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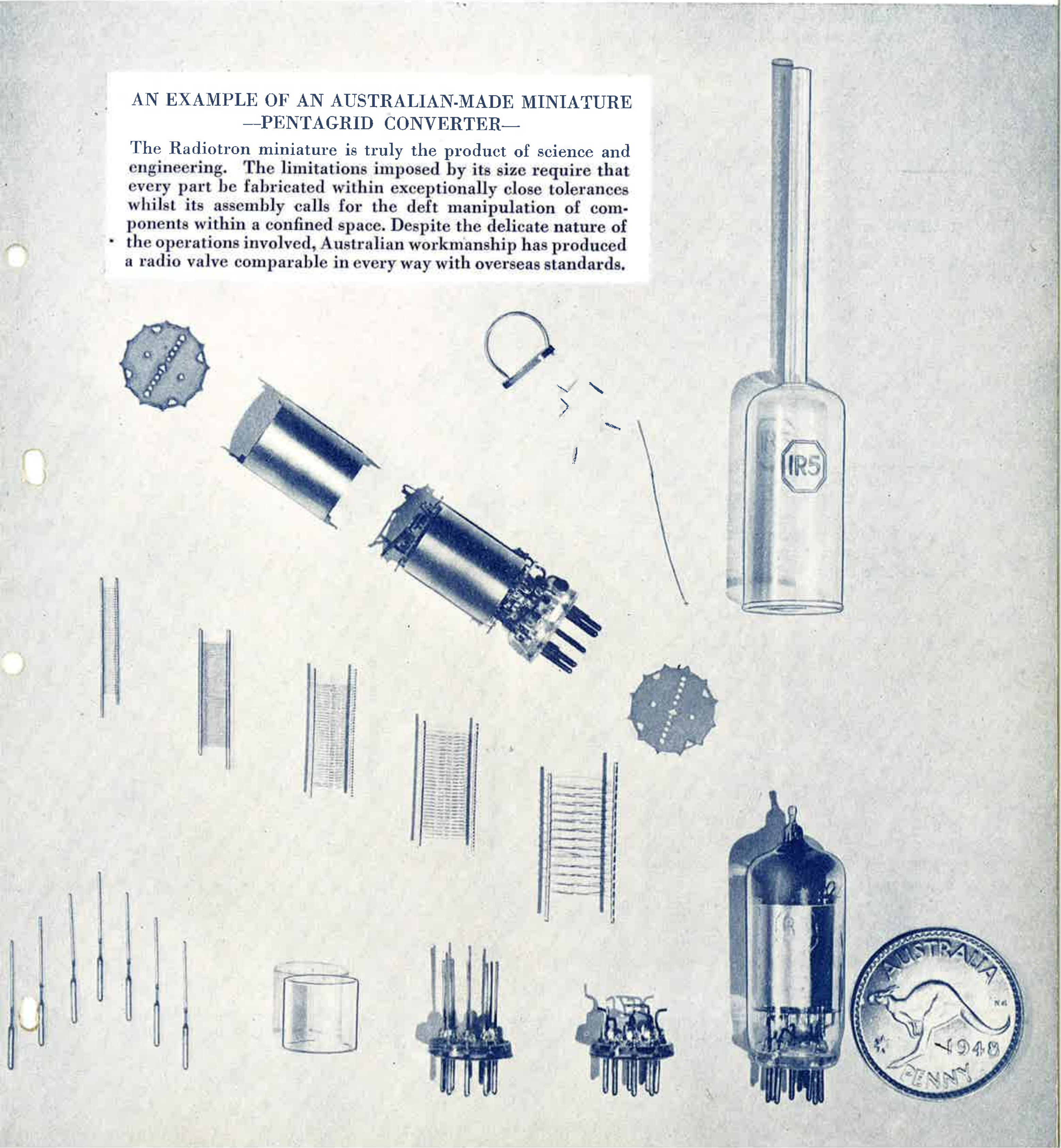
AN EXAMPLE OF AN AUSTRALIAN-MADE MINIATURE —PENTAGRID CONVERTER—

The Radiotron miniature is truly the product of science and engineering. The limitations imposed by its size require that every part be fabricated within exceptionally close tolerances whilst its assembly calls for the deft manipulation of components within a confined space. Despite the delicate nature of the operations involved, Australian workmanship has produced a radio valve comparable in every way with overseas standards.



Variable Bandwidth Crystal Filters

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

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(1) Behaviour of equivalent circuit.

Fig. 1 shows the generally accepted equivalent electrical circuit for a quartz crystal of the type used in the i-f stage of a communications receiver.

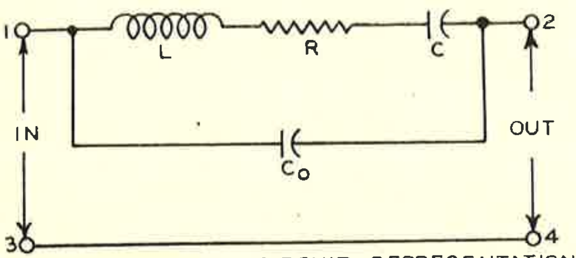


FIG. 1 ELECTRIC CIRCUIT REPRESENTATION OF QUARTZ CRYSTAL

In this network suppose we consider that R is zero, then between terminals 1 and 2 there is a reactance X equal to

$$X = -\frac{\omega L - \dfrac{1}{\omega C}}{\omega C_o\left[\omega L - \dfrac{1}{\omega}\left(\dfrac{1}{C} + \dfrac{1}{C_o}\right)\right]} \quad . \quad (1)$$

When $\omega L = \dfrac{1}{\omega C}$ the reactance is zero and the circuit is series resonant.

When $\omega L = \dfrac{1}{\omega}\left(\dfrac{1}{C} + \dfrac{1}{C_o}\right)$ the reactance is infinitely large, and the circuit is parallel resonant (anti-resonant).

The difference in frequency between f_p and f_r is given approximately by

$$\Delta f = \frac{C}{2C_o} f_r.$$

From the values of the 455 Kc/s crystal constants which are given in (iiiE) below, it will be seen that Δf is about 250 c/s in a typical case.

The presence of R will slightly modify the frequencies of series and parallel resonance and also the conditions for which parallel resonance occurs. For our purposes the conditions given are near enough.

It should be clear from the circuit that the parallel resonance frequency (f_p) will be higher than the series resonance frequency (f_r). For frequencies above f_r the resultant reactance due to L and C in series is inductive, and this is connected in parallel with a capacitive reactance due to C_o. A typical curve of output voltage versus frequency change, for this type of circuit, is shown in Fig. 2.

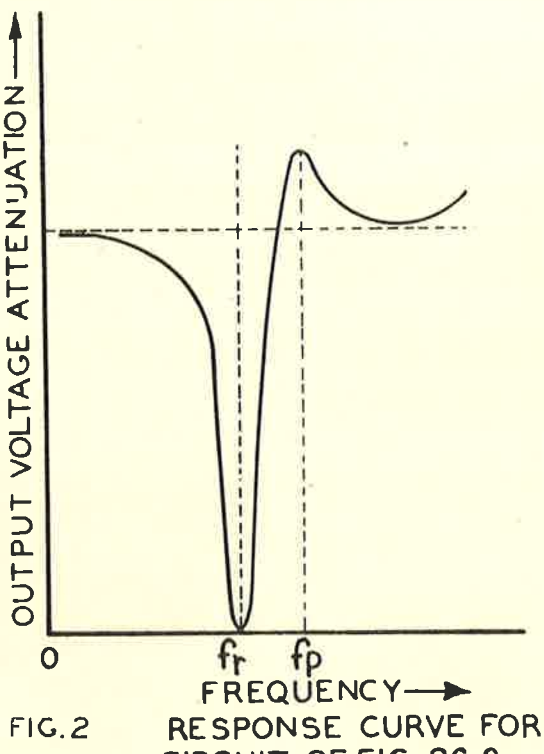


FIG. 2 RESPONSE CURVE FOR CIRCUIT OF FIG. 26·9

If the value of C_o could be altered as required, then the position of f_p would be variable. Suppose by some means we are able to connect a negative capacitance C_N (or a parallel inductance will give somewhat similar results) across C_o; then the value of the capacitance shunted across the series circuit will be

$$C' = C_o - C_N \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (2)$$

From this, it follows that C' can be reduced from its initial value of C_o (when C_N is zero) until it becomes zero, and C_N has then exactly neutralized C_o. The response curve would now be that for the series circuit (made up from L, R and C) alone, which behaves as a pure series resistance at f_r. If the magnitude of C_N is further increased, then C' becomes negative and the series circuit is shunted by a negative capacitance; which is equivalent to shunting an inductance across the series circuit (which behaves like a capacitive reactance for frequencies below f_r). This means that the parallel resonance frequency (f_p) will now be lower than the series resonance frequency (f_r). Any frequency above or below the series resonance frequency can now be chosen as the rejection, or parallel resonance, frequency and the current through the circuit, at this point, will be reduced; the values for f_p and the current depending on the circuit constants.

