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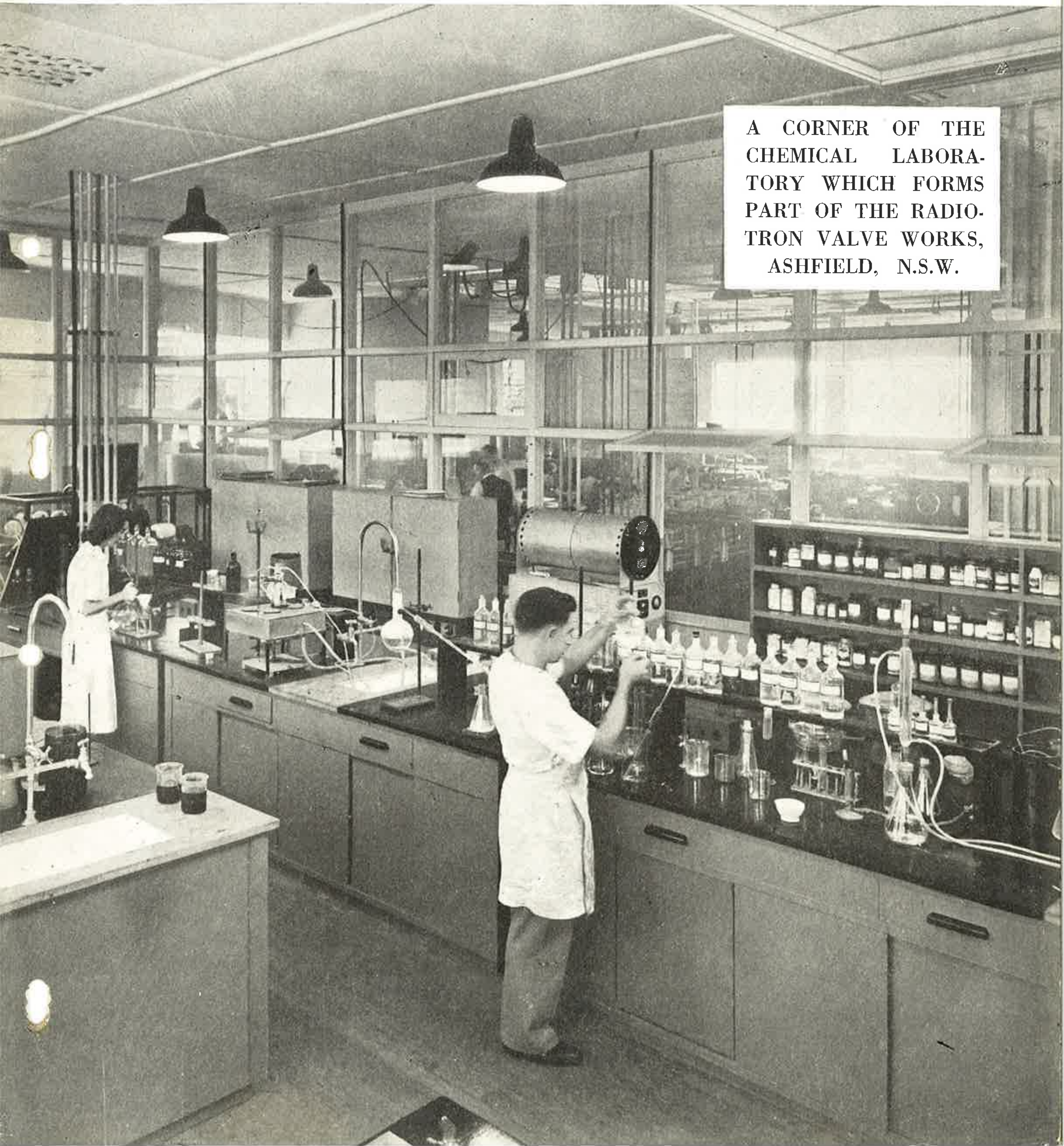

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RADIOTRONICS

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IN THIS ISSUE

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OUR COVER shows portion of the Chemical Laboratory at the Valve Works, the presence of which has proved not only desirable but absolutely essential to modern Valve manufacture. The work consists of the preparation of cathode spray, alundum spray for heaters, carbon spray for plates, and magnesium oxide spray for micas, and the manufacture of material for getters, chemical cleaning and electroplating of valve parts.

Design of I-F Transformers

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

INTRODUCTION

Methods are set out in detail for the design of mutual inductance coupled transformers. It will be shown that by placing a slightly different interpretation on the values of Q and L, obtained by the usual simplified i-f design methods, the design can be extended to give a number of useful practical results. The advantages will be clearly demonstrated in the worked examples, where in some cases the first impression may be that the design conditions cannot be fulfilled with a conventional i-f transformer.

Several appendices are added to show the development of additional formulae and the derivation of the methods of measuring k.

The source of the equations will be stated as the design proceeds and any departure from the original assumptions will be indicated.

This material is part of Chapter 26 of the 4th edition of the Radiotron Designer's Handbook (in preparation). Although references are made to other sections, all the necessary material is included here. The universal selectivity and phase shift curves of Chapter 9, Section 10, referred to in the text, are due to Maynard (Ref. 6).

Section 4:

DESIGN METHODS

(i) General.

(ii) Critically coupled transformers.
- (A) Design equations and table.
- (B) Example.
- (C) Design extension.
- (D) Conclusions.
- (E) k measurement.

(iii) Overcoupled transformers.
- (A) Design equations and table.
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(iv) Undercoupled transformers and single tuned circuits.
- (A) Single tuned circuit equations and table.
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(v) F-M i-f transformers.
- (A) Design Data.
- (B) Example.

(vi) I-F transformer construction.

(i) General.

The design procedure for i-f transformers can be greatly simplified by the use of charts and tables. If certain assumptions are made, which approximate to practical conditions, the design procedure can be reduced to a few routine operations. Here we will consider only the two winding transformer using mutual inductance coupling; the added capacitance coupling, which is always present, does not seriously affect the results particularly as its presence is taken into account when setting the coefficient of coupling (k).

The initial assumptions will be that the primary and secondary inductances L_1 and L_2 are given by $L = \sqrt{L_1 L_2}$. Also, it will be taken that the values of Q do not alter appreciably over the range in which the selectivity curves are taken. We will *not* take the primary and secondary Q's as being equal and the advantages to be gained will become clear as we proceed. The magnification factor, or Q, will be defined as

$$Q = \sqrt{Q_1 Q_2}$$

and provided the ratio of $\frac{Q_1}{Q_2}$ or $\frac{Q_2}{Q_1} \ngtr 2$ the error in the usual design equations is negligible for most practical purposes. As long as

$$\sqrt{Q_1 Q_2} \cong \frac{Q_1 + Q_2}{2}$$

the error in any of the usual approximate design equations will be small. If Q_1 and Q_2 differ by large amounts then the exact design equations are necessary and can be obtained from Refs. 2, 3, 6 and 8, or the design can be modified by using the universal selectivity curves to obtain the required results. It is of interest to note that the simplified equations given by Kelly Johnson (Ref. 1), Ross (Ref. 2) and Maynard (Ref. 6 and Figs. 9.17 and 9.18) are identical when it is assumed that $Q = Q_1 = Q_2$ and the various notations are made the same.

It may be thought that writing $Q = \sqrt{Q_1 Q_2}$ and $L = \sqrt{L_1 L_2}$ will be inconvenient since the i-f transformer, as constructed, will have its primary and secondary inductances and Q's equal. However, in the majority of cases $L = L_1 = L_2$ is applied, and the method is extended to fulfil the condition that the unloaded primary and secondary Q's should be equal while allowing the required $Q = \sqrt{Q_1 Q_2}$ to be obtained in the receiver, without further adjustment.

