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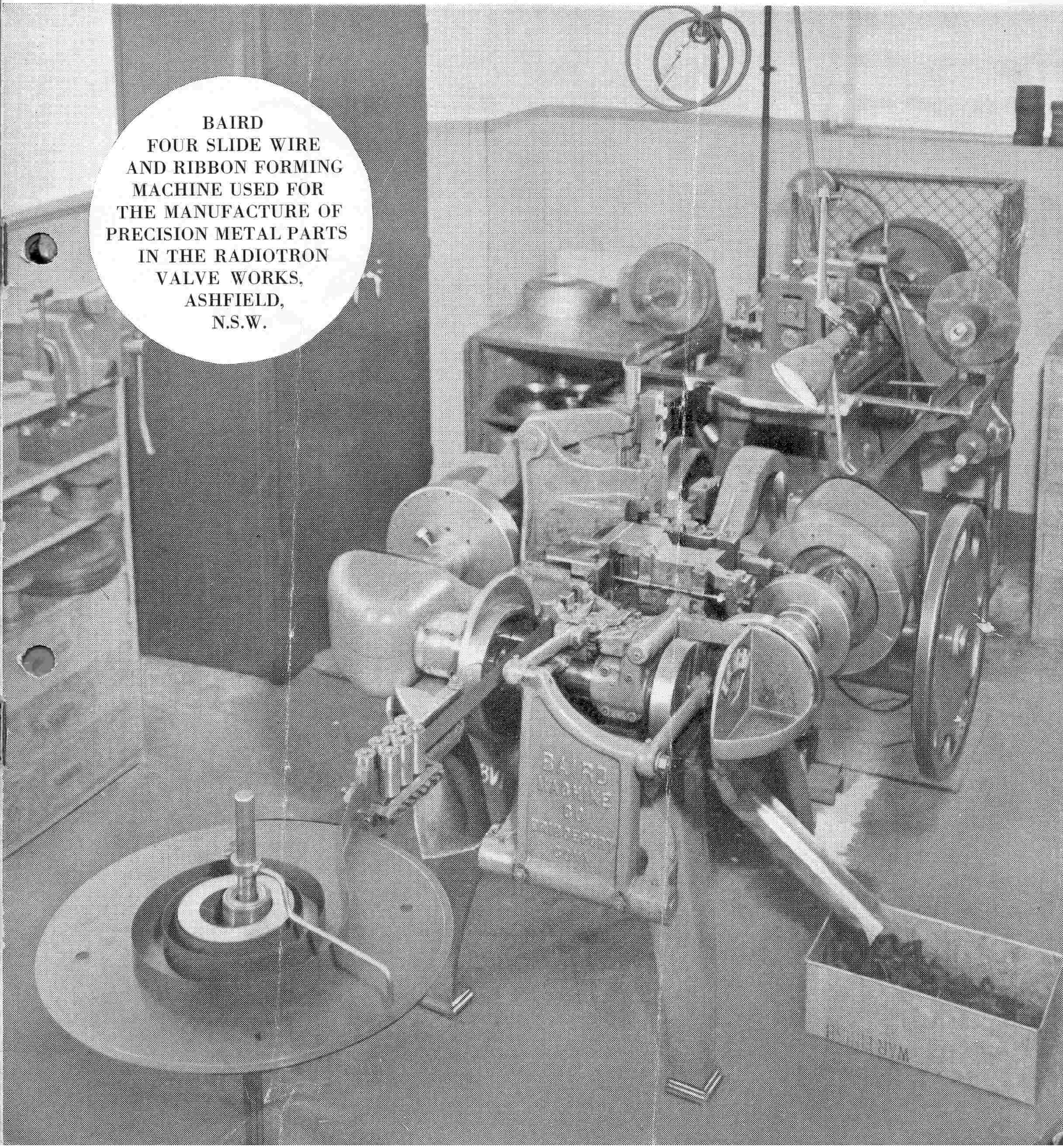

BAIRD FOUR SLIDE WIRE AND RIBBON FORMING MACHINE USED FOR THE MANUFACTURE OF PRECISION METAL PARTS IN THE RADIOTRON VALVE WORKS, ASHFIELD, N.S.W.

RADIOTRONICS

Number 128 **NOVEMBER — DECEMBER** **1947**

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Our Cover . .

Shows one of the Baird Four Slide Machines in use at the Ashfield Works.

This machine is set up for and is processing a completed lock seam plate in a series of progressive operations from the carbonised strip shown in the foreground.

These precision machines are tooled to cover a wide range of parts and are capable of producing large quantities of these parts to within very fine tolerances.

The Design of a High Fidelity Amplifier

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

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

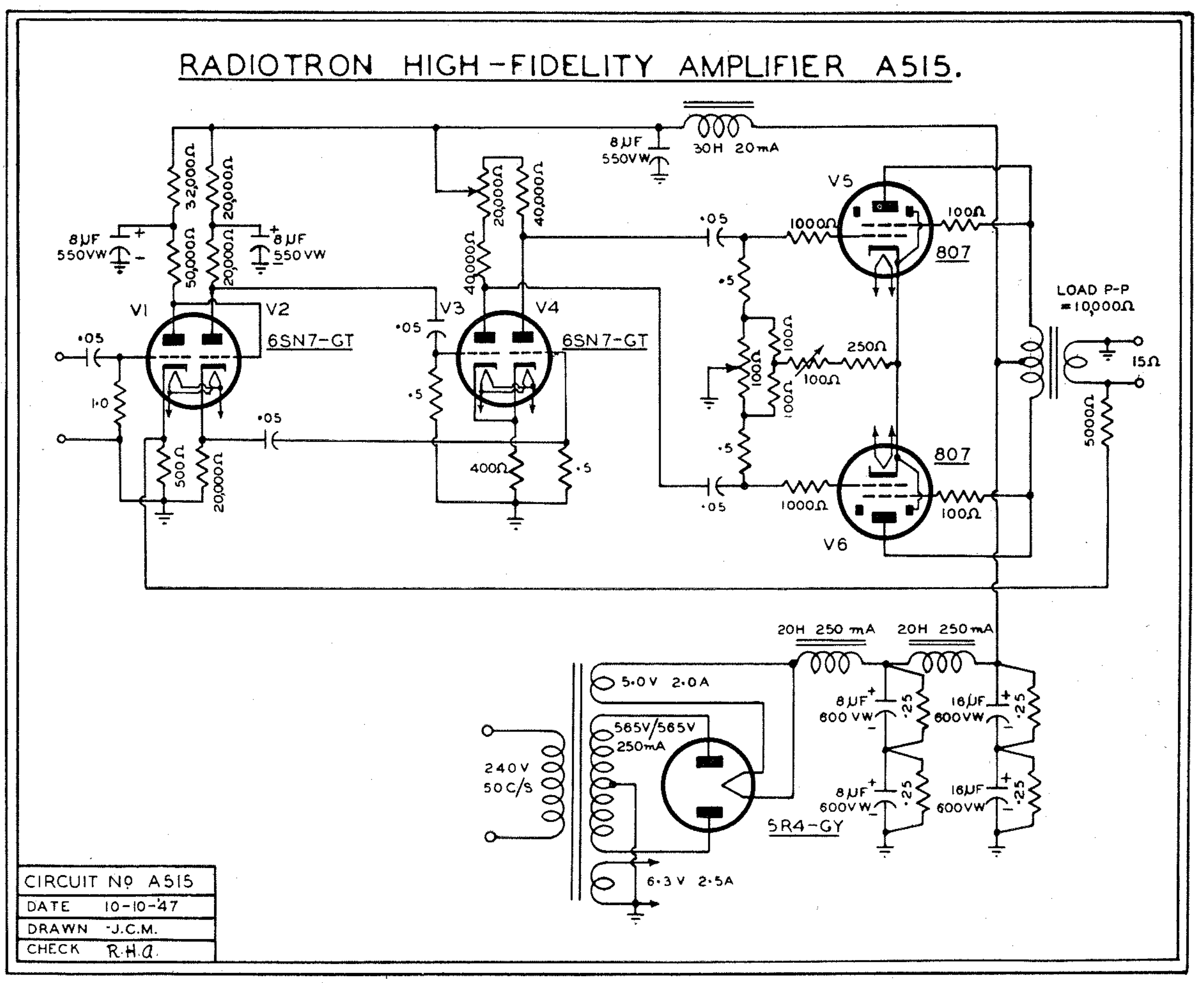
(3) A Design using Push-pull Triodes with Negative Feedback.