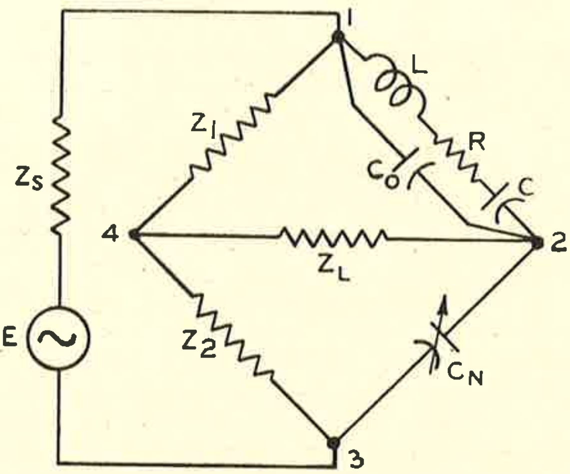


FIG. 3 CIRCUIT ARRANGEMENT FOR VARYING $C' (= C_O - C_N)$

To achieve the variation in C' the circuit of Fig. 1 (i.e. the crystal) is incorporated in the bridge circuit of Fig. 3, in which C_N is made variable to achieve the results discussed. In this circuit Z_L is the load impedance; Z_s is the impedance of the voltage source; Z_1 and Z_2 are any two impedances used to make up the resultant bridge circuit. It will be realized that the presence of Z_s and Z_L would have some modifying effect on our previous discussion, but the general principles remain unchanged; these impedances will be taken into account when the design of the crystal filted stage is carried out.

The arrangement of any practical circuit using a single crystal can be reduced to the general form of Fig. 3.

Two typical circuit arrangements are shown in Fig. 4. Fig. 4 (A) shows that Z_1 and Z_2 are obtained by using two capacitances C_1 and C_2 for the bridge arms. Fig. 4 (B) uses a tap on the i-f transformer secondary (L_1 and L_2) to obtain the bridge arms Z_1 and Z_2.

In any circuit arrangement, such as those shown, the best overall selectivity is obtained when C' is zero (i.e., $C_N = C_o$). Under this condition, if the circuit is designed with some care, the response curve of output voltage with frequency change will be reasonably symmetrical. If C_N is varied to place f_p above or below f_r, although we may improve the selectivity (and so the rejection) against an unwanted signal at f_p, the resonance curve is no longer symmetrical and the selectivity is decreased on the opposite side of the curve.

(ii) Variable bandwidth crystal filters.

It is often necessary, in practical receivers, to have means available for varying the bandwidth of the crystal filter circuit. This can be achieved in a number of ways, such as detuning the input and (or) output circuits or varying their dynamic impedances. Also, as well as having variable bandwidth it is very desirable that there should not be appreciable change in stage gain as the bandwidth is altered. Further, the overall response curve of voltage output versus frequency should remain as symmetrical as possible with changes in bandwidth when the bridge circuit is arranged so that C_o is neutralized by C_N.

To fulfil all of the conditions above, the simplest solution is to use a tuned circuit as the load for the filter circuit, and to alter its dynamic impedance by switching-in either series or parallel resistors. Simple detuning of the input circuit, or the output circuit, is not very satisfactory, and the best results could only be achieved by detuning both the input and output circuits in opposite directions by an amount depending on their relative Q's. (For a description of circuits using the latter method see Refs. 2, 17). To obtain constancy of gain it is necessary that the voltage source impedance be low, and this again suggests the switched output circuit when bandwidth is to be varied. The choice of series or shunt resistors to alter bandwidths is largely bound up in stray

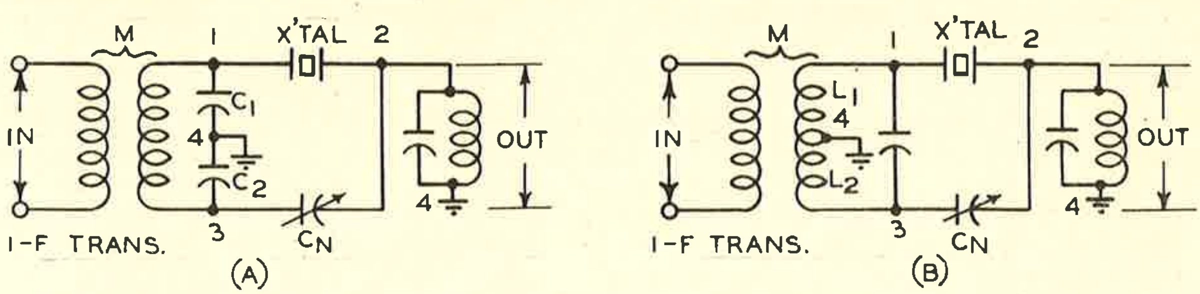


FIG. 4 PRACTICAL CIRCUIT ARRANGEMENTS FOR QUARTZ CRYSTAL FILTERS

capacitances across switch contacts and the resistors themselves; careful consideration here, suggests that resistors in series with the inductor, or the capacitor, of the tuned load will probably give the least detuning effects. Unless the values of the tuning capacitances for the input and load circuits are fairly large, alteration of C_N will have some appreciable effect on the resonant frequencies of these circuits, giving rise to asymmetry of the response curve and loss in sensitivity. This effect can sometimes be overcome, partly, by the use of a differential type of neutralizing capacitor. However, circuits using this arrangement should be carefully examined as usually there are still some detuning effects. For most practical cases it is sufficient to choose suitable values of tuning capacitance, particularly when operation is confined to an i-f of 455 Kc/s.

Two other points are worth mentioning. The first is that the crystal stage gives high selectivity around resonance, but the "skirt" selectivity may be quite poor; for this reason the other stages in the receiver must provide the additional "skirt" selectivity required. In addition, good "skirt" selectivity is a requirement of the i-f amplifier so as to minimize any possible undesirable effects which may arise because of crystal subsidiary resonances. The other point is that having C_N as a variable control is not necessarily a great advantage, and simpler operation is obtained when C_N can be pre-set to neutralize C_o. This cannot be done with all types of variable bandwidth circuits, since the conditions for neutralization may be altered as the bandwidth is changed.

(iii) Design of variable bandwidth i-f crystal filter circuits.

(A) Simplifying assumptions.

From Fig. 3, when a capacitance balance is obtained as far as C_o is concerned,

$$C_N \cong C_o \frac{Z_1}{Z_2} \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (3)$$

Since C_o is generally about 17 μμF it is convenient to make $Z_1 = Z_2$. Also, C_N must have very low losses if the attenuation at the rejection frequencies is to be high. The general design will call for all capacitors to be of the low loss type. When the condition of eqn. (3) is fulfilled, the equivalent circuit reduces to that of Fig. 5 (A). For purposes of analysis it is convenient to rearrange Fig. 5 (A) as shown in Fig. 5 (B), and, if we take $Z_1 = Z_2$, the value of Z'_s will be a quarter of the total dynamic impedance of the secondary circuit of the i-f input transformer. E'_s will be half the voltage developed across the two series capacitors tuning the i-f transformer secondary.

(B) Gain.

First we will derive an expression for the overall gain of the i-f stage.

From Fig. 5 (B),

$$\frac{E_o}{E'_s} = \frac{Z_L}{Z_L + Z'_s + Z_x} \quad . \quad . \quad . \quad . \quad (4)$$

where E_o = the output voltage across the load circuit applied to the grid of the following i-f amplifier valve

Z_L = the load impedance

Z'_s = the voltage source impedance

Z_x = crystal impedance (C_o neutralized) = $R + j\left(\omega L - \frac{1}{\omega C}\right)$

= R at resonance (i.e. i-f)

and E'_s = available voltage developed at tap on input i-f transformer secondary.

To compute the complete stage gain we must now consider Fig. 6. Here, since we require maximum gain, a critically coupled transformer is connected to a voltage amplifier valve (V_1) whose grid to cathode input voltage is E_g. Then since E'_s is half the total voltage appearing across C_1 and C_2

$$E'_s = \frac{1}{2}\left(\frac{g_m\, E_g\, Q\omega_o\, L}{2}\right) \quad . \quad . \quad . \quad . \quad (5)$$

where g_m = mutual conductance of V_1

$Q = \sqrt{Q_1 Q_2} = \frac{1}{k_c}$

$\omega_o = 2\pi \times$ i-f

and $L = \sqrt{L_1 L_2}$ (always provided that $L_1 C_1 = L_2 C_2$).