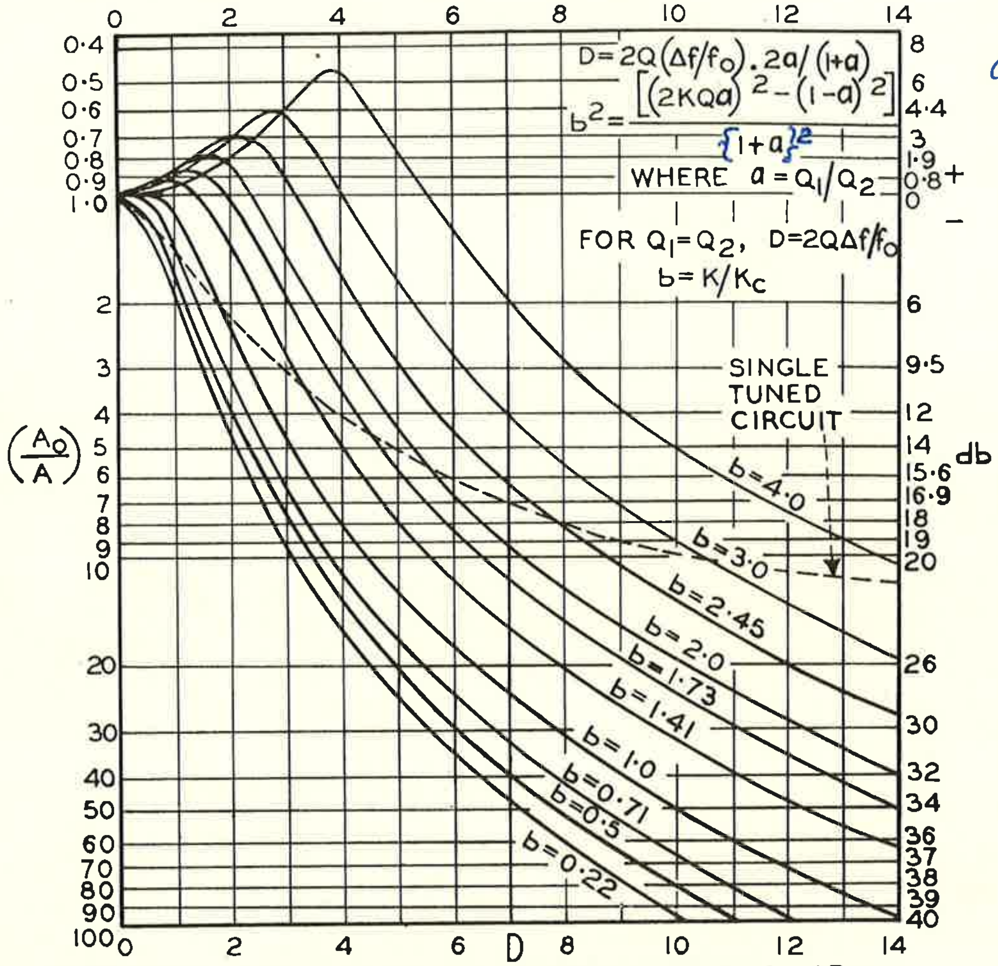

FIG. 9·17 UNIVERSAL SELECTIVITY CURVES FOR TWO COUPLED CIRCUITS

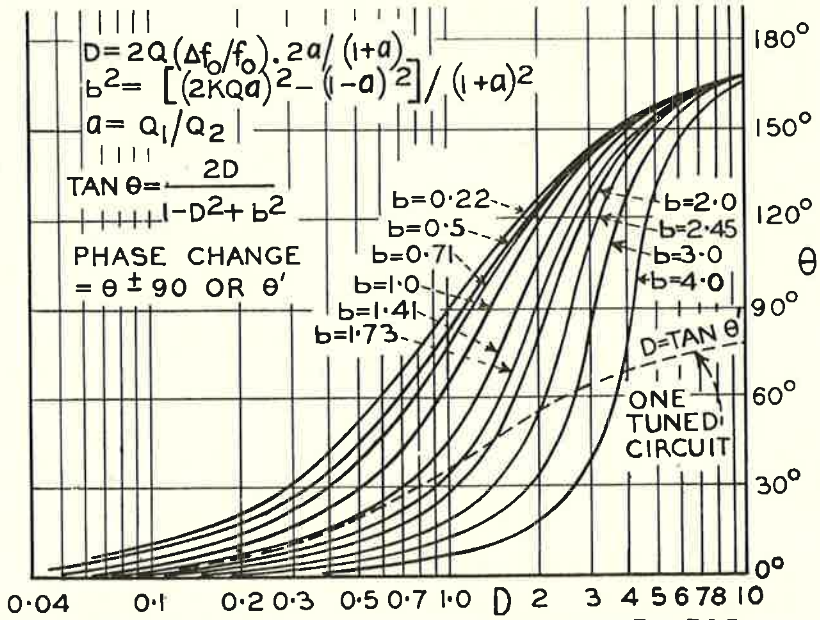

FIG. 9·18 UNIVERSAL PHASE SHIFT CURVES FOR TWO COUPLED CIRCUITS

Critical coupling, or a close approach to it, is most often employed in i-f transformers but there is little difficulty in designing transformers for almost any degree of coupling. All cases will be treated.

Universal selectivity and phase shift curves are given in Chapter 9, Section 10 (figs. 9.17 and 9.18). Additional charts and tables are given, to be used as described in the appropriate sections.

The design procedure generally consists of finding values of Q, k and stage gain for given bandwidths at some value of i-f; or in finding required bandwidths for values of Q and k previously determined. Design examples will be given for the more difficult cases; this is indicated particularly in the section on undercoupled transformers, where it is easy to take a given Q and k and find the bandwidth, but it is rather more difficult if the problem of finding Q and k is attempted. For clarity the cases of critical, over-, and under-coupled transformers will be dealt with separately. Single tuned circuits are also included as they are sometimes required in i-f amplifiers. Additional data for the design of f-m transformers will be given in section 4 (v).

Stagger tuning (Refs. 8, 13, 17 & 38) of i-f transformers, to give substantially the same bandwidths as overcoupled transformers, does not have a very wide application in f-m and a-m receivers (except in cases where variable selectivity is to be used) and will not be discussed in detail. The loss in gain to be expected generally suggests overcoupling as a more satisfactory arrangement. Stagger tuning of single tuned circuits is widely used in television receivers (Refs. 27 & 28), but this is a rather different application to where it is applied to the comparatively narrow bandwidths of ordinary sound receivers.

Since i-f transformers for television receivers involve special problems they are not treated here (See Refs. 27 & 28). Also, triple tuned transformers are not discussed but the article of Ref. 26 gives an excellent treatment.

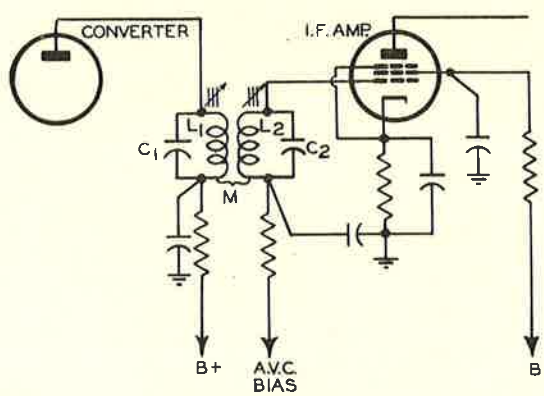

FIG. 26·1 TYPICAL TRANSFORMER COUPLED I.F. STAGE

Finally, the design methods do not make allowance for regenerative effects, nor should they be applied for finding the shape of resonance curves at frequencies far removed from resonance. The selectivity curve shapes are assumed to be perfectly symmetrical, although in practice it will be found that this is seldom true.

(ii) Critically Coupled Transformers

(*A*) Design Equations & Table

In general the procedure given is similar to that due to Kelly Johnson (Ref. 1) and Ross (Ref. 2). However, these procedures assumed that $Q = Q_1 = Q_2$; we shall take $Q = \sqrt{Q_1 Q_2}$ and the degree of approximation, when compared directly with the exact equations (Ref. 2), will be found to be somewhat less than is indicated by how closely $\sqrt{Q_1 Q_2}$ approaches $\frac{Q_1 + Q_2}{2}$. For most cases it is not advisable to use the equations given below for $\frac{Q_1}{Q_2}$ or $\frac{Q_2}{Q_1}$ greater than about 2, unless some additional adjustment is made from the universal selectivity and phase shift curves.

For N critically coupled transformers,

$$p = (1 + X^4/4)^{\frac{N}{2}} \qquad \text{.......... (1)}$$

$$X = \sqrt{2}\,(p^{\frac{2}{N}} - 1)^{1/4} \qquad \text{.......... (2)}$$

$$Q = \frac{X f_o}{2 \Delta f} \qquad \text{.......... (3)}$$

$$\text{and } \theta = \tan^{-1} \frac{2X}{2 - X^2} \qquad \text{.......... (4)}$$

where p = attenuation at Δf c/s off resonance.
N = number of identical transformers used ($N = 1$ for one transformer)
f_o = resonant frequency (the i-f in our case)
$Q = \sqrt{Q_1 Q_2} = \frac{1}{k_c}$ (in which Q_1 and Q_2 are the actual primary and secondary Q's; k_c is the critical coupling coefficient)
$2\Delta f$ = total bandwidth for a given attenuation (p)
θ = the phase shift between the secondary current at resonance and the secondary current at Δf c/s off resonance.

The required design information is given in equations (1) to (4). Usually either p is stated for a given bandwidth and a known i-f, or X can be found to allow p to be determined. Once these two factors have been found, the determination of k and/or Q is a simple matter.

Table 1 lists various values of p and X. Suppose p is known, then X is read from the table and used in equation (3) to find Q (since $2\Delta f/f_o$ is already known). The coefficient of coupling is then $k_c = 1/Q$.

If complete resonance and/or phase shift curves are required, then the universal curves of Figs. 9.17 and 9.18 (Chapter 9, Section 10) are used. These curves apply to one transformer only. For N identical transformers the attenuation in decibels is multiplied by N; for transformers which are not identical, the individual attenuations (in db) are added. Resonance and phase shift curves can also be determined directly, by using table 1 and equations (1) to (4).

In the application of the universal curves take $D = X = Q\ 2\Delta f/f_o$ (this Q being $\sqrt{Q_1 Q_2}$ as determined) and $b = k/k_c = Qk$ (for critical coupling $b = 1$, in this case); which are the same as for $Q_1 = Q_2$. If the values of Q_1 and Q_2 differ by more than about 2 to 1, then the more exact expressions for D and b^2 are applied to check how closely the required conditions are approached, it being carefully noted that in all expressions on the curves involving a that the Q shown is Q_2.