If it is desired to reduce distortion to the absolute minimum, there appear to be two possible approaches. The first of these has received the greater amount of publicity during recent years, and is the application of push-pull pentodes (or beam power valves) with a high degree of negative feedback. Although this is capable of providing extremely good results, the design is complicated owing to the extremely high degree of feedback and the precautions which require to be taken to avoid instability.

The second approach is through the application of push-pull triodes with negative feedback. In this case the initial distortion is considerably less, and the degree of feedback necessary to provide the required fidelity is not so great as to require abnormal precautions in the feedback loop.

We were interested to read an article by D. T. N. Williamson, "Design of a High Quality Amplifier" in the Wireless World, April and May 1947, using push-pull triodes with negative feedback. We therefore adapted it to suit our own valve types and then carried out tests to see whether the claimed low distortion limit of less than 0.1% total could be reproduced.

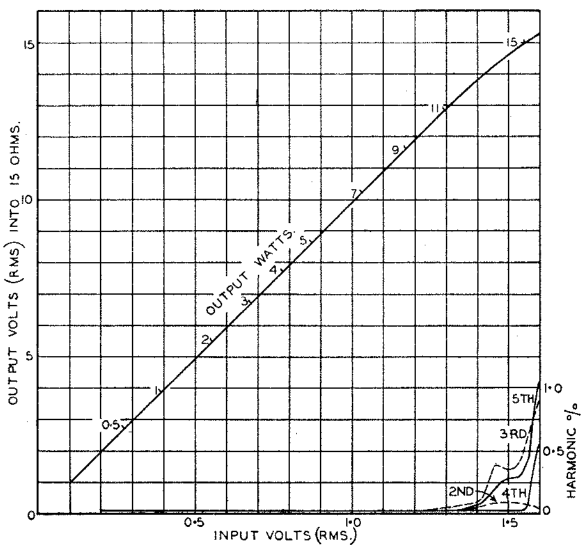
The original amplifier used one triode valve as a resistance coupled amplifier with its plate directly coupled to the grid of a second triode operated as a phase splitter with equal plate and cathode loads.



RADIOTRON HIGH-FIDELITY AMPLIFIER A515.

The plate and cathode of the second valve were capacitance coupled to the grids of the push-pull triodes, which in turn were capacitance coupled to the grids of the power valves. The latter were the English type KT66 which is somewhat close in characteristics to types 6L6 and 807. The KT66's were operated with plate to cathode voltages of about 400 volts which were in excess of the maximum triode ratings for either type 6L6 or 807.

In order to attempt to duplicate the results obtained in the English amplifier, tests were carried out on type 807 to see whether it could be used with the higher screen voltage in triode operation. During these tests we had no valve failures and subsequently arranged for a number of 807 valves to be placed on life test with triode connection and a combined plate and screen voltage of 400 volts to cathode. The



results of these tests will not be known for some months, but in the meantime there does not seem to be any reason why the valves should not be used under these conditions for individual amplifiers at the risk of the user, since the risk appears to be quite small. The screen dissipation is considerably below its maximum value and the only possible breakdown is through the increased screen voltage causing electrolysis or breakdown in the stem-press between adjacent leads. It would, of course, be possible for the output valves to be used within their maximum ratings by reducing the voltage from 400 to 300 but the power output would then be considerably smaller.

The valves used in the original voltage amplifier stages were type L63, which has characteristics resembling those of type 6J5-GT or one half of type 6SN7-GT. It was therefore decided to use two 6SN7-GT valves, one in place of each pair of L63 valves. This gave a valve complement of two type 6SN7-GT,* two type 807 and one type 5R4-GY rectifier. Special attention was given in the original design to the avoidance of phase shift. The direct coupling from the plate of the first valve to the grid of the second valve eliminated one possible cause of phase shift at low frequencies, while the cathode return circuit was purely resistive. All the cathode resistors are unbypassed so as to avoid any phase shift at low frequencies, leaving only the coupling condensers to the third and fourth stages, and the output transformer. The latter was required to have a primary inductance of at least 100 henrys, measured at 50c/s with 5 volts rms on the primary, a leakage inductance of not more than 30mH measured at 1000 c/s and a primary resistance of the order of 250 ohms. In order to avoid the delay in having a special transformer made for the job, we substituted various transformers which were on hand. It was found that the transformer supplied with the Goodmans 12″ loud speaker would give quite satisfactory results, and this was therefore adopted for the initial tests. An order has been placed for a transformer specially designed for this amplifier, which should be available at a later date. The amplifier was then wired up in accordance with the circuit diagram given herewith (A515).

The static plate currents in the final stage were balanced, and the drive to each valve was adjusted to give equal outputs at the plates when the valves were supplied with d.c. through separate chokes with individual loads of 5000 ohms each. The cathode bias resistors on the final stage were adjusted to keep the total plate dissipation less than 25 watts per valve. Without feedback, the amplifier was perfectly linear up to an output of 7.3 watts on a resistive load. With feedback, the amplifier was perfectly linear up to an output of 11.12 watts and it gave a smooth overload so that an output of 15 or 16 watts could be achieved without serious distortion.

The harmonic distortion for a power output of 11 watts was 0.01% 2nd harmonic, 0.04% 3rd harmonic, 0.01% 4th harmonic and 0.015% 5th harmonic. Even with an output of 17 watts as for an overload condition the distortion only reached the values of 0.46% 2nd harmonic, 0.31% 3rd harmonic, 0.53% 4th harmonic and 1.25% 5th harmonic. When tested on a speaker load it was found that for fairly large outputs at low frequencies a high frequency oscillation (about 60 Kc/s) would commence and be accompanied by a pulsed output of some other frequency. Both these incipient oscillations were cured by the addition of a 0.0005μF capacitor from earth to the screen of the 807 valve which tends to provide positive feedback to the cathode of the input stage. It was found that the by-pass was slightly more effective when connected to the screen than to the plate.

******If this type is not available, each 6SN7-GT may be replaced by two type 6J5-GT, or 6SJ7-GT (triode connected), 7193, 2C22, 9072, CV6 or any general purpose triode with a plate resistance of about 7000 to 8000 ohms and amplification factor about 20.*

The following tests were carried out with this addition. It was found practicable to apply sufficient feedback from the voice coil to the cathode of the input stage to provide 20db reduction in output with complete stability under all conditions. The phase shift between the input and output was quite small over a frequency range of 20c/s to 13 Kc/s.

The frequency response on a resistive load was constant from 20 to 10,000c/s and rose only 0.42db at 13,000c/s. On a speaker load the output was practically constant from 20 to 13,000c/s, using a Goodmans 12″ speaker, although the speaker was showing signs of frequency doubling with an input of 20c/s, as would be expected. As a matter of interest the resonant frequency of the speaker, when used in a vented baffle, was 45c/s.

It was found that harmonic distortion on the speaker load was very similar to that on a resistive load except at low frequencies, and the detailed values are tabulated at the end of the article. The distortion is extremely low down to a frequency of 40c/s.

At the time when these tests were carried out it was not possible to perform intermodulation distortion tests, but these will be carried out and published at the earliest opportunity. It is obvious that the total intermodulation distortion will be very low, and the tests are only intended to form a basis of comparison with other amplifiers.

SUMMING UP

This amplifier is by far the best which we have ever tested and we wish to give full credit to the original designer. It not only gives extraordinary linearity and lack of harmonic or intermodulation distortion but is comparatively simple and involves no special problems except the choice of output transformer. Unless the latter is specially designed for this circuit, or is one having extraordinarily low leakage inductance and exceptionally high primary inductance (such as the Goodmans transformer referred to in the article), the best results cannot be achieved.