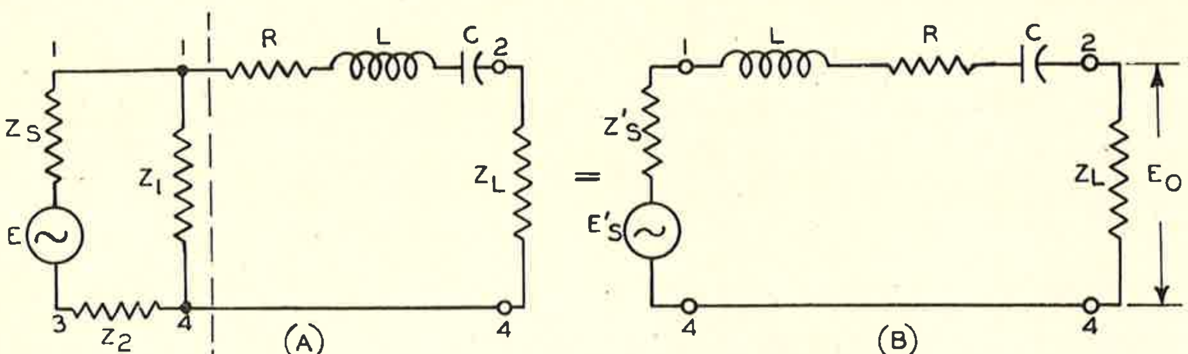


FIG 5 EQUIVALENT CIRCUITS FOR CRYSTAL FILTER WHEN $C_O = C_N$

Combining eqns. (4) and (5), the gain from the grid of V_1 to the grid of the next amplifier valve V_2 is

Stage Gain =

$$\frac{E_o}{E'_s} \times \frac{E'_s}{E_g} = \frac{E_o}{E_g} = \left(\frac{Z_L}{Z_L + Z'_s + Z_x}\right)\left(\frac{g_m Q \omega_o L}{4}\right) \quad . \quad . \quad . \quad . \quad (6)$$

Eqn. (6) gives a great deal of information about the circuit. For constancy of gain, it follows that Z_L should be very much greater than $(Z'_s + Z_x)$. Since Z_x (or R at resonance) cannot be altered by the receiver designer, it is necessary to make Z'_s as small as possible. However, as Z'_s is made smaller the overall gain will be affected since it is the centre tap on the output of the input i-f transformer. The

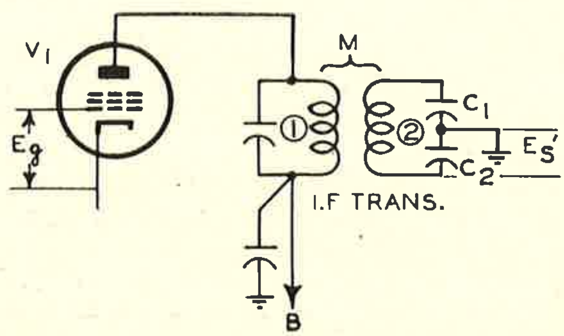


FIG.6 CIRCUIT FOR DETERMINING E_S

maximum value for Z_L will be limited by the maximum bandwidth requirements, and the permissible values of circuit constants. The other factors affecting gain are g_m, Q and L. For a given valve under a fixed set up operating conditions, g_m is practically outside the designer's control; the value of L $(=\sqrt{L_1 L_2})$ is made as large as possible consistent with the requirements of minimum permissible tuning capacitance; $Q = \sqrt{Q_1 Q_2}$ is adjusted so that Q_1 is made as high as possible, and since we desire Z'_s to be low, a suitable value is selected for Q_2. The method of determining Q has a large effect on the stage gain which can be obtained, as will be seen later in the illustrative example.

(C) Gain variation with bandwidth change.

For constancy of maximum gain at f_o (i.e. the i-f) it has been suggested that Z'_s should be small. There is a limit, however, to the minimum gain variation that can be obtained for a given change from maximum to minimum bandwidth. From the preceding equation it may be deduced that the maximum stage gain (A_{max}) is obtained when the bandwidth is greatest, and the minimum gain (A_{min}) when the bandwidth is least. This limit in gain variation (since R for the crystal is fixed) is the condition for which the voltage source impedance (Z'_s) becomes zero, which occurs when

$$\alpha_o = \frac{A_{min}}{A_{max}} = \frac{R_{T_2}\,(R_{T_1} - R)}{R_{T_1}\,(R_{T_2} - R)} \quad . \quad . \quad . \quad (7)$$

where R_{T_1} = total series resistance (at f_o) when the bandwidth is least ($= R + Z'_s + Z_L$; in which Z_L has its smallest value)

and R_{T_2} = total series resistance (at f_o) when the bandwidth is greatest ($= R + Z'_s + Z_L$; in which Z_L now takes up its largest value).

To determine the voltage source impedance (Z'_s) for the usual condition where Z'_s is not zero, we use

$$Z'_s = \frac{R_{T_2}\,(R_{T_1} - R) - \alpha\, R_{T_1}\,(R_{T_2} - R)}{R_{T_2} - \alpha\, R_{T_1}} \quad (8)$$

Which, for convenience, is rewritten in terms of eqn. (7) as

$$Z'_s = \frac{\alpha_o - \alpha}{\dfrac{\alpha_o}{R_{T_1} - R} - \dfrac{\alpha}{R_{T_2} - R}} \quad . \quad . \quad (8A)$$

where R_{T_1} and R_{T_2} have the values given immediately above,

R = equivalent series resistance of crystal

and $\alpha = \dfrac{A_{min}}{A_{max}}$ = gain variation desired, and must always be less than one.

If a value is selected for Z'_s, then

$$\alpha = \frac{R_{T_2}\,(R_{T_1} - R) - Z'_s\, R_{T_2}}{R_{T_1}\,(R_{T_2} - R) - R_{T_1}\, Z'_s} \quad (9)$$

(D) Selectivity.

Selectivity around resonance can be calculated by the methods to be outlined, but for most purposes it is sufficient to determine the bandwidth for a given attenuation at the half power points (3 db attenuation or $1/\sqrt{2}$ of the maximum voltage output) on each of the required selectivity curves.

Considerable simplification in the design procedure is possible, if the bandwidths for 1 db attenuation (or less) are known. In this case it would be sufficiently accurate to take Z_L and Z_s equal to their dynamic resistances at resonance, at least for most practical conditions, without introducing appreciable error. The advantage obtained being that, in what follows, $R_{T_1} = R_{T_1}$ and $R_{T_2} = R_{T_2}'$ for all conditions. The value of Q_x would be given by

$$Q_x = \frac{f_o}{4\Delta f}$$ for 1 db attenuation, and this expression would be used in place of eqn. (11). However, the procedure given is more general and there should be little difficulty in applying the simplified procedure, if necessary. Further, the method given illustrates (in reverse) how the bandwidths near resonance can be calculated for various amounts of attenuation.

To determine the bandwidth at the half power points proceed as follows: first, it may be taken that the equivalent inductive reactance of the crystal is very much greater than the inductive reactances of the load and source impedances. Also, provided the variations in gain are not allowed to become excessive, at and near resonance the value of Z'_s is very closely R'_s. For very narrow bandwidths, near resonance, it is also sufficiently close to take $Z_L = R_L$; for large bandwidths the resistive component (R_L) of Z_L at the actual working frequency will have to be found. From these conditions, we have

$$Q_x = \frac{\omega_0 L}{R_L + R'_s + R} = \frac{\omega_0 L}{R_T'} = \frac{1}{\omega_0 C R_T'} \quad (10)$$

where Q_x = the equivalent Q of the circuit of Fig. 5 (B)
L = equivalent inductance of the crystal
C = equivalent series capacitance of the crystal
R = equivalent series resistance of the crystal
R_L = resistive component of Z_L, the load impedance, at the frequency being considered ($= Z_L$ at resonance)
R'_s = resistive component of Z'_s the voltage source impedance
R_T' = $R_L + R'_s + R$ = total series resistance at frequencies away from f_o (for very narrow bandwidths $R_L \cong Z_L$ and $R_T' \cong R_{T_1}$)

and $\omega_0 = 2\pi \times f_o$ (where f_o is the i-f).

From the principles of series resonant circuits we know that the total bandwidth ($2\Delta f$) for the half power points (3 db att.) is given by

$$2\Delta f = \frac{f_o}{Q_x} \quad . \; . \; . \; . \; . \; . \; . \; . \quad (11)$$

It follows from eqns. (11) and (10) that for large bandwidths Q_x should be small and so R'_T should be large. For narrow bandwidths the reverse is true, and the limiting case for the narrowest bandwidth would be when $R'_T = R$ (i.e. for the crystal alone); a condition impossible to achieve in practice because of circuit requirements.

Since $Z_L \gg (Z'_s + R)$ for constancy of maximum gain at f_o, under conditions of varying bandwidth, it also follows that Q_x will be mainly determined by the magnitude of the resistive component of the dynamic impedance of Z_L.

Some additional selectivity is given by the input i-f transformer, and for more exact results the attenuation for a particular bandwidth would be added to that found for the crystal circuit. This additional selectivity is usually negligible around the "nose" of the resonance curve, but can be determined from the universal resonance curves given in Radiotronics 131 (p. 44) evaluating D and b for unequal primary and secondary Q's (Q_1 and Q_2) and noting that Q in these expressions is Q_2.

The resonance curve for the complete circuit is seldom necessary for a preliminary design. The procedure is rather lengthy but not very difficult. First it is necessary to find the resistive component of Z_L (i. e. R_L) at frequencies off resonance. This is carried out from a knowledge of $|Z_L|$ ($= Z_L$) at resonance, and by determining the reduction factors from the universal selectivity curves. Multiplying $|Z_L|$ by the indicated attenuation factors gives the required magnitudes of load impedance $|Z_L'|$. To find the resistive component, values of θ corresponding to the various bandwiths (and values of $|Z_L'|$ are read from the universal phase-shift curves given in Radiotronics 131 (P. 44). Then

$$R_L = |Z_L'| \cos \theta \quad . \; . \; . \; . \; . \; . \quad (12)$$

The resistive component of Z_s' can be found in a similar way from

$$R'_s = |Z'_s| \cos \phi$$

but since the impedance of Z'_s is generally small, and the circuit relatively unselective, R'_s can often be taken as equal to Z'_s, for a limited range of frequencies near resonance. When Z'_s is very low, it can usually be neglected in comparison with $R_L + R$.