The maximum stage gain is given by

$$\text{Gain} = \frac{g_m\ Q\ \omega_o\ L}{2} \quad \ldots\ldots\ldots\ldots (5)$$

where g_m = mutual conductance of i-f valve (if conversion gain is required, conversion conductance (g_c) is substituted for g_m)

ω_o = $2\ \pi\ \times$ resonant frequency (f_o) i.e. f_o = i-f.

$L = \sqrt{L_1\ L_2}$; L_1 and L_2 are primary and secondary inductances

$Q = \sqrt{Q_1\ Q_2}$; Q_1 and Q_2 are primary and secondary magnification factors.

(A condition, not specifically stated in the equations is that $L_1\ C_1 = L_2\ C_2$ in all cases).

The maximum gain is usually converted to decibels, so that the gain at any point on the resonance curve can be found by subtraction of the attenuation, also expressed in decibels.

Table 1.

CRITICALLY COUPLED TRANSFORMERS

For Use with Equations (1), (2) & (3).
N = Number of Transformers.

Attenuation (p) Times Down	db Down	$N = 1$ X	$N = 2$ X	$N = 3$ X
√2	3	1.41	1.14	1.01
2	6	1.86	1.41	1.25
4	12	2.76	1.86	1.57
7	17	3.73	2.21	1.81
10	20	4.46	2.46	1.95
20	26	6.32	2.96	2.26
40	32	8.96	3.54	2.56
70	37	11.9	4.08	2.82
100	40	14.1	4.46	3.02
1000	60	—	7.96	4.46
10000	80	—	14.1	6.66

(*B*) Example and Additional Design Extension

A 455 Kc/s i-f transformer using critical coupling, is required to give a total bandwidth of 20 Kc/s for an attenuation of 20 db (10 times).

(a) $\dfrac{f_o}{2\Delta f} = \dfrac{455}{20} = 22.75$

(b) From table 1 we have $X = 4.46$ (since $N = 1$)

(c) From Eqn. (3).
$Q = 4.45 \times 22.75 = 101.$

(d) $k_c = 1/Q = 0.0099.$

(e) Select a suitable value for C_1 ($=C_2$); a capacitance of 100 $\mu\mu$F is satisfactory (made up of fixed + stray capacitances).
Then

$$L = \frac{25.33}{f^2C} = \frac{25.33}{0.455^2 \times 100} = 1.22 \text{ mH}$$

(f in Mc/s; C in $\mu\mu F$)

and take $L = L_1 = L_2$ since this is convenient in this case.

(f) The i-f valve (e.g. type 6SK7) has g_m = 2mA/volt (=2,000 μ mhos).
From Equation (5)
Max. stage gain
$= \pi \times 2 \times 101 \times 0.455 \times 1.22$
$=$ 352 times (or 51 db).

(g) Some designs might stop here and the magnification factor would be taken as $Q = Q_1 = Q_2 = 101$. It would be realized that valve loading would have an effect and possibly nothing more would be done (or else some attempt would be made to allow for the plate and grid resistances by finding new values of Q_1 and Q_2).

Let us proceed further and ask if the transformer as it stands fulfils the design conditions in a radio receiver. The answer is that obviously it does not, since it would be connected in most cases between two i-f valves, a converter and i-f valve or between an i-f valve and a diode detector.

Suppose the connection between two i-f amplifier valves (type 6SK7 would be representative) is considered, since this appears a fairly innocuous case. The plate resistance (r_p) of a type 6SK7 under the usual conditions of operation is 0.8 megohm. The short circuit input resistance, also under one set of operating conditions, is 6.8 megohm (this is calculated from the data given in Chapter 23, Section 5); other effects, which will alter this value, will be ignored for simplicity, although they may not be negligible.

It is first required to determine what values of Q_1 and Q_2 are required to give $Q = Q_1 = Q_2 = 101$. This is found from

$$Q_u = \frac{Q\,R}{R - Q\,\omega_o\,L} \quad \ldots\ldots\ldots \quad (6)$$

where Q_u = unloaded Q
Q = loaded Q
R = parallel resistance across winding
$\omega = 2\,\pi\,f_o$; (resonant frequency $= f_o$)
and L = inductance.

For the primary
$Q = 101$;
$R = r_p = 0.8$ MΩ;
$\omega_o = 2\,\pi \times 455 \times 10^3$;
$L = L_1 = 1.22$ mH
and $Q\omega_oL = 101 \times 2\,\pi \times 455 \times 10^3 \times 1.22 \times 10^{-3} = 0.352$ MΩ.

$$\text{Then } Q_u = \frac{101 \times 0.8}{0.8 - 0.352} = 180.$$

For the secondary

$$Q_u = \frac{101 \times 6.8}{6.8 - 0.352} = 106.8.$$

(*C*) Design Extensions

The value $Q_u = 180$ could not be obtained very easily, if at all, with a normal type of i-f transformer. In addition, the disadvantage of unequal primary and secondary Q's should be apparent. For values of Q only about 10% higher than that given, or where the transformer is coupled to a diode detector, the situation becomes worse and it is clear that a revised approach is necessary. What is actually needed is

(1) A transformer with equal values of primary and secondary Q's when unloaded. These will be denoted by $Q_u = Q_{u1} = Q_{u2}$.

(2) The values of Q_{u1} and Q_{u2} to be such that when the transformer is connected into the i-f amplifier, and loaded by the valve output and input resistances, the desired value of $Q = \sqrt{Q_1\,Q_2}$ will be obtained.

(3) The required coefficient of coupling (k) (critical for this particular example) to be unchanged. It will be described later how k can be pre-set to any desired value for any two circuits coupled together and tuned to the same resonant frequency.

(4) Excessive values of Q_{u1} and Q_{u2} are to be avoided (see the previous method of determining Q_{u1}) as far as possible, because of the practical difficulties involved.

(5) The response curve of frequency versus attenuation to be that specified (or very close to it).

All of these conditions can be fulfilled very closely, provided the approximations made in deriving the design equations hold. A simple analysis of the circuits involved, and including the required conditions, gives

$$Q_u = \frac{\alpha + \sqrt{\alpha^2 + 2Q^2R_1R_2\beta}}{\beta} \quad \ldots\ldots \quad (7)$$

where $Q_u = Q_{u1} = Q_{u2}$ (unloaded primary and secondary Q)
$Q = \sqrt{Q_1\,Q_2}$ (in which Q_1 and Q_2 are loaded primary and secondary Q's)
R_1 = parallel resistance (r_p in our case) shunted across trans. primary.
R_2 = parallel resistance (grid input in our case) shunted across trans. secondary.
$\alpha = Q\,(Q\omega_oL)\,(R_1 + R_2)$; in all cases it will be taken that $L = L_1 = L_2$
and $\beta = 2[R_1R_2 - (Q\omega_oL)^2]$.

For our example:
$Q = 101$; $Q^2 = 1.02 \times 10^4$
$R_1 = 0.8$ MΩ $R_2 = 6.8$ MΩ
$Q\omega_oL = 0.352$ MΩ (found previously)
$\alpha = 101 \times 0.352 \times 7.6 = 270$;
$\alpha^2 = 7.29 \times 10^4$
$\beta = 2\,[5.44 - 0.124] = 10.63\ (\text{M}\Omega)^2$

$$Q_u = \frac{270 + \sqrt{7.29 \times 10^4 + 2 \times 1.02 \times 10^4 \times 5.44 \times 10.63}}{10.63} = 131.$$

so that $Q_{u1} = Q_{u2} = 131$.

To check that the transformer, when placed in the receiver, gives the desired value of $Q = \sqrt{Q_1\,Q_2}$ use

$$Q = \frac{Q_uR}{Q_u\,\omega_oL + R} \quad \ldots\ldots\ldots\ldots\ldots \quad (8)$$

from which

$$Q_1 = \frac{131 \times 0.8}{0.458 + 0.8} = 83.4$$

$$Q_2 = \frac{131 \times 6.8}{0.458 + 6.8} = 123$$

and so

$Q = \sqrt{Q_1\,Q_2} = \sqrt{83.4 \times 123} = 101$, which is the desired value (as determined previously).

(*D*) Conclusions

All that is required to design the specified transformer is to go through the simple steps (a) to (f) and, knowing R_1 and R_2, apply equation (7). Overall response and phase shift are determined from the universal curves, as explained previously.

It should be obvious that equation (7) will not hold under all practical conditions, but it is not limited by the ratio of Q_1/Q_2 *or* Q_2/Q_1, and failing cases can be checked by the condition for $\beta \leq 0$.

It has been assumed for simplicity that $L = L_1 = L_2$, but this is not essential, and the design equation could be extended to the case of $L = \sqrt{L_1 L_2}$. In a failing case, if it is essential to fulfil the specified conditions, Q_1 and Q_2 (and if necessary L_1 and L_2) are selected to give the desired values of Q and L; this will be illustrated in the section on the design of variable bandwidth crystal filters; from this the advantage of taking geometric mean Q and L values will be even more apparent than for the preceding example.