TEST RESULTS

Static Measurements

(with Avo model 7).

	No output	11W. output	15W. output
Ground to plate of V1	82V		
Ground to cathode of V1	1.78V		
Ground to plate of V2	222.0V		
Ground to cathode of V2	90.0V		
Ground to plate of V3	148.0V		
Ground to cathode of V3	4.1V		
Ground to plate of V4	152.0V		
Ground to cathode of V4	4.1V		
Ground to cathode of V5 and V6	39.0V	39.2V	28.8V
Cathode to plates of V5 and V6	400.0V	401.0V	425.0V
Screen dissipation of V5 and V6	2.16W	1.7W	1.8W
Plate dissipation of V5 and V6	47.4W	37.3W	36.4W

	No output	11W. output	15W. output
Total B+ voltage	440.0V	441.0V	452.0V
Total B+ current	142.0mA	140.0mA	112.0mA
Ground to filament of 5R4-GY	465.0V	565.0V	565.0V
Input voltage to V1 (rms)		1.34V	1.55V
Input voltage (rms) to V1 with no feedback		0.125V	
Input voltage (rms) to V2		1.95V	2.83V
Input voltage (rms) to V3		3.3V	4.4V
Input voltage (rms) to V5 and V6 (grid-to-grid)		59.0V	85.0V
Output resistance		0.3 ohms	
Damping factor (R_L/R_o)		50.0	
Hum output (across 15)	0.34mV at 50 c/s		
	0.06mV at 100 c/s		

Performance Measurements On A515

Linearity with *no* feedback. Constant B+ supply. Plate-to-plate resistive load of 10,000 ohms.

input volts	output volts	gain	watts
400 c/s			
0.01	30.0	3000	
0.02	60.0	3000	
0.03	90.0	3000	
0.04	120.0	3000	
0.05	150.0	3000	
0.06	180.0	3000	
0.07	210.0	3000	
0.08	240.0	3000	
0.09	270.0	3000	7.3
0.1	295.0	2950	8.7
0.11	324.0	2945	10.5
0.12	335	2790	11.2
0.13	358	2750	12.8

Linearity with feedback. Resistive load (15 ohms).

input volts	output volts	gain	watts
400 c/s			
0.1	1.0	9.95	
0.2	1.95	9.95	
0.3	2.96	9.95	
0.4	3.95	9.95	
0.5	4.95	9.95	
0.6	5.94	9.95	
0.7	6.95	9.95	
0.8	7.97	9.95	
0.9	8.95	9.95	
1.0	9.95	9.95	
1.1	10.9	9.95	
1.2	11.95	9.95	
1.3	12.95	9.95	11.12
1.4	13.8	9.86	12.65
1.5	14.7	9.8	14.3
1.6	15.3	9.57	15.6

Frequency Response

Constant input of 0.05V without feedback and 0.5V with feedback.

c/s	Output volts: speaker load		Output volts: 15 ohms resistive load	
	no feedback	feedback	no feedback	feedback
20	4.8	5.05	4.65	4.95
30	5.62	5.0	4.88	4.95
40	6.3	4.95	4.95	4.95
50	6.3	4.9	5.0	4.95
60	5.8	4.9	5.05	4.95
70	5.75	4.9	5.06	4.95
80	6.15	4.9	5.08	4.95
90	6.3	4.9	5.1	4.95
100	6.12	4.9	5.1	4.95
1000	5.55	4.9	5.1	4.95
3000	5.9	4.9	4.98	4.95
4000	5.95	4.92	4.92	4.95
5000	5.95	4.92	4.88	4.95
6000	5.95	4.95	4.72	4.95
7000	5.95	4.95	4.65	4.95
8000	5.95	4.95	4.5	4.95
9000	5.95	4.95	4.4	4.95
10,000	5.95	5.0	4.3	4.95
11,000	5.95	5.0	4.2	4.96
12,000	5.95	5.0	4.08	4.97
13,000	5.95	5.0	4.02	4.98

Bass resonant frequency of Goodmans 12″ speaker in vented enclosure was 45 c/s.

Harmonic Analysis

Resistive load of 15 ohms at frequency 400 c/s.

Output volts	Watts	Harmonics per cent. H_2	H_3	H_4	H_5
2.74	0.5	0.025			
3.88	1.0	0.02			
5.49	2.0	0.02			
6.72	3.0	0.02			
7.76	4.0	0.025	0.01		
9.5	6.0	0.025	0.01		
10.98	8.0	0.025	0.01		
12.28	10.0	0.01	0.015		
12.88	11.0	0.01	0.04	0.01	0.015
13.42	12.0	0.01	0.08	0.01	0.06
13.98	13.0	0.03	0.38	0.01	0.21
14.5	14.0	0.075	0.35	0.01	0.26
15.0	15.0	0.07	0.54	0.01	0.28
15.5	16.0	0.04	0.92	0.11	0.45
16.0	17.0	0.046	0.31	0.53	1.2
Oscillator distortion (approximate)		0.06	0.035		

Harmonic Analysis

Speaker load, with power output of 11 watts,measured across the voice coil.

Frequency c/s	Harmonics per cent. H_2	H_3	H_4	H_5
40	0.18	0.98	0.14	0.23
50	0.32	0.4	0.05	0.05
60	0.22	0.14		
75	0.18	0.07	0.02	0.075
100	0.18	0.07		
300		0.02		
1000	0.05	0.03		0.02
3000	0.04	0.07	0.01	0.09
5000	0.04	0.08		
7500	0.07			

Further harmonics were beyond the range of the analyzer.

k Measurements

by B. SANDEL, A.S.T.C.

It would be thought that the measurement of the coefficient of coupling k in coils for use in radio receivers would be such a well known process as not to require further comment. Experience has shown, however, that this is far from the truth and many receiver and coil designers rely on such vague processes as adjusting the primary to secondary coupling of i-f transformers for maximum stage gain. This method is obviously unsatisfactory as quite large changes in k have only a small effect on the gain of an i-f stage but a serious effect on the selectivity curve. The final criterion should always be the measurement of the overall frequency response of the stage being investigated, but simple methods of adjusting k to approximately the value required when assembling an r-f or i-f transformer offer obvious advantages.

Bridge methods of measurement of k, where the bridge operates at audio frequencies, are not very satisfactory because the effects of stray capacitive coupling do not show up to the extent that actually occurs at radio frequencies. Radio frequency bridges are satisfactory where the coefficient of coupling is large, but a high degree of accuracy is difficult to obtain where k is of the order of 0.01, or less, as for an i-f transformer. It seems much more satisfactory to devise a method where measurements can be made at the actual working frequency.

A large number of methods have been investigated but the best results were obtained using a "Q" meter. As this instrument is usually available in receiver laboratories the methods using this instrument will be detailed.

Aerial and R-F Transformers

The measurement of k in these cases is usually fairly simple as the coefficient of coupling is of the order of 0.2 in most cases. A number of methods using the "Q" meter can be applied but the most satisfactory appears to be as follows.

Connect the winding to be tuned in the receiver to the "Q" meter and tune the coil to resonance at the desired working frequency with some convenient value of capacitance. The second winding is left on open circuit. Note the capacitance reading (C_1) and then short circuit the second winding. Alter the "Q" meter capacitance until resonance is again obtained and note this new capacitance value (C_2). Then

$$k = \sqrt{\frac{C_2 - C_1}{C_2}}$$

The value of k obtained obviously includes capacitive as well as mutual inductive coupling and is practically always the figure that it is required to know.

Some care is necessary where aerial and r-f coils have high impedance primaries as self resonance of these windings in the open circuit condition can cause errors. In this case it is suggested that the same procedure be carried out firstly with the primary and then the secondary connected to the "Q" meter, thereby checking that the values of k obtained are the same in each case. Shifting the frequency slightly will also assist if this difficulty should be serious. A little practical experience with the method soon resolves difficulties of this nature.

I-F Transformers

The method used for r-f and aerial coils is not applicable to i-f transformers because of the low values of k normally involved. That this is so can be readily seen by the following example.