Then, since the resistance (R) and the inductance (L) of the crystal are known, eqn (10) can be applied to determine the values of Q_x corresponding to the various bandwidths. (Strictly f_o would be replaced by the actual operating frequency off resonance, but this is hardly necessary.)

Knowing the different values of Q_x, and since the bandwidths corresponding to each Q_x value are known, the various points for the complete selectivity curve can be found from the universal resonance curve for a single tuned circuit (it is unimportant that this is for a parallel tuned rather than a series tuned circuit, since the current-voltage relationship is reasonably linear over the range of frequencies near resonance for the circuit Q's involved). Additional selectivity due to the input transformer can be taken into account if this degree of accuracy is thought to be necessary.

(E) Crystal constants.

Before a complete design can be carried out, some data on the equivalent electrical constants of the

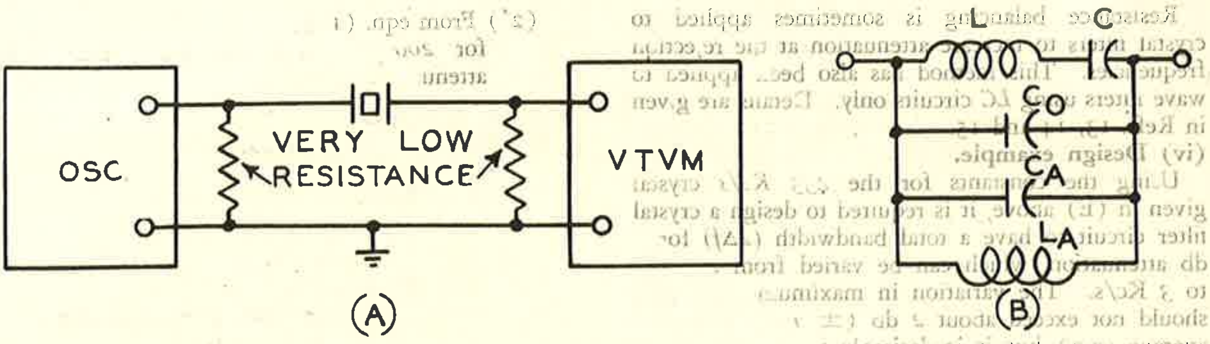


FIG. 7 APPROXIMATE METHODS FOR DETERMINING EQUIVALENT ELECTRICAL CONSTANTS OF QUARTZ CRYSTALS

quartz crystal must be available. In most cases details can be obtained from the crystal manufacturer.

Typical values for the electrical constants of *455 Kc/s* quartz crystals widely used in Australian Communications receivers are:

$R = 1500\ \Omega$; $C = 0.018\ \mu\mu F$; $C_o = 17\ \mu\mu F$.

The crystals are a special type of X-cut bar and have no subsidiary resonances for a range of at least ± *30* Kc/s from f_r. They are approximately *20* mils thick, $\frac{1}{4}''$ wide, $\frac{5}{8}''$ long and are mounted between flat electrodes with an air gap not greater than *1* mil.

If the required data cannot be obtained, then details of methods for measuring, first, the value of Q for the crystal, can be found in Ref. 19. The experimental set-up is shown in Fig. 7 (A) and Q is found from

$$Q = \frac{f_r}{2\,(f_p - f_r)}\sqrt{\frac{E_r}{E_p}} \quad . \quad . \quad (13)$$

where f_r = series resonance frequency
f_p = parallel resonance (or antiresonance) frequency
E_r = voltage across terminating resistance at series resonance
and E_p = voltage across terminating resistance at parallel resonance.

A knowledge of Q will allow the filter circuit to be designed, but if values for C and C_o are required these can be readily measured (Ref. 18) by using the arrangement of Fig. 7 (B) and the relations

$$\frac{C}{C_A + C_o} = \left[\frac{f_{p_2} - f_{p_1}}{f_r}\right]^2 \quad . \quad . \quad (14)$$

and

$$f_{pA} = \frac{1}{2\pi\sqrt{L_A C_A}} \quad . \quad . \quad . \quad (15)$$

where C = series capacitance of crystal
C_A = parallel capacitance of circuit $L_A C_A$
L_A = parallel inductance of circuit $L_A C_A$
C_o = shunt capacitance across crystal due to holder, etc.
f_{p_1} and f_{p_2} = parallel resonance frequencies of combined circuits. The combination has two parallel resonance frequencies, one above and one below f_r.
f_r = series resonance frequency. Measured with the combined circuit.
and f_{pA} = parallel resonance frequency of the circuit $L_A C_A$ alone.

C_o is determined by disconnecting the crystal and adding capacitance to $L_A C_A$ to retune the circuit to the frequency f_r.

(F) Position of filter in receiver.

In most receivers using crystal filters at least two i-f stages are included, since the gain of the crystal stage is often well below that for a normal i-f stage. Also, since threshold effects are not unknown with crystals (this is probably due to the mounting of the crystal in the holder) and since low stage gain immediately after the converter valve may have an adverse effect on signal-to-noise ratio, it is preferable to place the filter between the first and second i-f voltage amplifier valves.

It would not be satisfactory to place the filter between the last i-f valve and the detector because of the low impedance of the load in this case.

Since it is common practice to incorporate at least one r-f stage in receivers of the type mentioned, it is unlikely that any of the effects mentioned would be experienced. However, good design would suggest at least two i-f stages plus one or more r-f stages depending on circuit requirements.

(G) Other types of crystal filters.

There are several types of crystal filters suitable for use in radio receivers other than the simple bridge circuit to which attention has been confined. Typical examples are double-crystal and bridged T filters. These circuits are characterized by having two rejection frequencies usually placed with geometric symmetry about f_r. Details can be found in Refs. 3, 8, 9, and 15.

Resistance balancing is sometimes applied to crystal filters to increase attenuation at the rejection frequencies. This method has also been applied to wave filters using *LC* circuits only. Details are given in Refs. 13, 14 and 15.

(iv) Design example.

Using the constants for the *455 Kc/s* crystal given in (E) above, it is required to design a crystal filter circuit to have a total bandwidth $(2\Delta f)$ for *3* db attenuation, which can be varied from *200* c/s to *3* Kc/s. The variation in maximum stage gain should not exceed about *2* db (± 1 db about the average gain) but it is desirable to keep the stage gain as high as possible, consistent with stable operation with varying signal input voltages (i.e. large detuning of the i-f circuits should not occur when the signal voltages vary over a wide range).

To make the problem complete, it will be assumed that the filter is connected between two type *6SK7* pentode voltage amplifier valves. The complete circuit is shown in Fig. 8.

(*1'*) Since $Z_1 = Z_2$ and $C_o \cong 17\ \mu\mu F$, let us select a suitable capacitance range for C_N. The smallest residual capacitance for C_N will be about $3 \mu\mu F$. From this, we have to increase C_N a further $14\ \mu\mu F$ to neutralize C_o. In addition, it is desired to move the rejection frequency (f_p) below f_r, so that it is reasonable to allow C_N to increase at least a further $14\ \mu\mu F$. The total increment in C_N is thus $28\ \mu\mu F$; which gives a range of *3* to $31\ \mu\mu F$; for convenience this is made, say, *3* to $35\ \mu\mu F$ (or whatever is the nearest standard capacitance range).

(2') From eqn. (11):
for *200* c/s total bandwidth, and *3* db attenuation,

$$Q_{x_1} = \frac{455}{0.2} = 2275;$$

for *3* Kc/s,

$$Q_{x_2} = \frac{455}{3} = 151.6.$$

(*3'*) From eqn. (10) (and $C = 0.018\ \mu\mu F$)

$$R'_{T_1} = \frac{10^{12}}{2275 \times 2\pi \times 455 \times 10^3 \times 0.018} = 8{,}550\Omega$$

$$\text{and } R'_{T_2} = \frac{8550 \times 2275}{151.6} = 0.128\ M\Omega$$

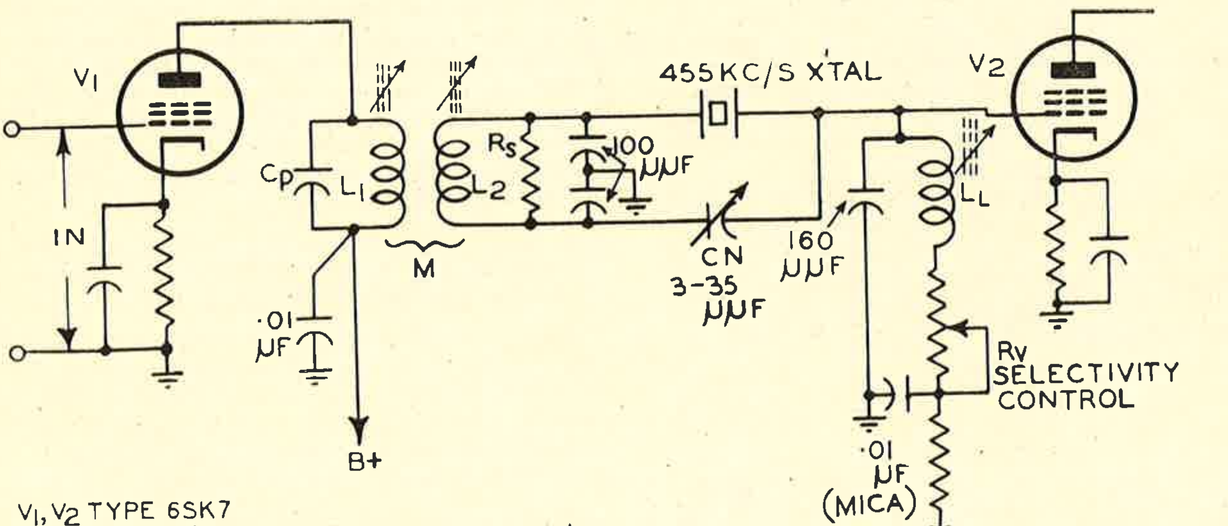
So that (since $R = 1500\Omega$)

$$R_{L_1} + Z'_s = 8550 - 1500 = 7050\Omega$$

$$R_{L_2} + Z'_s = 0.128 - 15 \times 10^{-3} = 0.126 M\Omega.$$

For narrow bands of frequencies $R_{L_1} \cong Z_{L_1}$, and we may write $R'_{T_1} = R_{T_1}$ and so $Z_{L_1} + Z'_s \cong 7050\Omega$.