(*E*) *k* **Measurement**

The coefficient of coupling, k, for two circuits resonant at the same frequency, can be set on a Q meter (provided Q_b lies within the useful working range) using the relationship

$$Q_b = \frac{Q_{u1}}{1 + Q_{u1} Q_{u2} k^2} \qquad \ldots\ldots (9)$$

if $Q_{u1} = Q_{u2} = Q$, *then*

$$Q_b = \frac{Q_u}{1 + (Q_u k)^2} \qquad \ldots\ldots (9A)$$

where

Q_{u1} = primary Q (sec. o/c or detuned by large amount)

Q_{u2} = secondary Q (pri. o/c or detuned by large amount)

Q_b = Q to be obtained when primary and secondary are coupled and both tuned to the same resonant frequency

(usual precautions as to can being earthed etc. to be observed).

In our example we desire a value for $k = 0.0099$ (for critical coupling when $Q = \sqrt{Q_1 Q_2} = 101$); $Q_u = 131$. Then from (9A)

$$Q_b = \frac{131}{1 + (1.297)^2} = 48.9.$$

All that is required is to adjust the spacing between the two resonant circuits until the Q meter reads 48.9 — the desired co-efficient of coupling has then been obtained. Alternatively, by transposing terms in the equation, k is given for any values of Q_b, Q_{u1}, and Q_{u2}. The method applies directly to under-, over-, or critically-coupled transformers and is useful within the limits set by the usable range of the Q meter. For overcoupled transformers additional methods are sometimes required, and the procedure will be indicated in the next section.

(iii) Overcoupled Transformers

(*A*) Design Equations & Table

Here the method to be followed is based on that due to Everitt (Ref. 3).

Fig. 26.5 illustrates the terms used regarding bandwidth.

It should be noted that when the primary and secondary Q's differ appreciably, two peaks of secondary output voltage do not appear immediately critical coupling is exceeded. The actual value of k,

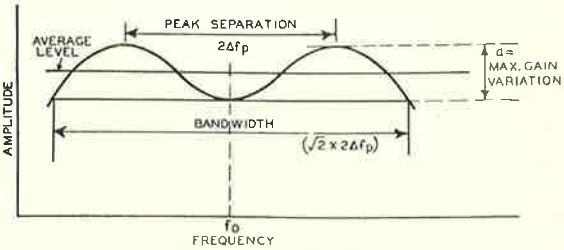

FIG. 26·5 ILLUSTRATION OF TERMS USED FOR OVER-COUPLED TRANSFORMERS

which corresponds to the condition for two peaks of secondary voltage, is called the transitional coupling factor (see Chapter 9, Section 6(V)), and Ref. 8.

In what follows we shall use $Q = \sqrt{Q_1 Q_2}$ and $L = \sqrt{L_1 L_2}$, as was done for critically coupled transformers, but this is not an approximation in the derivation of the design equations (10), (11) & (12) provided that $L_1 C_1 = L_2 C_2$. It will also be assumed, for simplicity, that the peaks of the response curve are of equal height and symmetrically placed in regard to f_o.

$$Qk = A + \sqrt{A^2 - 1} \qquad \ldots\ldots (10)$$

$$A = \frac{(Qk)^2 + 1}{2Qk} \qquad \ldots\ldots (11)$$

$$\frac{2\Delta f_p}{f_o} = k\sqrt{1 - 1/(Qk)^2} \qquad \ldots\ldots (12)$$

$$\theta = \tan^{-1}\frac{2X}{1 - X^2 + (Qk)^2} \qquad \ldots\ldots (13)$$

where

$Q = \sqrt{Q_1 Q_2} = 1/k_c$ (in which Q_1 and Q_2 are primary and secondary Q's respectively; k_c is coefficient of critical coupling)

k = any coefficient of coupling equal to or greater than critical.

A = gain variation from peak to trough (i.e. difference in transmission level)

$2\Delta f_p$ = bandwidth between peaks; $\sqrt{2}$ $(2\Delta f_p)$ is the total bandwidth for two other points on the resonance curve with the same amplitude as at f_o.

θ = phase shift between the secondary current at resonance and the secondary current at Δf c/s off resonance

$$X = \frac{2\Delta f}{f_o} Q$$

and f_o = resonant frequency of transformer (i-f).

The universal resonance and phase shift curves of Figs. 9.17 and 9.18 (Chapter 9, Section 10) are directly applicable; using the exact expressions if desired and taking Q as Q_2 for all terms involving a. It is more convenient, and sufficiently accurate, to use the conditions for $Q_1 = Q_2$ when Q_1/Q_2 or $Q_2/Q_1 \not> 2$; in this case $b = Qk$ (or k/k_c), $D = (2\Delta f/f_o)\ Q$ and since these expressions do not involve a, the value $Q = \sqrt{Q_1\ Q_2}$ as determined in the design problem is used. A check will reveal that it is difficult to read any difference from the curves whichever method is used.

It should be observed that five points on the resonance curve are given directly from the design equations.

To find the maximum stage gain which occurs at the peaks, the equation (5) as given for critically coupled transformers, is applied directly. Generally it is the average gain in the pass band ($\sqrt{2} \times 2\Delta f_p$) which is required and this is given by multiplying equation (5) by

$$\frac{(Qk + 1)^2}{2\ [(Qk)^2 + 1]} \quad \ldots\ldots\ldots (14)$$

If the gain at f_o (i.e. at the trough of the curve) is required, equation (5) is multiplied by

$$\frac{2Qk}{(Qk)^2 + 1}\ \left(= \frac{1}{A}\right) \quad \ldots\ldots\ldots (15)$$

Equations (14) and (15) can be evaluated directly from fig. 26.4 when Qk is known; the dotted line being for equation (14) and the solid line for equation (15). The gain reduction factor so found, is used as a multiplier with equation (5) (the gain reduction indicated by equation (15) can also be read directly from table 2).

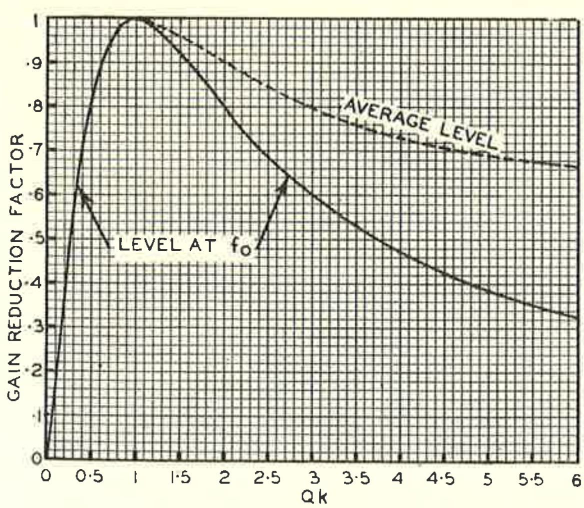

FIG. 26·4 GAIN REDUCTION FACTORS FOR COUPLING OTHER THAN CRITICAL

Table 2.

OVER COUPLED TRANSFORMERS

For Use with Equations (10), (11), (12), (14), (15).

A (= peak to trough gain variation)

db	Times Down	Qk	$\sqrt{1-1/(Qk)^2}$	$(Qk)^2$	$\frac{1}{A}$
0	1.00	1.00	0.0	1.00	1.00
0.25	1.03	1.27	0.616	1.61	0.971
0.50	1.06	1.41	0.707	2.00	0.943
1	1.12	1.73	0.817	3.00	0.893
1.9	1.25	2.00	0.866	4.00	0.800
2	1.26	2.02	0.869	4.08	0.794
3	1.41	2.41	0.910	5.81	0.709
3.1	1.43	2.45	0.913	6.00	0.670
4	1.59	2.81	0.935	7.90	0.629
4.4	1.67	3.00	0.943	9.00	0.599
5	1.78	3.25	0.952	10.56	0.562
6	2.00	3.72	0.963	13.84	0.500
6.6	2.13	4.00	0.968	16.00	0.469
7	2.24	4.24	0.972	17.98	0.446

(*B*) **Example**

A 455 Kc/s i-f transformer is required to pass a band of frequencies 16 Kc/s wide (i.e. ± 8 Kc/s). The variation in gain across the pass band is not to exceed 0.5 db.

(a) From fig. 26.5 it is reasonable to take the *total* bandwidth as 16 Kc/s and so the peak separation is

$$\frac{16}{\sqrt{2}} = 11.3 \text{ Kc/s.}$$

(b) $$\frac{2\Delta f}{f_o} = \frac{11.3}{455} = 0.0248.$$

(c) From Table 2, $\sqrt{1 - 1/(Qk)^2} = 0.707$.

(d) From Eq. (12), $k = \dfrac{0.0248}{0.707} = 0.035$

(e) From Table 2, $Qk = 1.41$

$$Q = \frac{1.41}{0.035} = 40.2$$

(f) Assuming a value for C_1 (= C_2) as 80 $\mu\mu$F (including strays)

$$L = \frac{25.33}{0.455^2 \times 80} = 1.53 \text{ mH}$$

and take $L = L_1 = L_2 = 1.53$ mH.

(g) To determine the average stage gain in the pass band. From fig. 26.4 (or eq. 14), with $Qk = 1.41$, read from the dotted curve, gain reduction factor equals 0.97. Assuming we use a type 6J8-G converter valve having a conversion conductance of 290 μmhos (0.29 mA/volt), then from eq. (5) and the gain reduction factor, Average stage gain

$= 0.97 \times \pi \times 0.29 \times 40.2 \times 0.455 \times 1.53$

$= 24.7$ times (or 27.9 db).