Suppose k is 0.01 in a particular case and that it is required that C_1 be 100 $\mu\mu F$ to tune one winding to 455 Kc/s. Then

$$C_2 = \frac{C_1}{1 - k^2}$$

$$= \frac{100}{1 - 0.0001}$$

$$= 100.01\ \mu\mu F$$

The change between C_2 and C_1 is so small that it cannot be read with any degree of accuracy on a "Q" meter and so the method is unsatisfactory.

A suitable method is to tune one winding, as previously, with the second winding open circuited and note the value of "Q" (Q_1). Now tune the second winding to resonance without in any way altering the "Q" meter setting or connections. The "Q" meter reading alters as the second winding is tuned and the minimum reading for "Q" (Q_2) is noted. Then

$$\frac{k}{k_{crit}} = \sqrt{\frac{Q_1 - Q_2}{Q_2}} \quad \text{..........} \quad \text{(A)}$$

This is satisfactory as fairly large changes in "Q" occur and it is usually desired to know k in terms of critical coupling.

For the usual i-f transformer.

$$k_{crit} = \frac{1}{Q_1}$$

so k can be directly determined, if desired, from equation (A).

Clearly if k equals k *critical* then

$$Q_2 = \frac{Q_1}{2}$$

i.e. the initial "Q" reading is halved.

(*Continued on page* 111)

Input Circuit Noise Calculations for F-M and Television Receivers

by WILLIAM J. STOLZE
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This article which has previously appeared in the periodical Communications for February, 1947, has been made available for reprinting in Radiotronics by kind permission of the Editor and publishers of Communications and of the author.

Maximum Receiver Sensitivity is not, in most cases, determined by the gain of the particular receiver but by the magnitude of the input circuit noise, which is generated by the antenna, the tuned input circuit, and the first tube. This is true of a-m, f-m, and television, except that in f-m and television the random noise effect assumes a far greater degree of importance than in the standard broadcast band. The reason for this is twofold:

(1) At the frequencies where these two services operate, 50 to 250 mc, the relative value of the several different noise sources assume entirely new proportions and the heretofore unimportant and little known *induced grid noise* becomes one of the predominant components of the total.

(2) Most random input and tube noise is proportional to the square root of the bandwidth. Both television, with a 4-*mc* band, and f-m, with a 200-*kc* band, occupy much wider sections of the frequency spectrum than anything previously encountered by the commercial receiver engineer.

THERMAL AGITATION NOISE

When an alternating electric current flows through a conductor, electrons do not actually move along the conductor but they are displaced, an infinitesimal amount, first in one direction and then in the other. A voltage is built up across the conductor equal to the magnitude of the current times its resistance.

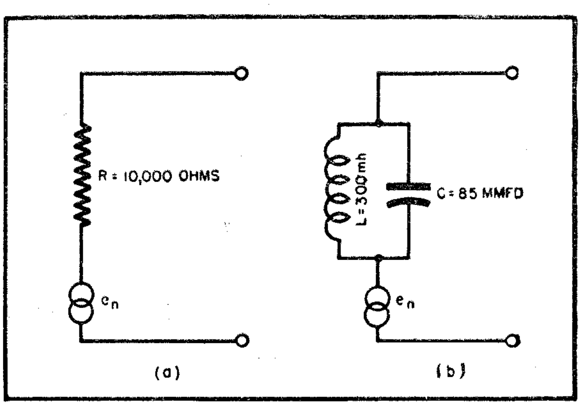


Figure 1.

Sample thermal noise circuits: Frequency, 1000 kc; Q, 100; L, 300 microhenries.

Applying heat to the conducting material agitates the molecules of the conductor and, consequently, varies the instantaneous position in space of the electrons. This random electron motion is, in a sense, a minute noise current flowing through the material and is known as *thermal agitation noise.* That is, the application of heat agitates the electron distribution of the substance thereby creating the noise.

The magnitude of the short-circuit noise current is given by

$$\bar{i}^2_n = \frac{4KT\Delta F}{R} \quad \ldots\ldots\ldots \quad (1)$$

where:

$\bar{i}^2_n$ = mean squared noise current (amperes^2)
K = Boltzmann's Constant (joules per degree Kelvin), 1.37×10^{-23}
T = temperature (degrees Kelvin)
ΔF = bandwidth (cps)
R = resistance (ohms)

All noise currents and voltages are random fluctuations and occupy an infinite frequency band. Because of the random effect, the most convenient terminology to use in expressing their magnitude is average noise-power output. Mean-squared noise current or mean-squared noise voltage, either of which is proportional to average power, is generally used.

In the expression for various noise components the term ΔF refers to the effective bandwidth of the circuit. This is determined from a curve of power output versus frequency by dividing the area under the curve by the amplitude of the power at the noise frequency in question. For most calculations, however, where only approximate values are desired, the bandwidth between half power points, or 0.707 voltage points, will give sufficient accuracy.

The equation below expresses thermal agitation noise as a voltage in series with a given resistor:

$$\bar{e}^2_n = 4KT\Delta FR \quad \ldots\ldots\ldots\ldots\ldots \quad (2)$$

Both the above forms are true of all resistive circuit elements or combination of elements including parallel and series-tuned circuits.

Referring to Figure 1 (a), let us suppose a resistance of 10,000 ohms were connected to the input of an amplifier with a 5 *kc* bandwidth, i.e., 5 *kc* between half power points or an audio band of 2.5 *kc*. At room temperature, 20° *C* or 293° *K*, the terms $4KT$ in the expressions for noise simplifies to 1.6×10^{-20}, which may be used in most receiver calculations. The noise in Figure 1 (a) is therefore:

$$e^2_n = 1.6 \times 10^{-20} \, \Delta FR$$

$$e_n = \sqrt{1.6 \times 10^{-20} \times 5{,}000 \times 10{,}000}$$

$$e_n = 0.89 \text{ microvolt.}$$

The noise bandwidth is generally determined by the narrowest element in the entire circuit under consideration. In the example for Figure 1 (b) the bandwidth of the amplifier is narrower than the tuned circuit and therefore its ΔF is used in the calculations.

Figure 1 (b) is a simple parallel-tuned circuit where the noise generating resistance is equal to the tuned circuit impedance. Again let us assume the bandwidth to be five kc per second.

$$R = Q\,(\omega L) = 100 \times 1900 = 190{,}000 \text{ ohms}$$

$$e^2_n = 1.6 \times 10^{-20} \, \Delta FR$$

$$e_n = \sqrt{1.6 \times 10^{-20} \times 5{,}000 \times 190{,}000}$$

$$e_n = 3.9 \text{ microvolts}$$

Thermal agitation noise voltage may be calculated easily with equation (2) but by using the graph shown in Figure 2 the room temperature values may be found directly.

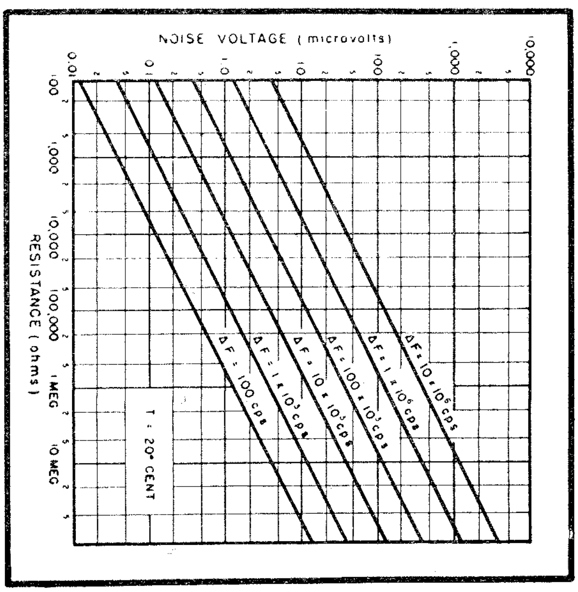


Figure 2.
Thermal agitation noise voltage versus resistance and bandwidth.