(4') To find Z_{L_2}. It should be clear, from the values just given (Z'_s remains unchanged) that for fairly large bandwidths $R_{L_2} \gg Z'_s$.



V1, V2 TYPE 6SK7
Rv=0–408Ω (FOR BANDWIDTHS STATED)

LL=0·586 mH	QL=140
L1=4·9 mH	Q1=140
L2=0·32 mH	Q2=140
kc=0·234	Rs=2460Ω

Cp=25μμF (INCLUDING STRAYS)

FIG. 8 CRYSTAL FILTER CIRCUIT DESIGNED FROM ILLUSTRATIVE EXAMPLE

Assume that, for the load circuit, $Q_L = 140$. Then by calculation, or from the universal resonance curves, we have for a single tuned circuit, and a frequency of 455 Kc/s,

$$\frac{Z_{L_2} \text{ at resonance}}{|Z'_{L_2}| \text{ at 3 Kc/s bandwidth}} = 1.37.$$

Also, the phase shift $\theta = 42°43'$.

Using these factors in conjunction with eqn. (12),

$$|Z'_{L_2}| = \frac{R_L}{\cos\theta} = \frac{0.126}{0.735}$$

and so $$Z_{L_2} = \frac{1.37 \times 0.126}{0.735} = 0.235 \text{ M}\Omega.$$

(If the maximum bandwidth required is too large it will be found that Z_{L_2} cannot be obtained with ordinary circuit components.)

From this, since $Z_{L_2} = Q_L \omega_0 L_L$,

$$L_L = \frac{0.235 \times 10^3}{140 \times 2\pi \times 0.455} = 0.586 \text{ mH.}$$

and $$C_L = \frac{25330}{0.455^2 \times 586} = 208 \ \mu\mu\text{F.}$$

(5′) If it is possible to make Z_L higher than 0.235 megohm then a voltage step up is possible using the arrangement of Fig. 9. This is not done here because it is considered that good circuit stability (which requires a large value for C_L) is more important, since the circuit bandwidth, and gain, is very critical to detuning. The main causes of detuning are capacitance changes at the input of V_2 (brought about as a result of alteration in a.v.c. bias voltage) and resetting of C_N; although this latter capacitance change is offset, to some extent, in the alternative circuit since it appears across only part of the tuned circuit capacitance.

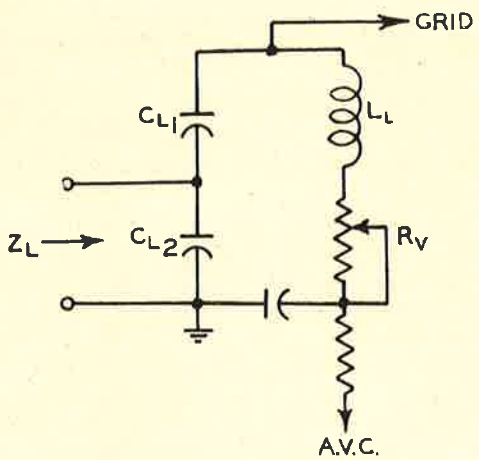


FIG. 9 ARRANGEMENT FOR INCREASING GAIN OF CRYSTAL FILTER STAGE

Any of the usual methods of reducing detuning effects could be applied here to offset changes in valve input-capacitance. Connecting the grid to a tap on the tuned load is an obvious means of reducing the effects of valve input capacitance variation. This method involves a loss in gain, but this is not necessarily serious, as in some receivers the total gain is more than can usefully be employed.

The actual fixed capacitance for C_L is found approximately as follows:

Valve input capacitance	=	6.5 $\mu\mu$F
Strays across coil + wiring, etc.	=	8.5 $\mu\mu$F
$C_N + C_o = 2C_o = 2 \times 17$	=	34 $\mu\mu$F (when C_o is neutralized by C_N)
Total Strays	=	49 $\mu\mu$F

Fixed capacitance req. = 208–49 = 159 $\mu\mu$F (say 160 $\mu\mu$F).

(6′) From eqn. (7); and since $R_{T_1} = 8550\Omega$ and

$R_{T_2} = 0.237$ MΩ (near enough; i.e. $Z_{L_2} + R$),

$$\alpha_o = \frac{0.237 \ (7050)}{8550 \ (0.235)} = 0.83$$

(which shows that the gain variation is practicable). Then, since α = 2 db down = 0.794, we have from eqn. (8A)

$$Z'_s = \frac{0.83 - 0.794}{\dfrac{0.83}{7050} - \dfrac{0.794}{0.235 \times 10^6}}$$

$$\simeq \frac{0.036 \times 7050}{0.83} = 306\Omega.$$

So that $Z_s = 4Z'_s = 1224\ \Omega$.

(7′) To determine the approximate range of the selectivity control R_v. For maximum bandwidth $R_v = 0$. For minimum bandwidth, since

$$Z_{L_1} \ (= R_{L_1}) = 7050 - 306 = 6744\ \Omega,$$

$$\text{then } Q_{L_1} = \frac{Z_{L_1}}{\omega_0 L_L} = \frac{6744}{1680} = 4$$

(This is only approximately correct: the error in R'_v *is about 3%.*)

$$\text{and } R'_v = \frac{\omega_0 L_L}{Q_L} = \frac{1680}{4} = 420\Omega.$$

Resistance already in circuit when Q is 140, is

$$R_L = \frac{1680}{140} = 12\Omega.$$

Range of R_v required is from 0 — 408Ω.

This is the range if R_v is connected in series with L_L. If R_v is to be connected in series with the capacitive arm of Z_L it should be realized that it would actually be in series with the fixed capacitance of 160 $\mu\mu$F only, and not the total capacitance, so that new values for the range of R_v would have to be calculated in this case.

(8′) The design of the input transformer is the next step. For maximum gain the transformer will be critically coupled.

Consider first the secondary circuit. The total secondary capacitance, if we select 100 μμF capacitors (connected in series) for the ratio arms of the bridge circuit, will be $\frac{100}{2} + 14.5 = 64.5$ μμF. The 14.5 μμF represents the approximate total for

$$\frac{C_N}{4} + \frac{C_o}{4} = \frac{C_o}{2}$$

(since $C_N = C_o$) = 8.5 μμF plus an allowance of 6 μμF for coil and circuit strays.

Then the apparent secondary inductance required

$$L_2 = \frac{25.33}{0.455^2 \times 64.5} = 1.9 \text{ mH.}$$

From step (6′) above $Z_s = 1224\Omega$. Since critical coupling will halve the actual value of Q_2, we must use an uncoupled value for Z_s of 2448Ω. This allows us to determine the uncoupled secondary magnification factor

$$Q_2 = \frac{2448}{\omega_o L_2} = \frac{2448}{5440} = 0.45.$$

Because of the low value of Q, the condition of $\omega^2 L_2 C_2 = 1$ is no longer sufficiently accurate. For cases such as this, where Q_2 is less than about 10, proceed exactly as before but modify the value of L_2 by a factor $Q^2/(1+Q^2)$. This gives the true condition for resonance (unity power factor) if Q is assumed constant and L_2 is variable, as it will be in most i-f transformers of the type being considered. The actual value required for L_2 is now

$$L_2 \text{ (actual)} = 1.9 \left[\frac{0.45^2}{1+0.45^2}\right] = 0.32 \text{mH.}$$

(9′) Since the secondary Q is very low, and L_2 is fixed by other considerations, it should be clear that if we require reasonably high stage gain that the primary Q ($=Q_1$) and the primary inductance L_1 should be as high as possible; since this will allow $QL = \sqrt{Q_1 Q_2 L_1 L_2}$ to be increased.

As the minimum capacitance across L_1 will be about 25 μμF (valve output + strays: which will be fairly high for a large winding), then

$$L_1 = \frac{25.33}{0.455^2 \times 25} = 4.9 \text{ mH.}$$

Of course, it may not be advisable to resonate the primary with stray capacitance only, but there are practically no detuning effects present (except those due to temperature and humidity variations). The gain lost in making L_1 equal L_2 is not very large, since a higher value of Q_1 is possible; however, L_1 and L_2 will be made unequal here to illustrate the procedure to be adopted.