(h) Assume that the transformer is connected between a type 6J8-G converter and a type 6SK7 voltage amplifier and that both valves are working under one particular set of operating conditions.
For the type 6J8-G the plate resistance $r_p = 4\text{M}\Omega = R_1$ and for the type 6SK7 the short circuit input resistance $= 6.8\ \text{M}\Omega = R_2$ (as determined from Chapter 23, Section 5).

From Equation (7)

$\alpha = 40.2 \times 0.17 \times 10.8 = 73.8;$

$\alpha^2 = 0.544 \times 10^4$

$\beta = 2\ [27.2 - 0.17^2] = 54.4$ approximately

$$Q_u = \frac{73.8 + \sqrt{0.544 \times 10^4 + 4.76 \times 10^6}}{54.4} = 41.4$$

So that $Q_u = Q_{u1} = Q_{u2} = 41.4$ which is the unloaded value for primary and secondary Q's before the transformer is connected between the two valves; the additional refinement in design is hardly necessary here, and it would be sufficient to make $Q = Q_1 = Q_2 = 40$ (approx.).

It can easily be checked, using the procedure set out for critically-coupled transformers, that the geometric-mean, of the loaded values Q_1 and Q_2, is 40.2 as required. Complete resonance and phase shift curves are plotted as previously explained. The value for k can be set in exactly the same way as explained in the section on critically-coupled transformers, if this is convenient. If k is very high it is preferable to apply the method detailed in Chapter 37. This uses the relationship

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

where: C_1 = capacitance required to tune the primary to resonance with the secondary open circuited

k = coefficient of coupling required (say k greater than 0.1 or so)

ΔC = increment in capacitance required to tune the primary to resonance when the secondary is short circuited.

As an illustration suppose $k = 0.2$ and $C_1 = 200\ \mu\mu\text{F}$ (the exact working frequency may not always be convenient) then $\Delta C = 8.34\ \mu\mu\text{F}$. Using a "$Q$" meter, the spacing between primary and secondary is adjusted until this increment in capacitance is obtained; this gives the required value of k.

With some odd transformers, neither of the two methods may be very convenient, and, in such cases, the gain variation in the pass band can be measured with a single stage amplifier using the relationship of eqn. (11).

(iv) Undercoupled Transformers & Single Tuned Circuits

(A) Single Tuned Circuits

It is sometimes necessary to use combinations of single tuned circuits or under-coupled transformers in conjunction with over-coupled circuits to give a substantially level response over the pass band. Another application for the under-coupled transformer often arises when an improvement in selectivity is needed, without an excessive loss in stage gain. It has been shown by Adams (Ref. 9) that the optimum conditions for selectivity and gain for a given Q, are obtained when the coefficient of coupling is approximately 0.82 of critical. However, it is readily shown for a transformer having k equal to 0.5 k_c that the loss in gain is only about 2 db (approx. 0.8 of maximum gain).

The design equations (Ref. 1) that follow are applicable to single tuned circuits, which will be considered first:

Single Tuned Circuits

$$p = (1 + X^2)^{\frac{N}{2}} \qquad \text{............ (16)}$$

$$X = (p^{\frac{2}{N}} - 1)^{\frac{1}{2}} \qquad \text{............ (17)}$$

$$Q = \frac{X f_o}{2\ \Delta f} \qquad \text{............ (18)}$$

$$\text{and } \theta = \tan^{-1} \frac{2\ \Delta f}{f_o} Q = \tan^{-1} X \quad \text{...... (19)}$$

where p = attenuation at Δf c/s off resonance

N = number of identical tuned circuits (for a single tuned circuit $N = 1$)

f_o = resonant frequency

and θ = phase shift between voltage at resonance and the voltage at Δf c/s off resonance.

The design of the single tuned circuit presents no difficulty and it is only necessary to use Table 3 in conjunction with eqns. (16), (17) and (18). Selectivity and phase shift can be found from the universal curves of figs. 9.17 and 9.18 (Chapter 9, Section 10), or selectivity can be evaluated directly from eqns. (16), (17) and (18) used in conjunction with Table 3.

Table 3.
SINGLE TUNED CIRCUITS

For use with Equations (16), (17) & (18).
N = Number of Identical Tuned Circuits.

Attenuation (p) Times Down	db Down	$N = 1$ X	$N = 2$ X	$N = 3$ X
$\sqrt{2}$	3	1.0	0.644	0.509
2	6	1.73	1.00	0.767
4	12	3.87	1.73	1.23
7	17	6.93	2.45	1.63
10	20	9.95	3.00	1.91
20	26		4.36	2.52
40	32		6.25	3.27
70	37		8.31	4.00
100	40		9.95	4.53
1000	60			9.95

(*B*) Example

Two single tuned circuits are required to give an attenuation of 17 db for a total bandwidth of 10 Kc/s. The i-f is 455 Kc/s.

(a) From Table 3, $X = 2.45$ (since $N = 2$)

(b) $Q = \dfrac{2.45 \times 455}{10} = 111.5$ from eqn. (18).

(c) Assuming $C = 200\ \mu\mu F$ (including all strays)

$$L = \frac{25.33}{0.455^2 \times 200} = 0.611 \text{ mH.}$$

(d) The unloaded Q depends on the combined effect of valve input and output resistance. Take the loading for the two transformers as being the same, for simplicity.
Then applying the assumed values $r_p = 0.8$ MΩ and grid input resistance = 6.8 MΩ, the effective shunt resistance is 0.715 MΩ.

From Eqn. (6)

$$Q_u = \frac{111.5 \times 0.715}{0.715 - 0.194} = 153$$

(e) The gain of each stage (twice that for a critically coupled transformer) is $g_m\ Q\ \omega_o\ L$, so that taking g_m = 2mA/volt (2000 μmhos) in each case, stage gain = $2\pi \times 2 \times 111.5 \times 0.455 \times 0.611$ = 390 times (or 51.8 db).

(f) Suppose we have the loaded Q given as 111.5 (as in our previous problem using $N = 2$), and we require the bandwidth for 6db attenuation; from table 3 obtain $X = 1.0$ and from eqn. (18) the total bandwidth ($2\Delta f$) is 4.08 Kc/s. In a similar manner the attenuation can be found when the bandwidth is stated. The resonance curves could also be used to find X ($= D$ in Chapter 9, in this case) and p.

(*C*) Undercoupled Transformers

If the transformers use very loose coupling, the methods for single tuned circuits could be applied (N = 2 for each transformer) in conjunction with eqn. (7). This approach does not lead to very accurate results since the values of coupling are seldom less than 0.1 of critical and more often are in the order of 0.5 to 0.8 of critical.

General design equations applicable to transformers having any degree of coupling are given below, but it will be seen that they are not quite as tractable as in previous cases unless an additional factor such as Q, k, or Qk (i.e. a given proportion of critical coupling) is specified. However, this will offer little difficulty.

$$X = (p^{\frac{2}{N}} - 1)^{\frac{1}{4}}\ [(Qk)^2 + 1]^{\frac{1}{2}} \quad . \quad (20)$$

$$p = \left[1 + \frac{X^4}{(Qk)^2 + 1}\right]^{\frac{N}{2}} \quad (21)$$

$$X = \frac{2\ \Delta f}{f_o}\ Q \quad \ldots\ldots \quad (22)$$

$$Qk = \left[\frac{X^2}{(p^{\frac{2}{N}} - 1)^{\frac{1}{2}}} - 1\right]^{\frac{1}{2}} \quad \ldots\ldots \quad (23)$$

$$\theta = \tan^{-1} \frac{2X}{1 - X^2 + (Qk)^2} \quad \ldots\ldots \quad (13)$$

where p = attenuation at Δf c/s off resonance

$Q = \sqrt{Q_1\ Q_2} = 1/k_c$ (in which Q_1 and Q_2 are the primary and secondary Q's and k_c is critical coupling coefficient).

k = any coefficient of coupling

N = number of identical transformers

f_o = resonant frequency

and θ = phase shift between secondary current at and off resonance. The restriction of Q_1/Q_2 or $Q_2/Q_1 \ngtr 2$ is applied, as previously explained.

It may be observed that these equations are the most general ones, e.g., if $Qk = 1$ the equations reduce to those for critical coupling.

Stage gain is given by evaluating eqn. (5) and multiplying by eqn. (15) (or reading the gain reduction factor from fig. 26.4).

The coefficient of coupling can be set as described for critically coupled transformers.

(*D*) Example

A 455 Kc/s i-f transformer is required to give a total bandwidth of 20 Kc/s for an attenuation of 4 times. The transformer is to be connected between a voltage amplifier, having a plate resistance of 0.8 MΩ (e.g. type 6SK7), and a diode detector having a load resistance of 0.5 MΩ.

In this case other loading effects due to the a.v.c. system etc. will be neglected. For a typical case where the a.v.c. diode plate is connected by a fixed capacitor to the i-f transformer primary, there will be appreciable damping of the primary circuit due to the diode circuit (approx. $R_L/3$ when diode is conducting). This damping will not be constant for all signal input voltages, particularly if delayed a.v.c. is used.