SHOT NOISE

Another important component of the total receiver noise is known as shot noise. This noise is generated inside the vacuum tube and is due to the random fluctuations in the plate current of the tube, or, to state it in another manner, random variations in the rate of arrival of electrons at the plate. When amplified, this noise sounds as if the plate were being bombarded with pebbles or as if a shower of shot were falling upon a metal surface, hence the name shot noise.

Although generated essentially in the plate circuit of the tube, which is not a convenient reference point for sensitivity or signal-to-noise ratio calculations, the shot noise is nearly always referred to as a noise voltage in series with the grid. Since the following equation is true,

$$e_g = \frac{i_p}{g_m} \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots \quad (3)$$

where

e_g = a-c grid voltage,
i_p = a-c plate current, and
g_m = transconductance,

by simply dividing the noise current in the plate circuit by the transconductance of the tube the shot noise may be referred to the grid and expressed in terms of grid voltage.

Another step is taken, however, to simplify the noise nomenclature. Suppose a given tube has a shot noise equal to e_n microvolts in series with its grid. It is perfectly valid to imagine that this voltage could be replaced by a resistance whose thermal agitation noise is equal to e_n (the shot noise) and considers the tube to be free of noise. This imaginary resistance, which when placed in the grid of the tube generates a voltage equal to the shot noise of the tube, is known as the shot noise equivalent resistance or just as the equivalent noise resistance of the tube. The advantage of this terminology is that when the equivalent noise resistance of the particular tube is known, the noise volts may be calculated directly for any given bandwidth by substituting values in the following formula:

$$e^2_n = 4KT\Delta FR_{eq} \quad \ldots\ldots\ldots\ldots\ldots \quad (4)$$

where R_{eq} = equivalent noise resistance, or at room temperature

$$e^2_n = 1.6 \times 10^{-20} \, \Delta FR_{eq} \quad \ldots\ldots\ldots \quad (5)$$

If the noise were expressed as a voltage or current its value would be correct only for one particular bandwidth.

By knowing the R_{eq} of any two given tubes their relative shot noise merit is also known regardless of what bandwidth they are to operate at, while if the noise voltages were given alone the operating bandwidth at which the calculation was made would also have to be noted if the relative merits of the two tubes were to be defined.

Noise-equivalent resistance values for a number of different tube types (triodes, pentodes, and converters) and for various circuit applications (amplifiers and mixers) can be calculated by applying the expressions presented in the chart, Figure 3.[1]

When the term converter is used it refers to a tube that is used for frequency conversion where the single tube acts as the local oscillator and the mixer (6SA7); the term mixer where two tubes are used, one as the mixer (6SG7), and one as the local oscillator (6C4).

[1] W. A. Harris, Fluctuations in Vacuum Tube Amplifiers and Input Systems, RCA Review; April 1941.

After the equivalent noise resistance is known the value of rms noise voltage at the grid of this tube can be calculated by applying the same expression that is used for thermal agitation noise,

$$e^2{}_n = 1.6 \times 10^{-20} \; \Delta FR$$

or by using the graph of Figure 2.

Figure 4 presents calculated equivalent noise resistance values for a number of commonly used tubes acting as various types of circuit elements. These are, of course, approximate figures.

It can be seen from Figures 3 and 4 that the noise resistance or voltage is at a minimum for a triode, increasing for the pentode and the multigrid tube, following in that order.

Shot noise is unique among the noise sources in the sense that the shot-noise voltage should be considered to exist in series with the grid inside the tube. The reason for this is that nothing can be done to the external grid circuit that will alter the magnitude of this component. Even though the shot noise must be tolerated, its effect can be minimized by designing the input circuit for maximum signal at the grid. This does not reduce the magnitude of the noise but does improve the signal-to-noise ratio of the receiver.

INDUCED GRID NOISE

Also present in the receiving tube is a third source of noise which is generated internally in the tube but whose magnitude and effect are determined partially by the external input circuit. Known as induced grid noise, this minute current is induced in the grid wires of the tube by random fluctuations in the plate current. It is known that a varying electron beam will induce a current in any nearby conductor. Therefore, the fluctuating plate current which is in a sense a varying electron beam, will induce a noise current in the nearby grid conductors.

TRIODE AMPLIFIER	$R_{EQ} = \dfrac{2\cdot5}{G_M}$
PENTODE AMPLIFIER	$R_{EQ} = \dfrac{I_B}{I_B+I_{G_2}}\left(\dfrac{2\cdot5}{G_M} + \dfrac{20 I_{G_2}}{G_M{}^2}\right)$
TRIODE MIXER	$R_{EQ} = \dfrac{4}{G_C} \qquad G_C = \dfrac{G_M}{4}$
PENTODE MIXER	$R_{EQ} = \dfrac{I_B}{I_B+I_{G_2}}\left(\dfrac{4}{G_C} + \dfrac{20 I_{G_2}}{G_C{}^2}\right)$
MULTIGRID CONVERTER OR MIXER	$R_{EQ} = 20\,\dfrac{I_B(I_K-I_B)}{I_K G_C{}^2}$

R_{EQ} = EQUIVALENT SHOT NOISE RESISTANCE

I_{G_2} = AVERAGE SCREEN CURRENT

G_M = GRID PLATE TRANSCONDUCTANCE

G_C = CONVERSION TRANSCONDUCTANCE

I_B = AVERAGE PLATE CURRENT

I_K = AVERAGE CATHODE CURRENT

Figure 3.
Equivalent shot-noise resistance formulas.[1]

The input impedance of a vacuum tube has a reactive and a resistive component. At relatively low frequencies the resistive component is very high (below about 30 *mc*); as the frequency is increased the resistive component decreases and its magnitude eventually becomes comparable to or even lower than the external grid circuit impedance. The resistive component is composed of two parts, the portion due to transit time effect, and the portion due to the inductance of the cathode lead.

An expression for induced-grid noise[2] for tubes with control grid adjacent to the cathode follows:

$$\bar{i}^2_{i.g.} = 1.4 \times 4\ KT_k\ \Delta F G_{elect}$$

or when expressed in the form of a voltage generator,

TUBE TYPE	APPLICATION	PLATE VOLTS*	SCREEN VOLTS	TRANSCONDUCTANCE MICROMHOS	EQUIVALENT NOISE RESISTANCE OHMS
6AC7	PENTODE AMPLIFIER	300	150	9,000	720
6AC7	PENTODE MIXER	300	150	2,200	2,800
6AG5	PENTODE AMPLIFIER	250	150	5,000	1,650
6AG5	PENTODE MIXER	250	150	1,250	6,600
6AG7	PENTODE AMPLIFIER	300	150	11,000	1,540
6AK5	PENTODE AMPLIFIER	180	120	5,100	1,880
6AK5	PENTODE MIXER	180	120	1,280	7,520
6AK6	PENTODE AMPLIFIER	180	180	2,300	8,800
6AT6	TRIODE AMPLIFIER	250	—	1,200	2,100
6AU6	PENTODE AMPLIFIER	250	150	5,200	2,660
6BA6	PENTODE AMPLIFIER	250	100	4,400	3,520
6BA6	PENTODE MIXER	250	100	1,100	14,080
6BE6	CONVERTER	250	100	475°	190,000
6C4	TRIODE AMPLIFIER	100	—	3,100	810
6C4	TRIODE MIXER	100	—	770	3,240
6C5	TRIODE AMPLIFIER	250	—	2,000	1,250
6C5	TRIODE MIXER	250	—	500	5,000
6J5	TRIODE AMPLIFIER	250	—	2,600	960
6J5	TRIODE MIXER	250	—	650	3,840
½6J6	TRIODE AMPLIFIER	100	—	5,300	470
½6J6	TRIODE MIXER	100	—	1,320	1,880
6K8	CONVERTER	250	100	350°	290,000
6SA7	CONVERTER	250	100	450°	240,000
6SB7-Y	CONVERTER	250	100	950°	62,000
6SC7	TRIODE AMPLIFIER	250	—	1,325	1,890
6SG7	PENTODE AMPLIFIER	250	125	4,700	3,100
6SG7	PENTODE MIXER	250	125	1,180	12,400
6SJ7	PENTODE AMPLIFIER	250	100	1,650	6,100
6SK7	PENTODE AMPLIFIER	250	100	2,000	11,000
6SL7	TRIODE AMPLIFIER	250	—	1,600	1,560
6SQ7	TRIODE AMPLIFIER	250	—	1,100	2,300

(*) VALUES OF PLATE VOLTAGE AND CURRENT AND SCREEN VOLTAGE AND CURRENT ARE FOR TYPICAL OPERATING CONDITIONS.

