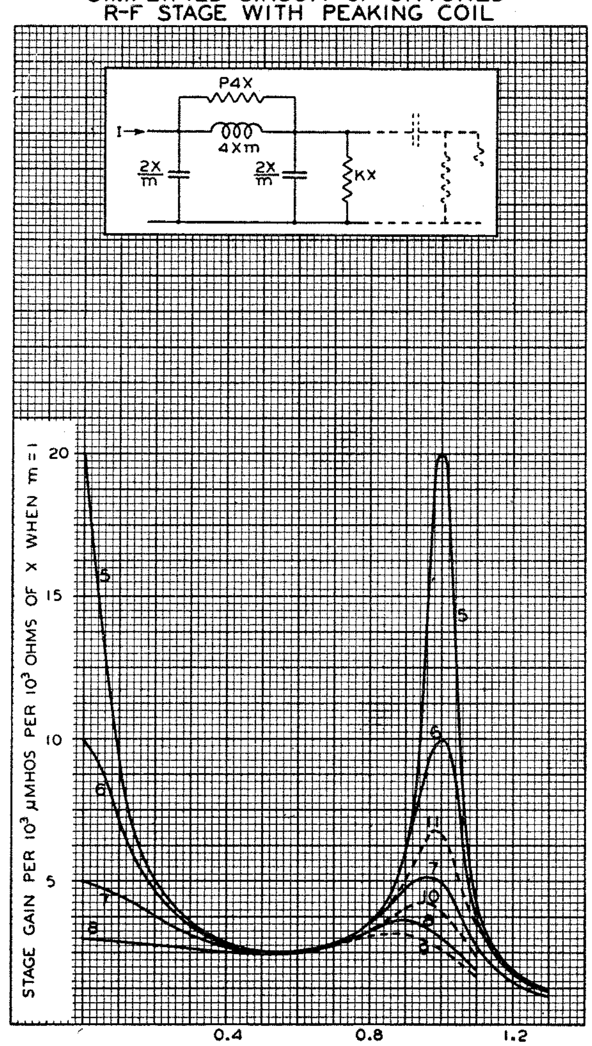


FIG. 2 m

In each diagram, a current I is shown entering the circuit. I is the signal current in the plate circuit of the r-f stage. It is simply the product of transconductance and the signal voltage impressed on the grid of the amplifier tube. This relation presumes that the plate impedance (R_p) of the r-f tube is in shunt with the network shown. However, in any likely untuned r-f stage this presumption may be disregarded without material error, because of the relatively low values of load impedance employed for such stages. Fig. 2 shows a resistor across the grid end of the network. Were this resistor to be placed across the plate end, as might be done in some instances, the characteristic curves would remain unchanged. When this resistor is placed as shown, it is understood to represent the effective a-c resistance of the load resistor and the usual shunt grid leak.

The circuit of Fig. 2 is formed from the circuit of Fig. 1 by adding an inductor. This inductor is commonly known as a "peaking coil"; its inductance accounts for the immediately obvious differences between the curves of Figs. 1 and 2. The peaking coil is, of course, an element of a low-pass filter, but an appropriate concept of its effect may be established by disregarding the resistors and then noting that series resonance obtains at a frequency where the reactance of the peaking coil is numerically equal to the reactance of the two condensers in series. At this frequency, I flows into a high value of parallel resonant impedance, and therefore produces a voltage rise at the input end of the system. The voltage rise affects the output end, but is limited by the power dissipated in either or both of the resistors shown.

Explanation of Terminology

The ordinates of the curves for both figures are labelled "stage gain per 10^3 micromhos, per 10^3 ohms of X when m equals one." The abscissas are labelled "m". On Fig. 2 m is the ratio of the impressed frequency to the frequency where the inductance of the peaking coil is resonant with the two capacitances in series; i.e., m is equal to unity at the nominal peak frequency of the system. For a nominal frequency of 16 megacycles, a value of m of 0.7 would correspond to 0.7 × 16, or 11.2 megacycles.

The reactance of the paralleled capacitances at the nominal peak frequency is X. If the total capacitance of the paralleled capacitances is 20 $\mu\mu$f, then, for a nominal peak frequency of 16 megacycles, X is 500 ohms.