(10′) Assume that unloaded Q values of 140, for the primary and secondary, can be obtained. The plate resistance (r_p) for the type 6SK7 is 0.8 MΩ for a set of typical operating conditions. Then the actual value of Q_1 is

$$Q_1 = \frac{Q_u\, r_p}{Q_u\, \omega_o L_1 + r_p} = \frac{140 \times 0.8}{1.96 + 0.8} = 40.6.$$

(11′) In order that the actual value of Z_s shall be 1224Ω (and so $Z_s' = 306\Omega$) the secondary circuit must be loaded with a resistance R_s given by

$$R_s = \frac{Q_u\, Q_2 \omega_o L_2}{Q_u - Q_2} = \frac{140 \times 2448}{140 - 0.45} = 2460\Omega.$$

where L_2 has the apparent value of 1.9mH.

$$(12') \quad Q = \sqrt{Q_1 Q_2} = \sqrt{40.6 \times 0.45} = 4.28.$$

The coefficient of coupling $= k_c = \frac{1}{4.28} = 0.234.$

$$L = \sqrt{L_1 L_2} = \sqrt{4.9 \times 0.32} = 1.25 \text{mH.}$$

The actual value of L_2 is required here.

The most satisfactory method for setting the value of k required is the open circuit/short circuit method described in Radiotronics No. 131 pages 50 and 57 (appendix 3).

(13′) For the valve types selected $g_m = 2{,}000$ μmhos (2mA/volt). Gain from grid V_1 to grid V_2, using eqn. (6), is

Gain (max. bandwidth) =

$$\left[\frac{0.235}{0.235 + 306 \times 10^{-6} + 1500 \times 10^{-6}}\right] \times \left[\frac{2 \times 4.28 \times 2\pi \times 0.455 \times 3.06}{4}\right]$$

$= 0.992 \times 7.7 = 7.6$ times (17.6 db).

Gain (min. bandwidth) $= \frac{6744}{8550} \times 7.7 = 6.07$ times (15.6 db).

The expression just used to determine stage gain (eqn. 6) is only correct when $L_1C_1 = L_2C_2$; a condition not fulfilled here, because the circuit constants for resonance when Q_2 is less than 10 does not call for XL_2 to be equal to XC_2. For cases of this type it is sufficiently accurate, for most practical purposes, to multiply eqn. 6 by the ratio XC_2/XL_2 (the error in the worst practical case likely to be encountered does not exceed about +20%). From this it follows that the gain figures determined above should be multiplied by *5.95* in each case (since $X_{C_2} = 5470$ Ω and $X_L = 917$ Ω). The actual stage gain is thus *45* times (*33* db) for maximum bandwidth, and *36* times (*31* db) for minimum bandwidth. If the exact stage gain is required then the complete equivalent circuit should be used (in our case the gain so determined is *38* times).

So that the maximum total gain variation (including all approximations) is about *2* db as specified i.e. ± *1* db about the average gain. The alternative circuit of Fig. 9 would allow the maximum gain to be increased to about 73 times. Larger variations in gain, if permissible, would also allow increased overall gain; the disadvantages have been discussed previously.

(14') For many purposes a standard i-f transformer using fixed capacitances of about *50* μμF, and the coupling increased to critical when the secondary and primary are correctly loaded, would give satisfactory results. The gain variation, and maximum gain, is largely controlled by the value of the damping resistor R_s connected across the secondary of the transformer.

To control bandwidth it is necessary to increase Z_L to increase the maximum bandwidth; R_v is increased to decrease the minimum bandwidth. If switched steps are required for bandwidth control, R_v can be calculated for each step; the remainder of the design is exactly as before.

The arrangement used for the input transformer is only one of many possible circuits. For cases such as the one given here (where the total gain is not a prime requirement) it may be preferable to leave the secondary circuit untuned and to use a tapped resistor to provide the ratio arms for the bridge circuit. The design procedure is readily developed fom the usual coupled circuit theory.

References

(1) Zepler, E. E., "The Technique of Radio Design" (p. 133), Chapman & Hall, London, 1943; Wiley & Sons, New York, 1943.

(2) Sturley, K. R., "Single Crystal Filters", Wireless Engineer 22.262 (July, 1945), 322.

(3) "The Amateur Radio Handbook" (2nd edit.), The Incorp. Radio Society of Gt. Britain, London.

(4) "The Radio Amateur's Handbook" (24th edit.), American Radio Relay League Inc., 1947.

(5) Oram, D. R., "Full-Range Selectivity with 455 Kc/s Quartz Crystal Filters", Q.S.T., 22.12 (Dec., 1938), 33.

(6) Erlich, R. W., "The Crystal Filter", Radio Craft, 17.6 (March, 1946), 398, Radio Craft 17.7 (April, 1946), 476.

(7) Colebrook, F. M., "High Selectivity Tone—Corrected Receiving Circuits", Brit. Radio Research Board Special Report No. 12, 1932 (Reprinted 1934).

(8) Builder, G., and Benson, J. E., "Simple Quartz Crystal Filters of Variable Bandwidth", A.W.A. Tech. Review 5.3 (Feb., 1941), 93.

(9) Mason, W. P., "Electrical Wave Filters Employing Quartz Crystals as Elements", B.S.T.J., 13.3 (July, 1934), 405.

(10) Robinson, J., "The Stenode Radiostat", Radio News, 12.8 (Feb., 1931), 682.

(11) Stanesby, H., "A Simple Narrow Band Crystal Filter", British P.O.E.E.J., Vol. 28, Pt. 1 (April, 1942), 4.

(12) Stanesby, Broad and Corke, "Calculation of Insertion Loss and Phase Change of 4—Terminal Reactance Networks", British P.O.E.E.J., Vol. 25, Pt. 3 (Oct., 1942), 88, Vol. 25, Pt. 4 (Jan., 1943), 105. (Gives transfer impedance, etc., for equivalent bridge circuit.)

(13) Mason, W. P., "Resistance Compensated Band-Pass Filters for Use in Unbalanced Circuits", B.S.T.J., 16.4 (Oct., 1937), 423.

(14) Builder, G., "Resistance Balancing in Wave Filters", A.W.A. Tech. Review, 3.3 (Jan., 1938), 83.

(15) Starr, A. T., "Electric Circuits and Wave Filters" (Textbook; 2nd edit.), Sir Isaac Pitman & Sons, London, 1940.

(16) Booth, C. F., "The Application and Use of Quartz Crystals in Telecommunications", Jour., I.E.E., Part 3, 88.2 (June, 1941), 97.

(17) Grisdale, G. L., and Armstrong, R. B., "Tendencies in the Design of the Communication Type of Receiver", Jour. I.E.E., Part 3, 93.25 (Sept., 1946), 365.

(18) Builder, G., "A Note on the Determination of the Equivalent Electrical Constants of a Quartz Crystal Resonator", A.W.A. Tech. Review, 5.1 (April, 1940), 41.

(19) Mason, W. P., and Fair, I. E., "A New Direct Crystal-Controlled Oscillator for Ultra-Short-Wave Frequencies", Proc. I.R.E., 30.10 (Oct., 1942), 464.

(20) Sandel, B., "Design of I-F Transformers", Radiotronics No. 131 (May/June, 1948), 43.

Triode Operation of 6AU6

If type 6AU6 is connected as a triode (screen connected to plate), the suppressor should be connected to the cathode and not to the plate.

CORRECTION RADIOTRONICS No. 131.

The expression for b^2 given in Fig. 9.17 (page 44), Universal Selectivity Curves for Two Coupled Circuits, should have the denominator squared. The equation should be identical with that for b^2 given directly below in Fig. 9.18.

✓ Noted.

REVISED PHOTOTUBE CHARACTERISTICS

A number of the commonly used R.C.A. phototube types have been registered with the R.M.A. in U.S.A. Since some of the characteristics listed differ slightly from those shown on previously issued data sheets, therevised data is shown below.