(°) CONVERSION TRANSCONDUCTANCE - MICROMHOS

Figure 4.
Approximate calculated equivalent noise resistance of various receiving-type tubes.

$$e^2_{i.g.} = 1.4 \times 4\ KT_k\ \Delta FR_{elect} \ \ldots\ldots \ (6)$$

where: T_k = cathode temperature (degrees Kelvin)
G_{elect} = electronic (transit time) component of input conductance
R_{elect} = electronic component of input resistance.

From equation (6) it can be seen that the induced grid noise is proportional to the electronic or transit time component of the input resistance. Measurement of the total input resistance is a comparatively simple matter with the use of a high frequency Q meter, but the separation of the electronic and the cathode inductance components, which are essentially two resistances in parallel between the grid and ground, is a very difficult matter. Since most high-frequency tubes are constructed with either two cathode leads or one very short lead, assuming the total measured input resistance to be electronic would not introduce too great an error. Another factor in favour of this approximation is that it would be the case for maximum induced grid noise and any error introduced would more than likely be on the safe side.

Cathode temperature in most receiving tubes, which almost exclusively use oxide-coated cathodes, is approximately five times the normal room temperature in degrees K. Equation (6) can be rewritten therefore as

$$e^2_{i.g.} = 5 \times 4KT\Delta FR_{elect} \ \ldots\ldots\ldots\ldots \ (7)$$

where: T = room temperature (degrees Kelvin), or when T = 300 degrees Kelvin

$$e^2_{i.g.} = 8 \times 10^{-20}\ \Delta FR_{elect} \ \ldots\ldots\ldots \ (8)$$

In circuit calculations this noise is essentially in series with a resistance equal to R_{elect} located between the grid and ground; Figure 5.

The approximate input resistance for a number of common receiving tubes in the frequency range of f-m and television is given in Figure 6. This chart can be used to find approximate input resistance values for induced grid-noise calculations.

[2] D. O. North, Fluctuations Induced in Vacuum-Tube Grids at High Frequencies, Proc. IRE: Feb. 1941.

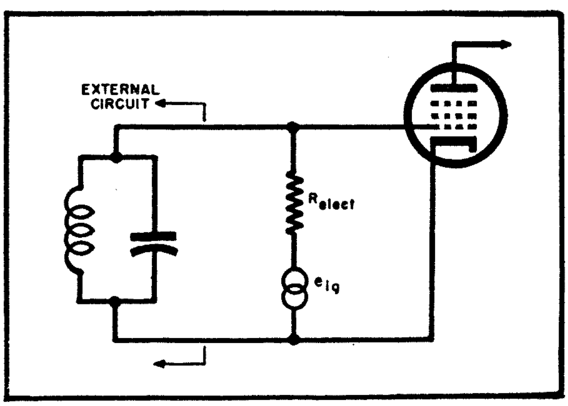


Figure 5.
Position of induced grid noise in vacuum-tube circuit.

TOTAL NOISE CALCULATIONS

Calculations of total input noise are made by using the grid of the input tube as a reference point. There are many sources of noise and each must be calculated and referred to the grid reference point before a summation is made. Since noise is a random effect and calculated on a power basis, the separate components cannot be added directly but as the square root of the sum of the squares.

$$\text{Total Noise} = \sqrt{e^2_1 + e^2_2 + e^2_3 + \text{etc.}} \ . \ (9)$$

The various noise voltages that must be referred to the first grid are:

(1) Thermal agitation noise of the antenna radiation resistance.

(2) Thermal agitation noise of the tuned grid circuit.

(3) Shot noise of the input tube.

(4) Induced grid noise of the input tube.

(5) Grid circuit noise of the following stages referred back to the first grid.

In Figure 7 (a) appears a diagram of a practical input circuit and the location of all the circuit parameters and noise voltages. Figure 7 (b) is essentially the same except that the antenna circuit is reflected through the transformer and considered to exist at the grid. This is the diagram that is most

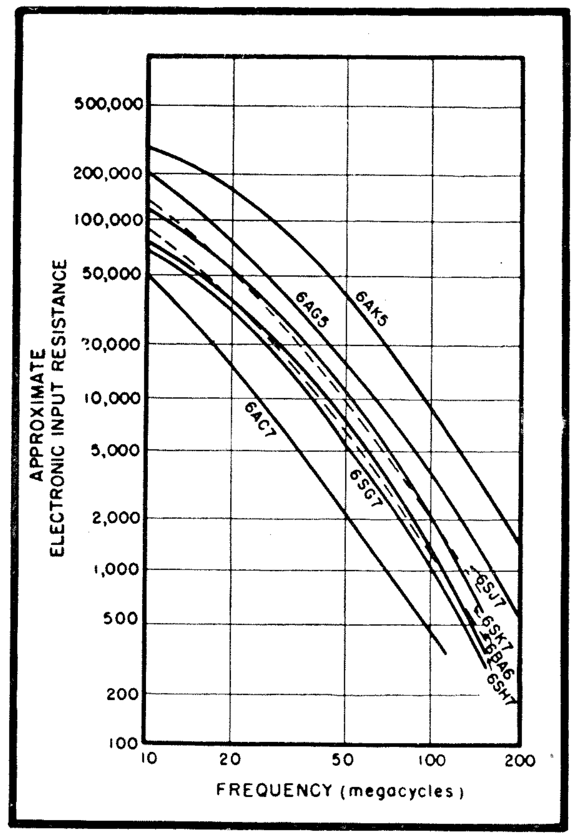


Figure 6.
Approximate electronic input resistance versus frequency.

useful in calculating the total input circuit noise.

The steps necessary to find specific values for each of these factors are shown in Figure 8. Antenna radiation resistance varies widely with the type of antenna chosen, but for f-m and television work it is generally in the order of 75 to 300 ohms. When the noise is known in terms of an equivalent resistance, as is the case here for the antenna, tuned circuit, and shot noise, the equivalent voltage can be either calculated or obtained directly from Figure 2.

In order to add the antenna, tuned circuit, and induced grid noise to the shot noise the effective voltage of these three components at the grid, or beween the points A and B, must be known. Each must go through what is essentially a resistive divider and may be calculated as shown in Figure 9.

After knowing the magnitude of the separate sources that exist between A-B, the total noise voltage is

$$e_{total} = \sqrt{\begin{aligned} e^2{}_{shot} &+ (e_{ant.}\ at\ A - B)^2 \\ &+ (e_{i\text{-}g.}\ at\ A - B)^2 \\ &+ (e_{ckt}\ at\ A - B)^2 \end{aligned}} \qquad (10)$$

One other factor may effect this total, however. If the total noise of the following stages, which is calculated similarly, ignoring the antenna of course, is appreciable, it must be added to the constants of Figure 9. In reflecting it to the first grid the second stage noise should be divided by the gain of the first tube. When the gain is about ten or more this factor may usually be neglected.

Effective signal voltage across A-B is calculated in the same way as the antenna noise in Figure 9. The signal-to-noise ratio is now also known.