K is the ratio of the effective shunt resistance to the total parallel capacitive reactance when m equals unity. For example, with m equal to unity at 16 megacycles, a total parallel capacitance of 20 $\mu\mu$f, X equal to 500 ohms, and a shunt resistance of 2,000 ohms, K is 2,000/500, or 4.

P is the ratio of the resistance of the damping resistor across the peaking coil to the reactance of the peaking coil when m is unity. Suppose that a total capacitance of 20 $\mu\mu$f is made up of two 10 $\mu\mu$f capacitances. These in series have a capacitance of 5 $\mu\mu$f and a reactance of 2,000 ohms at 16 megacycles. Therefore, the peaking coil should have a reactance of 2,000 ohms for resonance at 16 megacycles. A damping resistor 10,000 ohms across the peaking coil means that P is 10,000/2,000, or 5.

Suppose some other value of total capacitance, such as 18 or 25 $\mu\mu$f, is used, or some other value of nominal peak frequency, such as 15 or 18 megacycles. Then m is taken as equal to unity at whatever value of nominal peak frequency is chosen, and X is calculated only at that frequency.

The circuit elements of the diagrams are labelled with their impedances. Each capacitance is labelled 2X/m. This caption agrees with the use of X as the reactance of the total parallel capacitance. The peaking coil is labelled with its reactance 4Xm, a value which agrees with the condition of resonance that m is equal to unity.

The terminology of Fig. 1 is the same as that of Fig. 2, except that m on Fig. 1 does not designate a peak frequency. The value of m may be taken as unity for any chosen frequency. At this frequency, the ordinate approaches unity as KX is made large in comparison with X. X is evaluated only at the frequency where m is equal to unity, as in the case of Fig. 2.

An expression for the stage gain of the system shown in Fig. 1 is

$$\frac{(g_m)\ X\ K}{\sqrt{1 + (Km)^2}} \quad \ldots (1)$$

For Fig. 2, the stage gain is
(g_m) X K multiplied by

$$\sqrt{\frac{P^2 + m^2}{[P\,(1 - 2m^2) - Km^2]^2 + m^2\,[1 + KP\,(1 - m^2)]^2}} \quad \ldots (2)$$

These expressions have been used in plotting the curves shown in Figs. 1 and 2, and may be used for plotting other curves based on other combinations of K and P. The transconductance, (g_m), is in mhos. X is in ohms.

Stage-gain Characteristics

The curves of Fig 1, labelled 1, 2, 3, and 4, have been calculated for the values of K given below.

Curve	K
1	20
2	10
3	5
4	3

Suppose that m be equal to unity at 16 megacycles, and assume a typical value of 20 $\mu\mu$f as the total capacitance. In this case, X is 500 ohms. When K is 3, curve 4 shows the stage gain at one megacycle to be $3 \times 500/10^3 \times g_m$ (micromhos) $/10^3$. Hence, a transconductance of 2,000 micromhos results in a stage gain of 3; 4,000 micromhos gives a gain of 6. Moreover, it should be noted that curve 4 is essentially flat to an m of 0.1; i.e., the stage gain of this amplifier is uniform throughout the broadcast band. Consider curve 3, for K equal to 5. In this case, the assumed values of 20 $\mu\mu$f and 16 megacycles result in a stage gain of 4.6 at one megacycle when the transconductance is 2,000 micromhos, and of 9.2 for 4,000 micromhos.

When the other curves are considered in the same way, it becomes evident that variation of gain throughout the broadcast band is the chief difficulty in the use of high values of K in order to obtain high values of stage gain with low values of transconductance. However, since an untuned stage gain of more than 6 or 8 in the broadcast band is usually not desirable, and since a 2:1 variation of gain is usually tolerable, the development of an untuned r-f stage for the broadcast band is ordinarily a simple matter.

In many instances, an untuned r-f stage is called upon for amplification up to (say) 16 megacycles. When Fig. 1 is considered from this standpoint, it is noted that all of the curves merge as m is increased toward unity. Therefore K may be disregarded. On the basis of 20 $\mu\mu$f as the total capacitance and m equal to unity at 16 megacycles, the stage gain is only 1 at this frequency, for a transconductance of 2,000 micromhos, and 2 for 4,000 micromhos. It is because of these low values of stage gain that a peaking coil is desirable.