PHOTOTUBES

SINGLE- AND TWIN-UNIT TYPES

SPECTRAL CHARACTERISTICS				MAXIMUM RATINGS									
		MAX. RESPONSE VALUES								LUMINOUS SENSITIVITY u AMP./LUMEN			
COLOR	CURVE	WAVE-LENGTH ANG-STROMS	SENSI-TIVITY uAMP/uWATT	ANODE-SUPPLY VOLTS DC or PEAK AC	PEAK CATHODE CURRENT uAMP.	PEAK CATHODE CURRENT DENSITY uAMP/SQ.IN.	AVERAGE CATHODE CURRENT + uAMP.	AMBIENT TEMPER-ATURE °C	MAX. GAS AMPLIFI-CATION FACTOR	0 CYCLES	5000 CYCLES	10000 CYCLES	TYPE
													GAS TYPES
Red-Infrared	S-1	8000	0.009	100	20	100	5♦	100	8	90	77	67	**868**
Red-Infrared	S-1	8000	0.015	90	20	100	5♦	100	10.5	150	120	105	**918**
Red-Infrared	S-1	8000♠	0.010♠	90♠	6♠	50♠	2♦♠	100♠	9♠	100♠	85♠	74♠	**920**
Red-Infrared	S-1	8000	0.0135	90	10	100	3♦	100	10	135	119	108	**921**
Red-Infrared	S-1	8000	0.0135	90	10	100	3♦	100	10	135	111	101	**923**
Red-Infrared	S-1	8000	0.009	90	5	75	1.5♦	100	8.5	90	77	67	**924**
Red-Infrared	S-1	8000	0.0125	90	6	100	2♦	100	10	125	110	100	**927**
Red-Infrared	S-1	8000	0.0065	90	10	100	3♦	100	10	65	56	50	**928**
Red-Infrared	S-1	8000	0.0135	90	10	100	3♦	100	10	135	111	101	**930**
													VACUUM TYPES
Red-Infrared	S-1	8000	0.002	500	30	100	10	100	–	20	20	20	**917**
Red-Infrared	S-1			For additional data, refer to type 917.									**919**
Red-Infrared	S-1	8000	0.002	500	15	100	5	100	–	20	20	20	**922**
Red-Infrared	S-1	8000	0.0015	250	15	100	5	100	–	20	20	20	**925**
Violet-Green	S-3	4200	0.0016	500	15	100	5	100	–	6.5	6.5	6.5	**926**
Blue	S-4	4000	0.042	250	20	100	5	75	–	45	45	45	**929**
Blue	S-4	4000	0.028	250	12	100	4	75	–	30	30	30	**934**
Ultraviolet-Blue	S-5	3400	0.028	250	30	100	10	75	–	30	30	30	**935**

♦ At maximum anode-supply volts. For operation with anode-supply 20 volts less than the maximum rating, the average cathode current can be doubled.

♠ Each unit. + Averaged over any interval of 30 seconds maximum.

Receiver Sensitivity and Gain Measurements at High Frequencies

R.C.A. Application Note AN-132 reprinted by courtesy of the Radio Corporation of America.

Methods of measuring the sensitivity and gain of rf amplifiers and converters of receiving systems operating in the FM broadcast band or in the television bands are described in this Note. When conventional voltage-input methods used in the standard AM broadcast band are applied at high frequencies, difficulties result because the input voltage required to produce a given output is dependent on the point of input to the circuit being measured. The methods described in this Note, however, are based on the power input to a circuit rather than on the voltage input and are advantageous because the power input required to produce a given output is independent of the point of input.

General Considerations.

When a circuit such as that of Fig. 1 is part of a receiver, sensitivity measurements are conventionally made by connecting a standard signal generator supplying a modulated signal to terminals (1-1) through a specified "dummy antenna" network, and then adjusting the signal to produce a specified standard output from the receiver. In a low-frequency receiver, it is common practice to obtain additional data by connecting the signal generator successively to points (4-4), (3-3), and (2-2) through a low-impedance blocking capacitor. The frequency and voltage of the signal generator are adjusted for each test point to give the standard receiver output.

The voltage input at the intermediate frequency required at terminals (4-4) to give the standard output may be described as the voltage sensitivity of the receiver at the first if grid. Similarly, the inputs at the signal frequency required at points (3-3) and (2-2) may be described as the voltage sensitivities at the converter grid and at the rf grid, respectively. The ratio of the required input at (4-4) to the required input at (3-3) is the conversion voltage gain from converter grid to if grid provided that the receiver is nearly free of feedback effects. The ratio of the required input at (2-2) to the required input through the dummy antenna to (1-1) is frequently referred to as the antenna circuit gain, but it must be understood that the dummy antenna is considered as part of the antenna circuit for this definition.

High-Frequency Considerations.

At high frequencies the attempt to make these measurements with the foregoing method leads to erroneous and misleading results. The major difficulty is caused by the substantial reactances of even short pieces of wire at high frequencies. A signal generator is calibrated in terms of the open-circuit voltage across its terminals, but it is physically impossible to bring these terminals exactly to the points at which voltage-sensitivity measurements are desired, even when the terminals are at the end of a flexible cable.

It is possible, however, to introduce a measured amount of power into a circuit of a receiver without encountering similar difficulties. A method of introducing a measured amount of power is illustrated in Fig. 2. In this figure, a resistor R and an adjustable capacitor C are connected between the signal generator and the receiver tuned circuit. Maximum power will be transferred to the tuned circuit when capacitors C and C_1 are adjusted so that the impedance of the circuit between point (a) and ground is resistive and equal to r which is the sum of the added resistance R and the internal resistance of the generator. The method

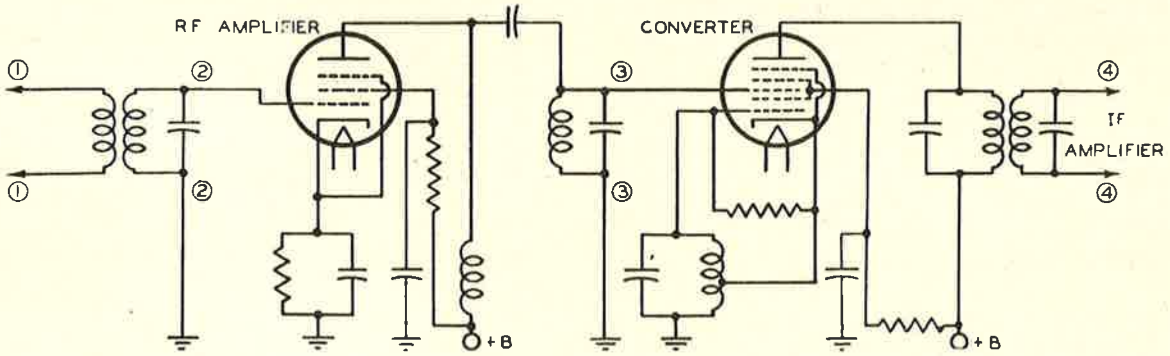


Fig. 1 – Typical RF Amplifier and Converter Circuit.

applies to either circuit of Fig. 2. Although the required capacitor adjustments will be different, the amount of power which can be transferred with a given signal-generator terminal voltage is the same for either circuit. When the adjustments for maximum output have been completed, the available power, P, is equal to the power transferred to the receiver circuit and is given by the equation

$$P = e^2/4r$$

where e is the open-circuit voltage at the generator terminals,

and r is the sum of the added resistance R and the internal resistance of the generator.

In practice, the resistor R is connected to the high-potential terminal of the signal generator, and the adjustable capacitor C is connected between the resistor and a point near the high-potential end of the receiver circuit under consideration. A value for resistor R of approximately 300 ohms has been found suitable for frequencies near 100 megacycles. At other frequencies, however, different resistor values may be more suitable. Two pieces of hook-up wire twisted together may be used for the adjustable capacitor C. The circuit is tuned to resonance with

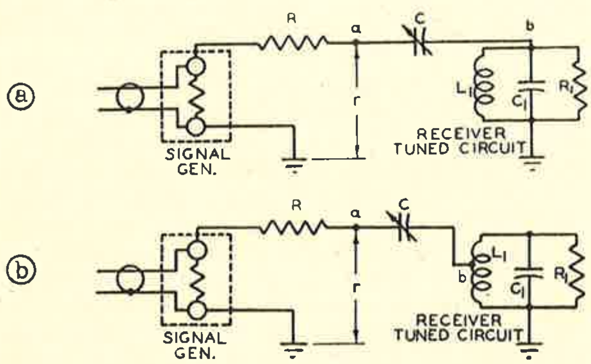


Fig. 2—Connection of Signal Generator to Resonant Circuit for Maximum Power Transfer.

the signal frequency by use of whatever tuning means are provided and the receiver output is noted. Various adjustments of the series capacitance are tried, with readjustment of the receiver circuit to resonance in each instance, until the adjustment giving maximum receiver output is found. The signal-generator voltage is then adjusted to the value giving standard power output from the receiver and this voltage is recorded. The power sensitivity can then be computed from the signal generator voltage and the resistance, r.

Example.

Measurements made on an FM receiver are given as an example of the application of this method. The receiver circuit used is not identical to that shown in Fig. 1, but it corresponds closely enough to permit use of Fig. 1 in the explanation of the data. The signal frequency used was 98 megacycles frequency modulated with 400 cycles. The receiver output was 50 milliwatts. The tubes used were type 12BE6 as a converter and type 6BJ6 as an rf amplifier. The resistor was 260 ohms and the output resistance of the signal generator was 26.5 ohms, giving a total resistance of 286.5 ohms. Since the antenna circuit of the receiver is designed for 300 ohms, this resistor can also be used for the dummy antenna. Connections corresponding to points (2-2) and (3-3) of Fig. 1 were made through a twisted-wire capacitor and connections to (1-1) were made through the resistor only. The measurements are tabulated below.

Point of Input	Signal Generator Output Volts	Available Power P Watts
(3-3)	125×10^{-6}	13.7×10^{-12}
(2-2)	25×10^{-6}	0.46×10^{-12}
(1-1)	25×10^{-6}	0.46×10^{-12}

Power Ratio, (3-3) to (2-2) 29.6

Power Ratio, (2-2) to (1-1) 1.0

The power ratio (3-3) to (2-2) is the effective power gain of the rf amplifier stage. This ratio represents the real advantage in sensitivity obtained by adding the rf stage to the receiver, and, therefore, conveys more significance to the designer than a measurement of grid-to-grid voltage gain.