Since the signal-to-noise ratio is determined by the signal strength and the total noise at the grid of the input tube, for a receiver that has a mixer, such as

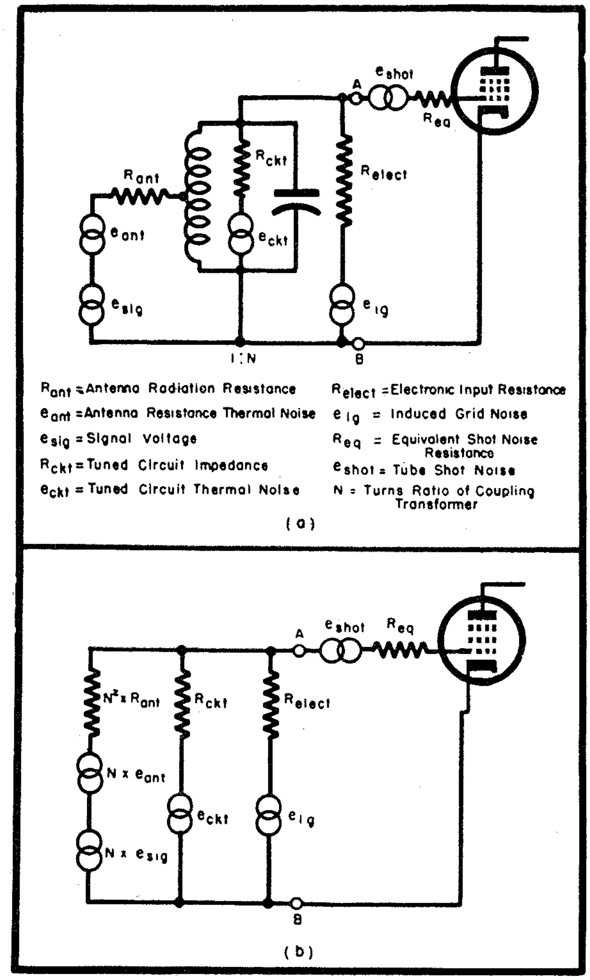


Figure 7.
Position of various noise sources in input circuit.

(1) R_{ANT} – DEPENDS UPON SPECIFIC ANTENNA

(2) $E_{ANT} = \sqrt{1{\cdot}6 \times 10^{-20}\, R_{ANT}\, \Delta F}$ – OR DIRECTLY FROM FIG. 2

(3) $R_{CKT} = Q\omega L = \dfrac{Q}{\omega C}$

(4) $E_{CKT} = \sqrt{1{\cdot}6 \times 10^{-20}\, R_{CKT}\, \Delta F}$ – OR DIRECTLY FROM FIG. 2

(5) R_{ELECT} – FROM ACCOMPANYING CHART, FIG. 6

(6) $E_{IG} = \sqrt{8 \times 10^{-20}\, R_{ELECT}\, \Delta F}$

(7) R_{EQ} – FROM ACCOMPANYING CHART, FIG. 4

(8) $E_{SHOT} = \sqrt{1{\cdot}6 \times 10^{-20}\, \Delta F\, R_{EQ}}$ – OR DIRECTLY FROM FIG. 2

Figure 8. Procedure for calculating various noise voltages.

6SK7, for the input tube, the signal-to-noise ratio may be considerably improved by the addition of an r-f tube, such as a 6SG7, which has considerably less total noise. By adding additional r-f tubes (6SG7s), however, since the total noise and signal at the grid will be the same, the signal-to-noise ratio will not be improved.

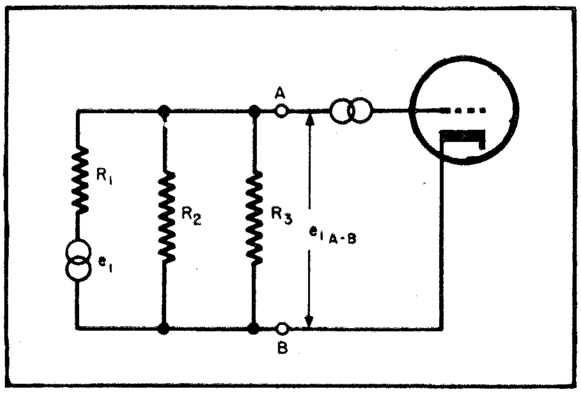


Figure 9.
Circuit for reflecting various voltages to the grid.

To find the effective voltage of the antenna, the tuned circuit, and the induced grid noise at the grid of the tube let R_1 equal one of the above noise resistances and e_1 its generated voltage. If R_2 and R_3 equal the other two noise resistances the effective voltage at the grid is

$$e_{1\ \text{A-B}} = \frac{e_1}{R_1 + \dfrac{R_2R_3}{R_2 + R_3}} \times \frac{R_2R_3}{R_2 + R_3}$$

This calculation must be performed for the three components in turn.

SAMPLE CIRCUIT CALCULATIONS

For a sample problem let us calculate the total noise at the grid of an f-m receiver r-f amplifier stage, assuming the circuit in Figure 10 (a) to be under consideration.

As a simplification of procedure the steps in the calculation will be numbered.

(1) $N^2\ R_{ant}$ = 1200 ohms (calculated)
(2) R_{elect} = 1200 ohms (Figure 6)
(3) R_{ckt} = $Q\omega L$ = 8000 ohms (calculated)
(4) R_{eq} = 3100 ohms (Figure 4)

At this point it will be convenient to redraw the circuit as shown in Figure 10 (b).

(5) $N\ e_{ant}$ = 2 microvolts (Figure 2).
(6) $e_{i.g.}$ = $\sqrt{8 \times 10^{-20} \times 200 \times 10^3 \times 1200}$ = 5 microvolts (equation (8))
(7) e_{ckt} = 6 microvolts (Figure 2)
(8) e_{shot} = 3.5 microvolts (Figure 2)

The next step is to find the effective voltage of each source between the grid and ground (or A – B) as shown in Figure 9.

$$(9)\quad e_{ant}\ A - B = \frac{2}{1200 + 1040} \times 1040 = 0.93 \text{ microvolt}$$

$$(10)\quad e_{i.g.}\ A - B = \frac{5}{1200 + 1040} \times 1040 = 2.3 \text{ microvolts}$$

$$(11)\quad e_{ckt}\ A - B = \frac{6}{8000 + 600} \times 600 = 2.5 \text{ microvolts}$$

And the total noise is therefore

$$(12)\quad e_{total} = \sqrt{3.5^2 + 0.93^2 + 2.3^2 + 2.52^2} = 4.9 \text{ microvolts}$$
(equation (10))

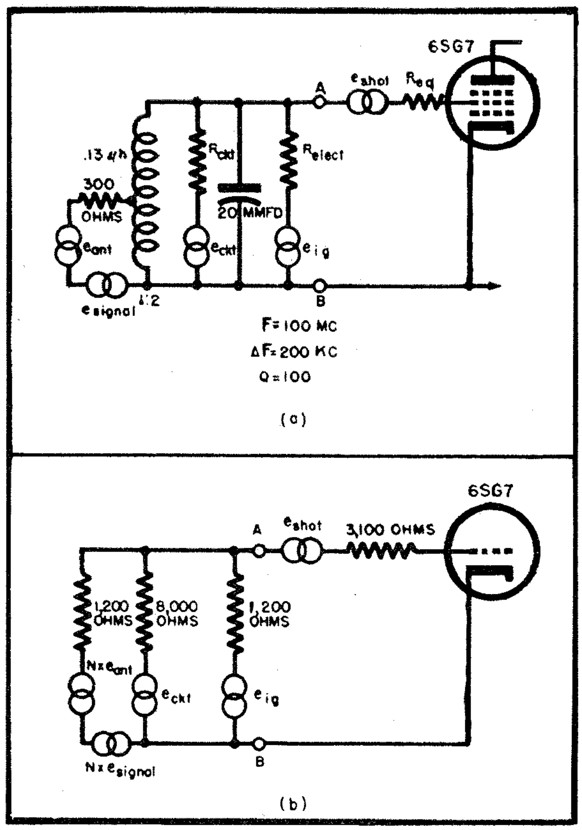


Figure 10.
Typical f-m receiver input circuit.

CONCLUSIONS

Selection of an input tube for a television or f-m receiver is dependent upon many varying circuit conditions and individual requirements. The choice of using balanced or unbalanced input, permeability or capacitor tuning, noisy pentodes or quiet triodes that possibly require neutralization, among others, lies entirely with the design engineer. Considering these reasons and various engineering and economic compromises, no particular tube can be chosen and defined as *the input tube.* Complete noise information about the circuits involved is necessary, however, as this is one of the determining factors for good sensitivity and signal-to-noise ratio.