The peaking coil has no appreciable effect at m-values less than 0.4. Consequently, the portions of curves 5, 6, 7, and 8 on Fig. 2 for low values of m are respectively the same as the corresponding portions of curves 1, 2, 3, and 4 on Fig. 1, when K has the same values on both figures. In general, K is the only controllable circuit property which can determine the stage gain at low values of m. Even though a value of K is chosen to determine the stage gain in the broadcast band, K also affects the gain at the high-frequency portion of the desired frequency range. This state of affairs is shown by making P so large that KX is essentially the only dissipative element in Fig. 2. Then, curves 5, 6, 7, and 8 have the same ordinate values when m is unity as when m is zero. Thus, it is apparent that K limits the high-frequency gain. Also, high values of K give rise to an obviously excessive sharpness of response at the peak frequency. Were a high value of K to be used, this sharpness could be removed by using a suitably low or moderate value of P. (This expedient would decrease the ordinate for m equal to unity.) However, it is seldom necessary to depend on P for this result, because K itself is likely to be low enough to prevent excessive sharpness when it is low enough for suitable gain in the broadcast band. Curves 7 and 8, based on K-values of 5 and 3, respectively, show a worthwhile degree of gain at the high-frequency end of the range, without excessive sharpness. These and all of the full-lined curves of Fig. 2 depend solely on K, P being infinite; i.e., in this example, the peaking coil dissipates no power in or across itself. P and the resistance of the peaking coil itself, are quite important from the standpoint of high-frequency gain.

Effect of P and Resistance of Peaking Coil

Peaking coils, as ordinarily found in commercial receivers, are by no means "low loss" inductors. The "coil form" is often the resistor which provides the shunt damping, and it is usually wound with small wire. For such a coil, Q should not be assumed to be greater than about 20, even though P (in Fig. 2) may be so high that the coil resistance itself is the limiting factor. Accordingly, were P to be taken as 20, the true meaning in practice is that the series and shunt resistance are as though P is 20 when the resistor is connected across a lossless coil. The highest value assigned to P in this Note is 20.

Consider curves 7 and 8 of Fig. 2. These curves are of interest in practice, but their ordinates at the high-frequency end of the range are based on a peaking coil having no series resistance and no shunt damping. Consequently, it is important to determine the effect on these curves when P has its highest likely value of 20. The result is shown by the dotted curves 10 and 9. These correspond respectively to curves 7 and 8. Fortunately, when P is 20, curves 7 and 8 are not affected severely.

Curves 5 and 6, based on K-values of 20 and 10, respectively, are included for reference use with low transconductance; e.g., 1,000 micromhos, or less. From curve 6, which has a K-value of 10, the dotted curve 11 is obtained for P of 20. This same dotted curve 11 is also obtained from curve 5 for P of 10. Curve 11 may be regarded as roughly indicative of the maximum degree of sharpness which is acceptable in practice.

Stage Gain at Medium Values of m

All of the curves of Fig. 2 are substantially coincident between m-values of about 0.4 and 0.8. This condition means that any likely values of K and P have no material effect on the stage gain throughout the wide range of frequencies received by an all-wave receiver; for example, from roughly 6.5 to 12.5 megacycles when m is unity at 16 megacycles. The ordinates in Fig. 2 for these medium values of m have a minimum value of 2.4. This ordinate value is accepted as a natural property of the system. On this basis, the expected stage gain in practice can be increased only by increasing the transconductance, provided the circuit has its irreducible minimum value of capacitance. When this capacitance is 20 $\mu\mu$f, and when m is unity at 16 megacycles, X is 500 ohms, and the stage gain for medium values of m is taken as 2.4 $\times$ 500/10^3 $\times$ g_m (micromhos) 10^3, or, 1.2 per thousand micromhos of transconductance. Thus, 500 micromhos gives a stage gain of only 1.8, while 4,000 micromhos gives a stage gain of 4.8. A high-transconductance tube is required for worthwhile stage gain in the range of frequencies corresponding to medium values of m.