(a) The difficulty first arises in evaluating X. If we select a suitable value for Q the problem becomes quite straightforward.
To select a value for Q it is necessary to realize that an unloaded value of 150 would be about the absolute maximum with normal types of construction, and even this figure is well on the high side unless an "iron pot" or a fairly large can and former are used. Assume $Q = 150$ for this problem (so far as the procedure is concerned it is unimportant if a lower value is selected).
The next point is that circuit loading will set a limit to the value of

$$Q\ \omega_o\ L\left(=\frac{Q}{\omega_o\ C}\right)$$

Now L will be set, normally, by the minimum permissible value of C.
Suppose $C = 85$ $\mu\mu$F including strays, then $L = 1.44$ mH; and we will take $L = L_1 = L_2$ and $C = C_1 = C_2$ for practical convenience. Better performance would be possible by making $L_1 > L_2$ but the improvement is only small, and hardly worthwhile unless the secondary load is very small.

(b) For our problem
$\omega_o L = 1/\omega_o C = 4{,}120\Omega$; also $R_1 = 0.8\ M\Omega$
and $R_2 = 0.5/2 = 0.25\ M\Omega$ (half the diode load resistance).
Then applying eqn. (8)

$$Q_2 = \frac{150 \times 0.25}{0.25 + (150 \times 4.12 \times 10^{-3})} = 43.3$$

$$\text{and } Q_1 = \frac{150 \times 0.8}{0.8 + 0.618} = 84.6$$

so that

$$Q = \sqrt{Q_1\ Q_2} = \sqrt{84.6 \times 43.3} = 60.5.$$

(c) From eqn. (22)

$$X = \frac{20 \times 60.5}{455} = 2.66$$

(d) From eqn. (23)

$$Qk = \left[\frac{2.66^2}{(4^2-1)^{\frac{1}{2}}} - 1\right]^{\frac{1}{2}} = 0.91$$

$$\text{and } k = \frac{0.91}{60.5} = 0.015$$

If the selectivity requirements are too severe a negative value will be found for Qk. A value for Qk is always possible when

$$X > (p^{\frac{2}{N}} - 1)^{\frac{1}{4}}.$$

(e) From eqns. (5) and (15) (the solid curve of fig. 26.4)

$$\text{Stage gain} = \frac{(2 \times 60.5 \times 4.12)}{2}\ 0.994$$

= 248 times; or 47.9 db.

(f) The completed transformer has primary and secondary Q's of 150 (before connection into the receiver), a coefficient of coupling equal to 0.015 (which is 0.91 of critical coupling when the transformer windings are loaded), primary and secondary inductances of 1.44 mH and tuning capacitances of 85 $\mu\mu$F (including strays).

The stray capacitances across the primary would be valve output (7 $\mu\mu$F for type 6SK7) plus distributed capacitance of winding, plus capacitances due to wiring and presence of shield can; across the secondary there would be diode input capacitance (about 4 $\mu\mu$F for a typical case), plus distributed capacitance of secondary winding plus wiring and shield can capacitances. The total capacitances can be measured in the receiver or estimated using previous experience as a guide; typical values would be 10 – 20 $\mu\mu$F depending on the type of i-f transformer, valves etc. If the second valve is not a diode, the input capacitance should also include that due to space charge, Miller effect etc. as discussed in Chapter 2, Section 8; Chapter 23, Section 5; and Section 7 of this Chapter. Changes in input resistance which would affect the loading across the i-f transformer secondary, are also discussed in these same sections. Input capacitance changes with a.v.c. are considered in Section 7 of this Chapter.

(g) A complete resonance curve can be obtained from Fig. 9.17 (Chapter 9, Section 10), by

$$\text{taking } D = Q\ \frac{2\ \Delta f}{f_o}\ (= X) \text{ and}$$

$b = Qk = k/k_c$, or directly from the design equations (20), (21) and (22).

(v) F-M i-f Transformers

The design methods given so far are applicable to both f-m and a-m transformers, but there is additional data available which will be of assistance.

Bandwidth requirements are of importance, and it is fairly generally accepted that the i-f amplifier should be capable of passing all the significant sideband frequencies of the frequency modulated wave; where significant sidebands are taken as those having amplitudes which are greater than about 1% of the unmodulated carrier amplitude. The bandwidths for this condition can be found from Table 4 (see also Ref. 19) for commonly occurring values of modulation index.

Table 4.

BANDWIDTHS FOR USE WITH F-M TRANSFORMERS

$$\text{Modulation Index} = \frac{\Delta F}{f} = \frac{\text{carrier frequency deviation}}{\text{audio modulating frequency}}$$

Values for $\frac{\Delta F}{f}$ may be interpolated with sufficient accuracy.

$\frac{\Delta F}{f}$ =	0.01–0.4	0.5	1.0	2.0	3.0	4.0	5.0	6.0
Bandwidth =	$2f$	$4f$	$6f$	$8f$	$12f$	$14f$	$16f$	$18f$
$\frac{\Delta F}{f}$ =	7.0	8.0	9.0	10.0	12.0	15.0	18.0	21.0
Bandwidth =	22f	$24f$	$26f$	$28f$	$32f$	$38f$	$46f$	$52f$

As an example; for the f-m broadcast band $\Delta F = \pm$ 75 Kc/s and the highest audio frequency is 15 Kc/s, then

$$\frac{\Delta F}{f} = \frac{75}{15} = 5.$$

Then the required bandwidth is 16 $\times$ 15 = 240 Kc/s.

The highest audio frequency is chosen because this imposes the most severe requirements on bandwidth; e.g. suppose we had taken f = 7.5 Kc/s, then the modulation index would be 10, and the bandwidth = $28f$ = 28 $\times$ 7.5 = 210 Kc/s.

The bandwidths actually employed in a receiver should also make allowance for possible drift in the oscillator frequency. A reasonably good oscillator should not drift by more than about $\pm$ 20 Kc/s when operating around 110 Mc/s; so that an additional 40 Kc/s should be added to the bandwidth. Of course, this is only a rough approximation, since the oscillator frequency variation is random and would introduce additional f-m; the determination of the true bandwidth would be quite a difficult problem, unless several simplifying assumptions are made.

From what has been said, it appears that the receiver total bandwidth should be about 280 Kc/s to fulfil the most severe requirements. However, most practical receivers limit the total bandwidth to about 200 Kc/s, which is not unreasonable since it is seldom that the frequency deviation would be 75 Kc/s when the audio frequency is 15 Kc/s; on the average the deviation is $\pm$ 50 Kc/s or less.

For such large bands of frequencies to be passed through tuned circuits, it is necessary to have some criterion which will allow the permissible amount of attenuation to be estimated for the required bandwidth. To eliminate non-linear distortion the circuits should provide a uniform amplitude and a linear phase characteristic over the operating range. Curvature of the phase characteristic of the tuned circuits will cause non-linear audio distortion, while curvature of the amplitude characteristic may cause additional distortion if the amplitude happens to drop below the operating voltage range of the amplitude limiting device incorporated in the receiver. A suitable criterion can be determined from the phase angle/frequency characteristics of the tuned circuits (the phase angle being that between the secondary current at resonance to that at Δf c/s off resonance). Inspection of universal phase shift curves will show that the greatest range of linearity of phase shift versus frequency change, is given by critically coupled transformers (Ross, Ref. 2); but slight overcoupling does not lead to excessive non-linearity. Overcoupling has some advantages, in particular slightly greater adjacent channel selectivity can be obtained; but there is the disadvantage of more difficult circuit alignment. As an extention of this work, Ross (Ref. 2) has also shown that for a critically coupled transformer a suitable criterion of permissible non-linearity is that

$$X \not> 2 \qquad \ldots\ldots\ldots\ldots\ldots\ldots \quad (24)$$

where we will take

$$X = \frac{2\Delta f}{f_o} Q$$ (this is the same X as previously)

$2\Delta f$ = total bandwidth

f_o = central carrier frequency

and $Q = \sqrt{Q_1 Q_2}$ (where Q_1 and Q_2 are the primary and secondary magnification factors).

The amount of introduced amplitude modulation can be estimated from

$$m = \frac{p - 1}{p + 1} \qquad \ldots\ldots\ldots\ldots \quad (25)$$

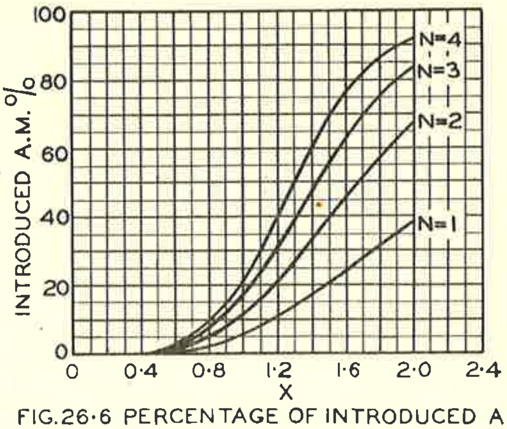

FIG.26·6 PERCENTAGE OF INTRODUCED A-M IN AN F-M CURRENT FOR N CRITICALLY COUPLED TRANSFORMERS.

where m = amplitude modulation factor
and p = attenuation at the specified bandwidth.

Fig. 26.6 (Ref. 2) shows directly values of m for various values of X. Values for m are of importance since they allow an estimate to be made of the amplitude limiting requirements demanded from whatever device is incorporated in the receiver to "iron out" amplitude variations. Usual f-m receiver design allows 30 to 50% of introduced amplitude modulation at the greatest bandwidth required, but this is quite arbitrary and no set standards are available.