R.C.A. Tube Types Not Recommended for New Equipment Design

The following recommendation has been issued by R.C.A. and is reprinted for the information of our readers, although conditions in Australia differ considerably from those in U.S.A.

An asterisk has been added opposite receiving types which are recommended for new equipment design in Australia at the present time.

Certain tube types should be avoided in the design of new equipment because they are approaching obsolescence or have limited or dwindling demand. Such types are listed below for the benefit of equipment designers.

RECEIVING TUBE TYPES

OZ4-G	2A6	6P5-GT	37
1A4-P	2A7	6Q7-G	38
1A6	2B7	6S7-G	39/44
1B4-P	2E5	6T7-G	49
1B5/25S	6A3	6U7-G*	50
1C6	6A4/LA	6W7-G	53
1C7-G*	6A6	6Z7-G	55
1D5-GP	6AC5-GT	12A7	56
1D7-G	6B5	15	57
1E5-GP	6B6-G*	19	58
1E7-G	6B7	24-A	59
1F4	6B8-G	26	75
1F5-G	6C6	27	76
1F6	6D6	30	77
1F7-G	6D8-G	31	78
1G5-G	6J7-G*	32	79
1H4-G	6J8-G*	32L7-GT	81
1H6-G	6K7-G	33	84/6Z4
1J5-G	6K8-G	34	85
2A5	6L5-G	35	89
	6L7-G	36	

TRANSMITTING TUBE TYPES

10-Y	804	850	893A-R
203-A	830-B	851	898
204-A	835	852	1608
207	838	858	1610
211	841	860	1619
217-C	842	861	1623
800	843	862-A	1624
801-A	846	865	1626
803	849	893-A	8012-A

CATHODE-RAY TUBE TYPES

2AP1-A	5BP1-A	12AP4	913
3AP1-A	5HP1-A	902-A	1840
	9AP4	908-A	

PHOTOTUBES

923	924

THYRATRONS

629	885	2051

MISCELLANEOUS

2V3-G	874	878

(Concluded from page 103)

Some care is necessary, such as taking all readings with the transformer in the can, and the can should be connected to the earth terminal of the "Q" meter.

It should be remembered that as the coefficient of coupling is low, stray coupling when the i-f transformer is wired into a receiver will modify the measured degree of coupling. A figure of about 0.8 of critical has been found fairly satisfactory with most ordinary broadcast receivers [although this figure may be modified in particular cases] since stray coupling increases this value to approximately critical. A figure of 0.95 has been used with 10.7 *Mc/s* i-f's in an f-m receiver where the stray coupling had been reduced to negligible proportions by careful lay-out.

It is necessary to emphasise that the bandwidths must be checked, in the receiver, to ensure that the desired degree of coupling has been obtained. Over-coupling can be readily determined by the unsymmetrical shape of the response curve, but in any case a calculated curve is always of assistance as it serves as a direct comparison for the desired curve. Methods of calculating i-f response curves have been given in Radiotronics 125, and if charts are used, as suggested, the process is a simple one and well worth the labour involved.

The special case of an i-f transformer having identical primary and secondary windings but unequal Q's because of circuit loading may be of interest. Measurement of the required coefficient of coupling is still possible, using the method described above, by loading the primary and secondary windings to give equal Q's and then determining k from the relationship

$$k'_{crit} = k\ crit \sqrt{\frac{Q_1 - Q_2}{Q_2}}$$

In this case k is made to have the value required to give critical coupling in the circuit where the Q's are unequal.

This required value of k is given by

$$k'_{crit} = \sqrt{\frac{1}{2}\left(\frac{1}{Q^2_p} + \frac{1}{Q^2_s}\right)}$$

This procedure is quite valid as the actual value of k is a physical property of the circuit. Whether this value of k gives over, under, or critical coupling merely depends on the tuned circuit operating conditions.

New R.C.A. Releases

Radiotron type 6S8-GT is a multi-unit valve combining three diodes and a high-mu triode in one bulb. One of these diodes has a separate cathode while the other two diodes have a single cathode which is also common to the triode unit. This arrangement of units provides for the detection and amplification of either AM or FM signals without switching detector circuits.

Radiotron type 1945 is a highly evacuated, ionization gauge tube sensitive to hydrogen but not to other gases or vapors. It is particularly useful in detecting and locating leaks in vacuum enclosures.

Unlike conventional ionization gauge tubes, the 1945 is constructed with a palladium plate which serves when cold as a vacuum-tight barrier to the vacuum system. As a result, the 1945 is exhausted to a much better vacuum than normally exists in a vacuum system. However, when heated, the palladium plate serves as a permeable membrane which permits hydrogen deliberately introduced in the vacuum system to flow into the 1945.

Because of its high vacuum, the 1945 can detect far smaller leaks than feasible with conventional ionization gauges operating at the same pressure as the vacuum system. Actually, an increase in hydrogen pressure of less than 0.0000001 mm of mercury (0.0001 micron) can be detected with the 1945 on systems using rotary vacuum pumps.

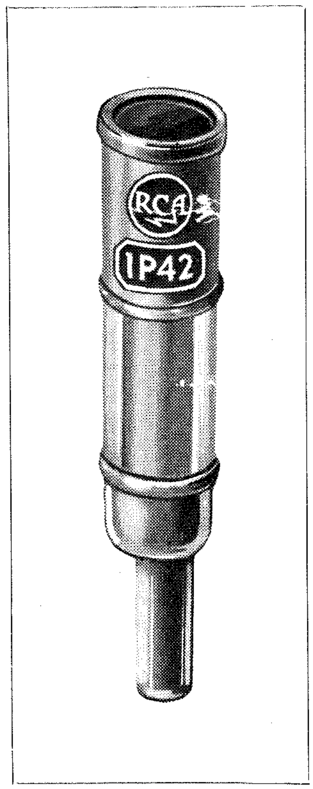
Radiotron type 1946 is a thermocouple type of hard-glass, vacuum-gauge tube useful in measuring gas pressures in the range from 1 mm to 0.0001 mm of mercury (1,000 microns to 0.1 micron), but having greatest sensitivity in the range from 1 mm to about 0.001 mm. Since the 1946 is not damaged under operating conditions involving loss of vacuum, it can conveniently be used as a protective device in vacuum systems to prevent application of voltages to electronic components operating within a vacuum enclosure until the pressure in the enclosure has been reduced to the desired value.

Radiotron type 1947 is a Pirani type of soft-glass, vacuum-gauge tube for measuring gas pressures in the range from approximately 0.5 mm to 0.01 mm of mercury (500 to 10 microns), but it can be used with reduced sensitivity to measure pressures above and below this range. The 1947, like the 1946, may be used as a protective device in vacuum systems.

Radiotron type 1949 is an ionization type of hard-glass, vacuum-gauge tube for measuring gas pressures in the range below 0.0001 mm of mercury (0.1 micron). When operated with a grid voltage of +110 volts and a plate voltage of —22.5 volts, it has a sensitivity of 110 microamperes per micron. It is, therefore, very sensitive for detecting minute leaks in a vacuum system.

Radiotron type 1950 is like the 1949 in application, but is constructed with soft glass and has somewhat different dimensions.

Radiotron type 5655 is a television camera tube recommended for studio use. It is similar to type 2P23 but differs in that its photocathode has practically no infrared response, its resolution is somewhat better, its signal to noise ratio has been improved about twice, and its response to half-tones is more natural. It does not, however, cover as wide a light range as the 2P23.



Radiotron type 1P42 is a small head-on type of high-vacuum phototube. The cathode surface is blue-sensitive but has negligible sensitivity to infra-red radiation. The maximum overall length is 1 13/32″ and the maximum diameter ¼″; the window area is 0.030 square inch.

This phototube is particularly useful where space considerations are a prime factor, and it can be operated in any desired mounting position.

QUICK REFERENCE CHART

A quick reference chart of miniature valve types is now available and may be obtained on application to Amalgamated Wireless Valve Co. Pty. Ltd., Sales Promotion, 47 York St. (Box 2516, G.P.O.), Sydney, N.S.W.