Effect of Transconductance on Shape Of Stage-gain Characteristic

It is evident from curves 8-9 that substantially uniform gain is readily obtainable when a high-transconductance tube is used to provide a worthwhile amount of gain. On the other hand, low-transconductance tubes lead to a consideration of curves 7-10, 6-11, etc. Although these curves lack the uniformity of gain shown by curves 8-9, they show a worthwhile amount of gain for the broadcast band and at the high-frequency end of the chosen range. In these cases, the low order of gain at medium values of m is a natural property of the system. This condition is sometimes regarded with enforced complacence, because signals in the broadcast band and in the high-frequency portion of the range are ordinarily more important than signals at frequencies corresponding to medium values of m. However, a gain of less than 1 at these less important frequencies is difficult to accept. For this reason, it is reiterated that regardless of the chosen curves of Fig. 2 the expected gain for medium values of m is found to be only 1.2 per thousand micromhos, when m is unity at 16 megacycles and when the total capacitance is 20 $\mu\mu$f.

Example—Calculation of Gain for 6SG7 Stage

Suppose that an amplifier stage having a gain of approximately 7 is required for the broadcast band, and that this order of gain is desirable at all frequencies up to 16 megacycles. Since the stage gain is to be substantially uniform, rather than low at medium values of m, a high-transconductance tube is essential. The 6SG7 is well suited to the purpose.

It has a transconductance of 4,700 micromhos for a plate voltage of 250 volts, a screen voltage of 125 volts, and a grid bias of —1 volt. Its output capacitance of 7 $\mu\mu$f, the probable input capacitance of the following tube, and a reasonable allowance for circuit wiring, suggest that 10 $\mu\mu$f be assigned for each capacitance shown in Fig. 2; i.e., a total capacitance of 20 $\mu\mu$f, and a series capacitance to resonate with the inductance of the peaking coil at 16 megacycles, of 5 $\mu\mu$f. This latter value has a reactance of 2,000 ohms at this frequency. The peaking coil, therefore, must have a reactance of 2,000 ohms at 16 megacycles and hence, an inductance of 20 microhenries.

A choice of 3 as the value of K leads to curve 8 of Fig. 2 as the form of the stage-gain characteristic for low and medium values of m. The ordinate when m is 0.05 shows that the stage gain is approximately the desired value of 7 at 0.8 megacycle, for a transconductance of 4,700 micromhos. Since curve 9 is suitable at the high-frequency end of the range, 20 is chosen as the value of P. Curve A of Fig. 3 shows the gain of the stage plotted against frequency. Curve B applies if K is 5, all other factors being the same as for curve A.

THEORETICAL STAGE-GAIN CURVES FOR CIRCUIT OF FIG. 2 WHEN USING THE 6SG7

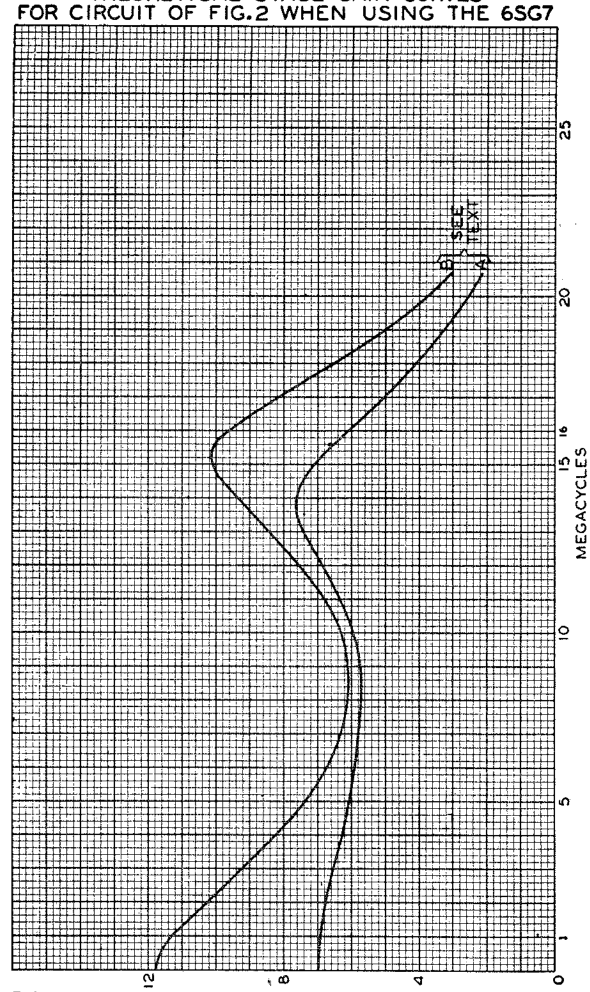


FIG. 3

Double Barrelled Type Numbers

The use of double barrelled type numbers for receiving valves has now been discontinued, and all such types will in future be known by their GT number. For example, Type 25L6-GT/G will become 25L6-GT.