It is also worth noting, before leaving this section, that the carrier frequency should be regarded as a reference point only, since unlike amplitude modulation, its amplitude varies and becomes zero under some conditions of modulation. This is the basis of a method due to Crosby (Ref. 25) used for measuring frequency deviation.

(*B*) Example

An f-m i-f amplifier is required using three critically coupled transformers. The i-f is 10.7 Mc/s and the frequency deviation ± 75 Kc/s; the highest audio modulating frequency is 15 Kc/s (critical coupling has been selected in this case but a combination of critical and overcoupled transformers might lead to a better solution).

The converter valve to be used has an r_p of 1.5 MΩ and a conversion conductance of 475 μ mhos; and the two i-f valves each have plate resistances of 2MΩ and g_m = 5,000 μ mhos.

(a) For simplicity it will be taken that any additional selectivity, due to the other tuned circuits in the receiver, is negligible in the i-f pass band. Also, as will be illustrated, the dynamic impedances of the i-f transformers will be so low as to render additional damping due to plate and grid input resistances negligible; this is not true, however, if the final transformer is connected to a limiter stage because of grid current damping — some consideration will be given to this later.

(b) From previous considerations regarding bandwidth, in connection with Table 4, we will adopt 220 Kc/s as a compromise. If we design on the limit of X = 2, then from eqn. (3)

$$Q = \frac{2 \times 10.7 \times 10^6}{220 \times 10^3} = 97.4$$

and $k_c = 1/Q = 0.0103$.

(c) From eqn. (1) (or with sufficient accuracy by interpolating in Table 1) since X = 2 and $N = 3$, $p = 11.4$.
From eqn. (24) (or fig. 26.6)

$$m = \frac{11.4 - 1}{11.4 + 1} = 0.84 \text{ or } 84\%$$

amplitude modulation.

(d) This is a severe requirement for limiters etc., although the condition for linearity of phase shift with frequency is fulfilled, and it would be more suitable to take $m = 0.5$ as a compromise. Then

$$p = \frac{1 + m}{1 - m} = \frac{1.05}{0.5} = 3$$

and from eqn. (2) (or interpolation in Table 1) X = 1.44 and from eqn. (3) $Q = 70$; $k_c = 0.043$.

(e) Select suitable values for C_1 and C_2. To obtain the highest possible dynamic impedance these are usually made rather small. Take $C_1 = C_2 = 60$ μμF (including strays)

$$L \ (= L_1 = L_2) = \frac{25330}{10.7^2 \times 60} = 3.68 \ \mu\text{H}.$$

There would be an advantage in making $C_1 < C_2$ and $L_1 > L_2$ but this is awkward for winding i-f's on a machine.

(f) The dynamic impedance of each winding (considered uncoupled from one another) is
$R_o = Q\omega_o = 70 \times 2\pi \times 10.7 \times 3.68 = 17{,}300\Omega$
which is very much less than the valve plate or input resistances in typical stages.

(g) To find the overall gain of the i-f stages. For the first stage, connected to the converter, we are concerned with conversion gain. From eqn. (5) and step (f), Conversion

$$\text{gain} = \frac{475 \times 10^{-6} \times 17{,}300}{2} = 4.12$$

times (12.3 db).
For the second and third stages in each case,

$$\text{Gain} = \frac{4.12 \times 5}{0.475} = 43.4 \text{ times (32.8 db)}.$$

Hence the overall gain is 12.3 + (2 × 32.8) = 77.9 db.

(h) If the third transformer connects to a limiter its design should be modified for best results. Grid current will alter the effective input capacitance of the limiter valve and cause very appreciable detuning of the transformer secondary. To overcome this detuning it may be necessary to make the pass band of this transformer somewhat greater than for the other two, or else to use a very large condenser (in the order of 600 μμF) to tune the secondary. It has also been suggested that Q_1 should equal Q_2 in this case (Ref. 21). The important point, apart from possible distortion, is that the susceptibility to certain types of impulse noise is increased if the tuned circuits are not accurately aligned to the centre frequency of the discriminator. Grid current damping, of course, would tend to offset the effect to some extent (See also Refs. 21, 24). Since this transformer would probably be nonstandard in any case, it would be advantageous to make L_1 as large as possible, to assist in keeping the stage gain high. This follows as a result of the stage gain being directly proportional to $\sqrt{L_1 L_2 Q_1 Q_2} = LQ$.

(vi) I-F Transformer Construction

The methods of (i) to (v) in this section will allow the necessary design data for an i-f amplifier to be collected together. The final step is to determine the winding details and physical arrangement of the transformers. It is not proposed to discuss the merits of various types of windings but merely to give a few details which have proved helpful in practice.

For transformers working at the higher frequencies, the windings are quite often solenoids and the determination of the number of turns required is a simple matter. Satisfactory results can be obtained by applying Hayman's modification of Wheeler's formula as given in Chapter 10. Methods are also set out for determining the number of turns per inch, and suitable wire diameters for obtaining optimum values of Q. Usually wire gauges between 18 and 28 SWG are suitable as they are not so heavy as to be awkward to bend and they do not tear the usual type of coil former when the construction requires the leads to be passed through the inside of the former down to the base connections. Some error in the calculated number of turns will be apparent unless allowance is made for the inductance of leads. The number of turns required is finally determined experimentally in any case so that the calculated number of turns provides a good starting point.

Measurements must always be made with the coils in the cans because the effects of the can, brass mounting bosses, slugs etc. on inductance and Q are quite large.

Powdered iron "slugs" are commonly used for setting the inductance values and for 10.7 Mc/s, in particular, the iron must be very finely divided if the coil Q is not to be seriously changed as the cores are moved through the windings. Sufficient inductance variation also requires that the "slugs" have a certain minimum size.

Silvered mica fixed condensers are to be preferred for good frequency stability with temperature changes, particularly at high frequencies such as 10.7 Mc/s. Cheaper mica types are very often used at the lower frequencies around 455 Kc/s. If capacitance trimmers are to be used, care must be taken in their choice as pressure types are often mechanically unstable and sometimes have very low Q values. Suitable wax or other treatment should be applied to fixed condensers to offset the effects of changes in humidity.

Coils should be baked to remove moisture and then given suitable wax or varnish treatment to prevent humidity changes offsetting their properties. The electrical characteristics of the coils will be altered by wax etc. and it is essential to check the final values for Q, k etc. after the treatment is complete.

Typical former diameters for i-f transformers range from about $\frac{3}{8}''$ to $\frac{3}{4}''$.

The design of transformers with pie windings (a larger number of pies reduces distributed capacitance and affects Q) is not as simple as for solenoids since most equations require a knowledge of coil dimensions which are not readily available. There are also optimum sizes of winding to give the highest possible Q (e.g. a coil of approximately square cross section) — see Chapter 10. A method which has proved satisfactory in practice is to make measurements on various types of coils which are likely to be used fairly often and apply the relationship

$$N = B\sqrt{L} \quad \text{.......} \quad \text{.......} \quad (26)$$

where N = turns per pie

L = inductance in μH

and B = experimentally determined constant.

As an example: A winding is to be made to have an inductance of 1.44 mH. Previous experiments have shown that, in the frequency range of 300-900 Kc/s and for an inductance of about 0.5-2 mH, a two pie winding on a $\frac{9}{16}''$ former (each pie $^5/_{32}''$ wide, with $^3/_{32}''$ spacing between the pies and using 5/44 B&S Litz wire) has the factor B = 4.33. Then from eqn. (26)

$$N = 4.33\sqrt{1440} = 165 \text{ turns per pie;}$$

the two pies each of 165 turns, being connected series aiding. The same method can be applied, in cases where it is convenient, to any type of winding.

Since the presence of the iron "slug" will affect inductance (and r-f resistance), it is necessary to determine its effect and also to calculate the variation in inductance which can be made. The turns required are found for the condition with the "slug" in the winding and in the position giving the mean inductance value. This means that the value for L used in eqn. (25) will be less than the calculated value by the increase due to the iron.

The value of the inductance (L) using a powdered iron core (e.g. magnetite) can be found from

$$L = L_0 \left[1 + a\left(\frac{r_1}{r_2}\right)^2 \frac{l_1}{l_2}(\mu eff - 1)\right] \text{ . } (27)$$

where: L_0 = inductance of air cored coil

r_1 = radius of iron core

r_2 = mean radius of coil

l_1 = length of core

l_2 = length of coil

μeff = effective permeability of iron core (Refs. 86 & 87 list values for μ and μeff for various types of iron powders; a typical value is about 3 for μeff.

a = 0.8 when $l_1 < l_2$

and a = 1 when $l_2 < l_1$.

The iron cores in common use are about $\frac{3}{8}''$ to $\frac{3}{4}''$ in diameter and range in length from about $\frac{1}{4}''$ to 1″. An inductance change of about ± 10% when the "slug" is moved through the winding, is generally sufficient for most requirements. The dimensions required for solving eqn. (27) are available if the experimental procedure previously suggested has been carried out on air cored coils; or the whole procedure can be carried out experimentally.

Some manufacturers make up 455 Kc/s i-f transformers completely enclosed in powdered iron pots. There is often little difficulty with this construction in obtaining Q's in the order of 150. Stray capacitances are often large, however, and sometimes lead to very unsymmetrical resonance curves.