The power ratio (2-2) to (1-1) indicates the degree of coupling and the efficiency of the antenna transformer. The observed value of unity indicates that a close impedance match was obtained and that the additional losses obtained when the antenna winding is used are negligible within the limits of accuracy of the measurements.

Advantages.

When measurements are made at the input circuit of the converter tube (point (3-3), Fig. 1), an important advantage is obtained by this method because the signal is introduced with only a slight disturbance of the circuit by the measuring equipment. The impedance of the input circuit to the signal frequency is reduced to half its normal value, but the impedance of the circuit to the oscillator frequency changes very little. At high frequencies, the amount of oscillator-frequency voltage induced in the signal-grid circuit is frequently an important factor in determining the performance of the converter tube. Consequently, a method of measurement which does not affect the induced voltage gives a better indication of tube performance than a method in which the signal grid is effectively short-circuited to ground.

The power-measurement method described can also give data concerning the resonant impedance of the circuits to which connections are made. When a resistance match is obtained at point a, (Fig. 2a or 2b), the resistance components from points b to ground for the generator and the circuit are also equal. For the generator, the admittance is

$$Y = \frac{1}{r + \frac{1}{j\omega C}}$$

$$= \frac{j\omega C(1 - rj\omega C)}{1 + r^2\omega^2C^2}$$

$$= \frac{r\omega^2C^2}{1 + r^2\omega^2C^2} + \frac{j\omega C}{1 + r^2\omega^2C^2}$$

The conductive component $\frac{r\omega^2C^2}{1 + r^2\omega^2C^2}$ can be

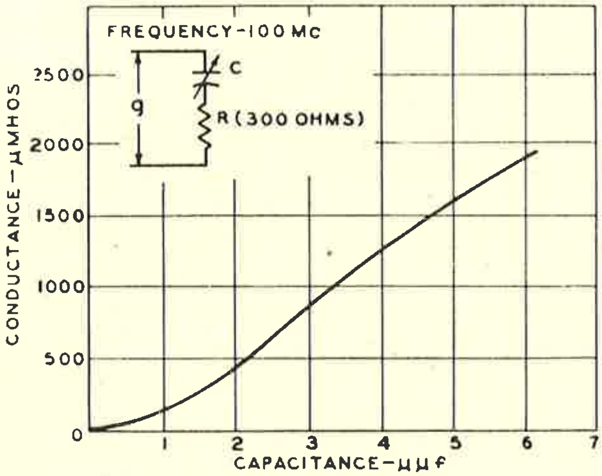


Fig. 3—Variation of Conductance with Capacitance of Series RC Circuit.

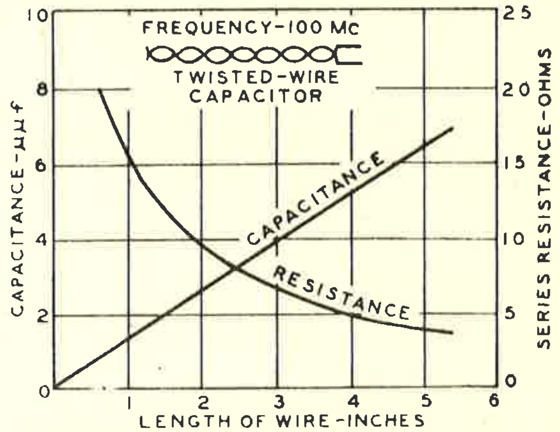


Fig. 4—Capacitance and Effective Series Resistance of Twisted-Wire (Belden 8861) Capacitor.

evaluated when the capacitance is known, and is known, and is equal to the conductance of the circuit measured between the points to which connections are made. The curve of Fig. 3 shows the variation of conductance with capacitance for a resistance of 300 ohms at a frequency of 100 megacycles. It is frequently sufficient in practice to note that an increase in the required capacitance for matching represents an increase in conductance. When a tube or other component is changed, the effect of the change on the circuit impedance can be quickly evaluated by this method.

The capacitance and effective series resistance of a capacitor formed of two pieces of Belden #8861 rf hook-up wire at a frequency of 100 megacycles is given in Fig. 4. Measurements were made on a high-frequency Q meter. The series resistance can usually be disregarded, but it may be measured and taken into account when greater accuracy in power or impedance measurements are desired.

New R.C.A. Releases

Radiotron types 6BA7 and 12BA7—are high gain pentagrid converters of the miniature 9 pin-type designed especially for F-M broadcast service. They perform simultaneously the function of a mixer and of an oscillator in superheterodyne circuits.

These valves, which are identical except for heater rating, feature high conversion gain because of their high conversion transconductance of 950 micromhos with 250 volts on the plate. They have a separate connection for direct grounding of the suppressor.

The types 6BA7 and 12BA7 have characteristics similar to those of the metal type 6SB7-Y.

Radiotron type 5FP7-A—is a high-vacuum cathode ray tube of the magnetic focus and magnetic deflection type having a long-persistence, cascade (two layer) fluorescent screen 4¼" in diameter. It is intended especially for oscillographic applications where a temporary record of electrical phenomena is desired.

During excitation by the electron beam the screen of the 5FP7-A emits blue light (fluorescence). After excitation, the screen exhibits a greenish-yellow phosphorescence which persists for several minutes provided the duration and intensity of the excitation has been adequate and the intensity of the ambient light is not too high.

The type 5FP7-A supersedes the type 5FP7.

Radiotron type 7BP7-A—is a high-vacuum cathode ray tube with a fluorescent screen 6" in diameter. Its general description is similar to that for the type 5FP7-A.

The type 7BP7-A supersedes the type 7BP7.

Radiotron type 12DP7-A—is a high-vacuum cathode ray tube with a fluorescent screen 10" in diameter. The general description is similar to that for the type 5FP7-A.

Type 12DP7-A supersedes type 12DP7.

Equipment Types for 1949

We give below the Radiotron range of Equipment Types recommended for use by receiver and amplifier manufacturers in new equipment. Type U52/5U4-G has been added in place of type 5V4-G for large amplifiers. Type X61M triode hexode has been added as an alternative to type 6J8-GA where high gain is required. Type 6J8-GA has replaced type 6J8-G for an indefinite period. Type 6SN7-GT twin triode has been added to complete the range.

The four miniature 7 pin A.C. types are primarily intended for use in auto. sets and sets incorporating F-M, but may also be used in ordinary A-M receivers. Type 6BE6 is an almost exact equivalent of type 6SA7-GT, which has now been dropped from the list of recommended equipment types. Type 6BA6 may be used in R.F. and I.F. amplifiers as an alternative to type 6SK7-GT; it is particularly valuable as an untuned R.F. amplifier. Type 6AV6 has improved characteristics, but is otherwise an equivalent of type 6B6-G or 6SQ7-GT.

Type 6J7-G/1620 takes the place of the older type 1603 as a non-microphonic amplifier, while type 807 remains in the list as a high power amplifier.

A complete range of AC/DC valves with a heater current of 0.16 ampere has been added. These are all octal based. Printed characteristics are now in course of preparation.

EQUIPMENT TYPES

The following types are recommended for use in new equipment (1949):

1.4. Volt Miniature Battery Range.

1R5	Converter.
1S5	Diode, pentode.
1T4	Remote cut-off R.F. pentode.
3S4	Power amplifier pentode.
3V4	Power amplifier pentode.

2 Volt Battery Range.

1C7-G	Pentagrid converter.
1H4-G	General purpose triode.
1J6-G	Class B twin triode.
1K5-G	R.F. pentode.
1K7-G	Duo-diode, pentode.
1L5-G	Power amplifier pentode.
1M5-G	Remote cut-off R.F. pentode.

Rectifiers.

5Y3-GT	Full wave, directly heated.
6X5-GT	Full wave, indirectly heated.
U52/5U4-G	Full wave, directly heated.

0.16 amp. AC/DC Range (Octal base).

X76M	Triode-hexode converter.
W76	Remote cut-off R.F. pentode.
DH76	Duplex-diode triode.
KT71	Power output tetrode.
U76	Half-wave high-vacuum rectifier.

A.C. Range.

6A8-G	Pentagrid converter.
X61M	High gain triode hexode.
6J8-GA	Triode-heptode converter.
6G8-G	Duo-diode remote cut-off pentode.
6J7-G	R.F. pentode.
6SJ7-GT	R.F. pentode.
6SK7-GT	Remote cut-off R.F. pentode.
6U7-G	Remote cut-off R.F. pentode.
6SN7-GT	Twin triode.
6SQ7-GT	Duo-diode high-mu triode.
6B6-G	Duo-diode high-mu triode.
6V6-GT	Beam power amplifier.

A.C. Miniature Range.

6AU6	R.F. pentode.
6BA6	Remote cut-off R.F. pentode.
6BE6	Pentagrid converter.
6AV6	Duo-diode high-mu triode.

High-Power Amplifier.

807	Beam power amplifier.

Non-Microphonic Amplifier.

6J7-G/1620	Triple-grid amplifier.

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Technical Editor
F. Langford-Smith B.Sc., B.E.

Assistant Technical Editor
B. Sandel, A.S.T.C.

All communications in Australia should be directed to Amalgamated Wireless Valve Co. Pty. Ltd. (Sales Promotion Department), 47 York Street, Sydney, N.S.W.—in New Zealand to Amalgamated Wireless (Australasia) Ltd., P.O. Box 830, Wellington, C.1, N.Z.

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