The double barrelled number was used during the transition period between the earlier G types and the later GT types. The GT type superseded and replaced the earlier G types, and the suffix to the type number was used to indicate this fact.

It may be some considerable time before present stocks of the double barrelled type numbers are exhausted, but it should be understood that the valve is the same whether the type number is GT or GT/G.

This remark does not apply to transmitting valves, some of which will continue to use double barrelled type numbers, e.g., 866A/866.

NEW R.C.A. RELEASES

Radiotron types 5UP1, 5UP7, 5UP11—the 5U series of cathode ray tubes consists of three five-inch types—5UP1, 5UP7, 5UP11—differing only in the spectral-energy emission and persistence characteristics of their respective phosphors P1, P7, and P11. All of these types use electrostatic deflection and electrostatic focus.

Radiotron type 9C26—is a forced air cooled power triode intended for F-M service. It has a maximum rated plate dissipation of 7.5 kilowatts, and may be operated with full plate voltage, plate input, and grid current at frequencies as high as 30 Mc/s. Operation at frequencies up to 108 Mc/s is permissible with reduced ratings.

Radiotron type 889-A—is an improved water and forced air cooled transmitting triode superseding the earlier type 889.

Radiotron type 889R-A—is an improved forced air cooled transmitting triode superseding the earlier type 889-R.

Radiotron type 914-A—is an improved nine inch cathode ray tube superseding the earlier 914.

R.C.A. TYPES WITHDRAWN

Type 3EP1/1806-P1 (obsolete).
Type 9JP1/1809-P1 (obsolete).
Type 907 (obsolete).
Type 914—replaced by type 914-A.
Type 889—replaced by type 889-A.
Type 889-R—replaced by type 889R-A.
Type 7193—replaced by commercial equivalent type 2C22.

THE SELECTION OF VALVES FOR USE AS CATHODE FOLLOWERS

Not all types of valves are entirely satisfactory as cathode followers, or even with an unbypassed cathode bias resistor. No trouble usually occurs with triodes, but care should be taken to see that a pentode should have the internal screen, if any, connected to some pin other than the cathode pin. It is important that the internal screen be effectively earthed and this cannot be done with a cathode follower if the screen is connected to the cathode inside the valve.

This is the case with types 6AG5 and 6AK5, so that neither of these types is desirable for use as a cathode follower. Types 6AU6 and 6 BA6 are entirely satisfactory for use as cathode followers, since their internal screens are connected to the suppressor grid and not to the cathode. In this case the pin leading to the suppressor grid and internal shield should be returned to a point of approximately the same d.c. voltage as that of the cathode, but should be adequately bypassed to earth.

Radiotron types 6U7-G and 6D6 differ from valves of the same type numbers by other manufacturers in that the internal screen is connected to pin number 5 (suppressor) as indicated on the A.W.V. data sheets. This is specially to permit the use of these valve types in certain equipment in which the cathode is not earthed.

Radio engineers are advised to pay careful attention to the connection of the valve internal screen in all cases in which the cathode is not effectively earthed.

PREFERRED LIST OF TRANSMITTING VALVES

As a guide to the designers of transmitting equipment in Australia, we have prepared a Preferred List of transmitting valves. This list covers all normal applications up to several hundred megacycles per second, but it is not intended to restrict the availability of other types, particularly for experimental or developmental purposes. The list is necessarily a fluid one and will be brought up to date at intervals. In preparing the list, the requirements of F-M Broadcasting and mobile communication were borne in mind, as well as certain other applications on the ultra high frequencies.

A good proportion of the types shown in this list are already manufactured in Australia, while the balance is being considered for local manufacture when the demand becomes sufficient.

Those interested in making use of these valve types are invited to communicate with the Applications Department of our Company for technical data and advice regarding the best choice of types for particular purposes.

All orders for radio transmitting valves and rectifiers should be directed to the Engineering Sales Department of A.W.A.

A.W.V. PREFERRED TYPES LIST — TRANSMITTING VALVES AND RECTIFIERS

TYPE.	DESCRIPTION.	MAX. PLATE INPUT (watts)	MAX. PLATE DISSIPATION (watts)	MAX. FREQUENCY. full ratings (Mc/s.)	MAX. FREQUENCY. reduced ratings (Mc/s.)	COMMENTS.
807	Beam ampl.	60	25	60	125	Australian-made
809	Triode	75	25	60	120	Australian-made
811	Triode	155	40	60	100	Being considered for local production.
828	Beam ampl.	200	70	30	75	
805	Triode	315	125	30	80	Australian-made
810	Triode	500	125	30	100	Australian-made
8003	Triode	330	100	30	50	Australian-made
813	Beam ampl.	360	100	30	120	Australian-made
250TH	Triode	—	250	40	—	Being considered for local production.
833-A	Triode	1800	400	20	75	Australian-made

VALVES FOR HIGHER FREQUENCIES

7193	Triode	—	3.3	200	300	Identical with 2C22
2E24	Beam ampl.	40	13.5	125	175	ICAS ratings only
2E26	Beam ampl.	30	10	125	175	CCS ratings
815	P.P. beam ampl.	60	20	125	200	CCS ratings
829-B	P.P. beam ampl.	120	40	200	250	CCS ratings
826	Triode	125	60	250	300	CCS ratings
4.125A/ 4D21	Tetrode	675	125	120	250	CCS ratings

WATER-COOLED OR RADIATOR-COOLED TYPES

		KW	KW			
892	Triode	30	10	1.6	20	CCS ratings
892-R	Triode	18	4	1.6	20	CCS ratings
889R-A	Triode	16	5	25	100	CCS ratings
9C21	Triode	150	40	15	25	CCS ratings
9C22	Triode	100	20	5	25	CCS ratings
6C24	Triode	1.5	0.6	160	—	CCS ratings
7C24	Triode	5	2	110	—	CCS ratings

RECTIFIERS

TYPE.	MAXIMUM Peak Inverse Voltage.	Peak Plate Current.	Average Plate Current.	COMMENTS.
866A/866	10,000	1.0A	0.25A	Australian-made.
872A/872	10,000	5.0A	1.25A	Australian-made, for replacement purposes only. Type 8008 should be used in new equipment.
8008	10,000	5.0A	1.25A	Planned for Australian production; to supersede type 872A/872 for new equipment.
673	15,000	6.0A	1.5 A	Will be manufactured in Australia when the demand justifies local production; will be imported for the immediate future.

COMMENTS ON INDIVIDUAL TYPES

Type 7193 is a direct equivalent of type 2C22, and both types may be used interchangeably. Type 9072 is electrically somewhat similar but the grid and plate top caps are reversed.

Type 2E24 is designed for intermittent operation only, and has a quick heating filament.

Type 2E26 has performance almost identical to that of type 2E24 and is intended for use under continuous operating conditions.

Type 826 has been included solely for use above 200 Mc/s., since its special characteristics are only of advantage on frequencies of the order of 250 to 400 Mc/s.

Type 250TH is being considered for local production, and we hope shortly to make an announcement in this regard.

RADIOTRON 12AU7 TWIN-TRIODE AMPLIFIER

RCA-12AU7 is a heater-cathode type of medium-mu, twin-triode amplifier featuring a small glass envelope with integral button 9-pin base, separate terminals for each cathode, and a mid-tapped heater to permit operation from either a 6.3- or 12.6-volt supply.

Having characteristics which are very similar to those of the larger types 6SN7-GT and 12SN7-GT the 12AU7 like these types is useful in many diversified applications including multivibrators, synchronizing amplifiers, oscillators, mixers, and numerous industrial control devises. In such equipment, the 12AU7 can be used to advantage because of its compact size, its separate cathode terminals, and its economical consumption of heater power at either of the two voltages.

General Data

Electrical:

Heater for Unipotential Cathodes:

Heater Arrangement	Series	Parallel	
Voltage (AC or DC)	12.6	6.3	Volts
Current	0.15	0.3	Ampere

Direct Interelectrode Capacitances:*

	Triode Unit T_1	Triode Unit T_2	
Grid to Plate	1.5	1.5	$\mu\mu f$
Grid to Cathode	1.6	1.6	$\mu\mu f$
Plate to Cathode	0.50	0.35	$\mu\mu f$

Mechanical:

Mounting Position	Any
Maximum Overall Length	2-3/16"
Maximum Seated Length	1-15/16"
Length from Base Seat to Bulb Top (excluding tip)	1-9/16" ± 3/32"
Maximum Diameter	7/8"
Bulb	T-6-1/2
Base	Small Button Noval 9-Pin

* With no external shield.

CLASS A1 AMPLIFIER

Values are for each unit

Maximum Ratings, Design-Centre Values:

PLATE VOLTAGE	300 max.	Volts
PLATE DISSIPATION	2.75 max.	Watts
CATHODE CURRENT	20 max.	Ma.

Peak Heater-Cathode Voltage:

Heater negative with respect to cathode	180 max.	Volts
Heater positive with respect to cathode	180 max.	Volts

Characteristics:

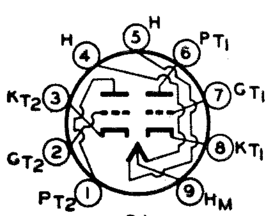
Plate Voltage	100	250	Volts
Grid Voltage	0	−8.5	Volts
Amplification Factor	19.5	17	
Plate Resistance	6250	7700	Ohms
Transconductance	3100	2200	Micromhos
Plate Current	11.8	10.5	Ma.

Maximum Circuit Values (for maximum rated conditions):

Grid-Circuit Resistance:

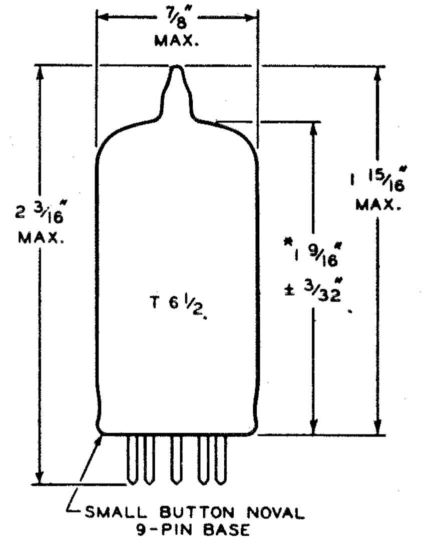
For cathode-bias operation	1.0 max.	Megohm
For fixed-bias operation	0.25 max.	Megohm

SOCKET CONNECTIONS
Bottom View



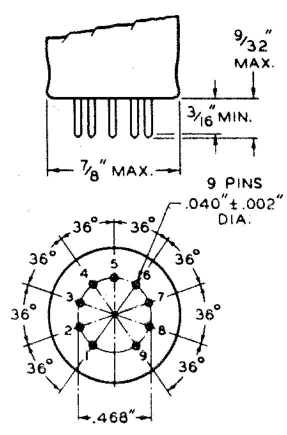
PIN 1 - PLATE (TRIODE No.2)
PIN 2 - GRID (TRIODE No.2)
PIN 3 - CATHODE (TRIODE No.2)
PIN 4 - HEATER
PIN 5 - HEATER
PIN 6 - PLATE (TRIODE No.1)
PIN 7 - GRID (TRIODE No.1)
PIN 8 - CATHODE (TRIODE No.1)
PIN 9 - HEATER MID-TAP

DIMENSIONAL OUTLINE



* MEASURED FROM BASE SEAT TO BULB-TOP LINE AS DETERMINED BY RING GAUGE OF 7/16" I.D.

SMALL BUTTON NOVAL 9-PIN BASE



THE PINS WILL FIT A FLAT-PLATE GAUGE HAVING THICKNESS OF 1/4" AND TEN HOLES 0.0520" ± 0.0005" SO LOCATED ON A 0.4680" ± 0.0005" DIAMETER CIRCLE THAT THE DISTANCE ALONG THE CHORD BETWEEN ANY TWO ADJACENT HOLE CENTERS IS 0.1446" ± 0.0005".

THE DESIGN OF SOCKET SHOULD BE SUCH THAT CIRCUIT WIRING CAN NOT IMPRESS LATERAL STRAINS THROUGH THE SOCKET CONTACTS ON THE BASE PINS. THE POINT OF BEARING OF THE CONTACTS ON THE BASE PINS SHOULD NOT BE CLOSER THAN 1/8" FROM THE BOTTOM OF THE SEATED TUBE.