At the lower frequencies (up to about 1 or 2 Mc/s) Litz wire is advantageous for obtaining high Q values and 3, 5, 7 and 9 strands of about 44 B&S (or near SWG or B&S gauges) wire are common; Q values greater than about 120 will require some care in the transformer construction, and the size of can selected will materially affect the value obtainable (see Ref. 89 for illustration).

The required values of L_1, L_2, k etc. for developmental purposes are conveniently found using a Q meter. Methods for setting k, using a Q meter, have already been outlined in this section; further details will be given in Chapter 37.

It should be noted that some variation in k can be expected in the receiver because of regeneration, alteration of tuning slug positions etc., and it is usual to make the value somewhat less than is actually required (often about 0.8 to 0.95 depending on the receiver construction and i-f).

For the capacitance and mutual inductance coupling to be aiding, the primary and secondary windings are arranged so that if the plate connects to the start of the primary, then the grid (or diode plate) of the next stage connects to the finish of the secondary winding; both coils being wound in the same direction.

The cans to be used with i-f transformers should be as large as is practicable. They are generally made from aluminium, although copper was extensively used at one time. Cans should preferably be round and seamless. Perfect screening is not obtained, in general, and care is necessary in the layout of the various stages to ensure that the transformers are not in close proximity to one another. When mounting the transformer into a can, if there is a choice as to the position of the leads (although this is largely determined by the valve type available) it is always preferable to bring the connections for each winding out to opposite ends, as this reduces stray capacitance coupling. The effects of the coil shield on inductance and r-f resistance can be calculated (see Ref. 4, P134) and the results serve as a useful guide, but direct measurement on the complete transformer is the usual procedure. Mechanical considerations generally ensure that the can is thick enough to provide adequate shielding (for considerations of minimum thickness see Ref. 4, P135).

Methods for determining gear ratios, winding pitch and so on, for use with coil winding machines are discussed in the literature (Refs. 29, 30, 31, 32).

Finally, the measure of the success of any i-f transformer design will be how closely the predicted performance approaches the actual results obtained when the transformer is connected into the receiver.

APPENDIX 1.

Derivation of eqn. (7)

The problem is to set up conditions which will allow an i-f transformer to be designed so that the unloaded primary and secondary Q's are equal, but when the transformer is connected between a voltage source and a load, both of which damp the windings by different amounts, the resultant loaded primary and secondary Q's are to have as their geometric mean a value of Q which will meet previously determined conditions. For simplicity, it is taken that $L_1 = L_2 = L$, but the equations could be extended to include the case for $L = \sqrt{L_1 L_2}$.

Let:

Q_u be the unloaded value for the primary and secondary Q's, i.e.

$$Q_u = Q_{u1} = Q_{u2}.$$

$$Q = \sqrt{Q_1 Q_2}$$

Q_1 = loaded primary Q

Q_2 = loaded secondary Q

R_1 = parallel primary loading resistance

R_2 = parallel secondary loading resistance

and ω_o = $2\pi \times$ resonant frequency of both tuned circuits.

Then:

$$Q_1 \omega_o L = \frac{Q_u \omega_o L}{Q_u \omega_o L + R_1}$$

$$\therefore Q_1 = \frac{Q_u R_1}{Q_u \omega_o L + R_1}$$

Similarly

$$Q_2 = \frac{Q_u R_2}{Q_u \omega_o L + R_1}$$

Also, since

$$Q = \sqrt{Q_1 Q_2}$$

$$Q^2 = Q_1 Q_2 = \frac{Q_u^2 R_1 R_2}{(Q_u \omega_o L + R_1)(Q_u \omega_o L + R_2)}$$

Expanding and rearranging

$$Q_u^2 [R_1 R_2 - (Q\omega_o L)^2] - Q_u Q(Q\omega_o L)(R_1 + R_2) - Q^2 R_1 R_2 = 0.$$

Write

$$\beta = 2 [R_1 R_2 - (Q\omega_o L)^2]$$

$$\alpha = Q (Q\omega_o L)(R_1 + R_2)$$

then

$$Q_u^2 \frac{\beta}{2} - Q_u \alpha - Q^2 R_1 R_2 = 0$$

Applying the usual quadratic formula

$$Q_u = \frac{\alpha \pm \sqrt{\alpha^2 + 2Q^2 R_1 R_2 \beta}}{\beta} \quad (7)$$

The positive root is the one required for our purpose.

Evaluation of Q_u is not as involved as appears at first sight, since $Q\omega_o L$ is already known; or is required in any case for checking gain.

APPENDIX 2.

Derivation of eqn. (9)

With the primary uncoupled from the secondary the magnification factor is

$$Q_p = \frac{\omega_o L_p}{R_p}$$

where L_p = primary inductance

and R_p = the primary series resistance.

When the two circuits are coupled together and both are tuned to the same resonant frequency, the magnification factor is

$$Q_b = \frac{\omega_o L_p}{R_p + \dfrac{\omega_o^2 M^2}{R_s}}$$

where M = mutual inductance = $k \sqrt{L_p L_s}$

and R_s = secondary series resistance.

Then

$$Q_b = \frac{\dfrac{\omega_o L_p}{R_p}}{1 + \dfrac{\omega_o^2 k^2 L_p L_s}{R_p R_s}}$$

$$= \frac{Q_p}{1 + Q_p Q_s k^2}$$

where $Q_s = \dfrac{\omega_o L_s}{R_s}$ = secondary magnification factor.

Writing $Q_1 = Q_p$; $Q_2 = Q_s$ gives eqn. (9).

When $Q_p = Q_s$ then $Q_b = \dfrac{Q}{1 + (Qk)^2}$ .. (9a)

Clearly it does not matter what the values of Q_1, Q_2 and k may be, a value for Q_b is always possible (although not necessarily readable on a "Q" meter).

Since Q_1 and Q_2 can be measured directly, and the required value of k has been previously determined, then the value of Q_b is directly available. Primary to secondary spacing is adjusted until Q_b is obtained.

For the reverse case, where two circuits are coupled together and resonant at the same frequency, Q_b is read from the Q meter. Q_1 and Q_2 alone are then measured. From this k is given by

$$k = \sqrt{\frac{Q_1 - Q_b}{Q_1 Q_2 Q_b}}$$

$$\text{or} \quad \frac{k}{k_c} = \sqrt{\frac{Q_1 - Q_b}{Q_b}}$$

APPENDIX 3.

Derivation of expression for measuring high coefficients of coupling (0.1 or greater).

For the primary circuit resonant (with the secondary on open circuit)

$$\omega_o^2 L_p C_1 = 1 \quad (1)$$

With secondary short circuited and assuming that $\omega_o L_s \gg R_s$, the reflected reactance is given by

$$\frac{\omega_o^2 M^2}{j\omega_o L_s} = -j\omega_o k^2 L_p \quad (2)$$

To keep the circuit resonant at ω_0 the value of C_1 is increased by ΔC. Then from (1) and (2)

$$\omega_0^2 L_p (1-k^2)(C_1 + \Delta C) = 1 \quad \ldots\ldots (3)$$

Divide (1) by (3)

$$\frac{C_1}{(1-k^2)(C_1 + \Delta C)} = 1 \quad \ldots\ldots\ldots\ldots (4)$$

Transposing terms,

$$\Delta C = \frac{C_1 k^2}{1 - k^2} \quad \ldots\ldots\ldots\ldots\ldots\ldots (5)$$

the required equation.

The alternative form is

$$k = \sqrt{\frac{\Delta C}{C_1 + \Delta C}} = \sqrt{\frac{\Delta C}{C_2}} \quad \ldots\ldots (5A)$$

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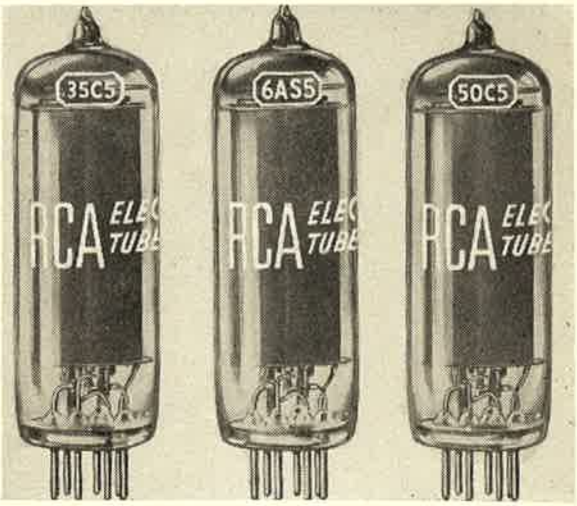
New R.C.A. Releases

Radiotron type 6AS5 — is a miniature beam power amplifier intended for use in the output stage of low cost automobile and a-c operated receivers. It is capable of delivering 2.2 watts at plate and screen voltages of 150 and 110 volts respectively.

Radiotron type 35C5 — is a miniature beam power amplifier of the heater-cathode type intended for use in the output stages of ac/dc receivers. Except for slightly higher ratings and different basing arrangements the type 35C5 is the performance equivalent of the miniature type 35B5, and within its maximum ratings, the glass-octal type 35L6-GT.

Radiotron type 50C5 — is a miniature beam power amplifier for use in the output stages of ac/dc receivers. Except for slightly higher ratings and different basing arrangements, it is the performance equivalent of the type 50B5, and within its maximum ratings of the type 50L6-GT.

Radiotron types 5691, 5692, 5693 ("Special Red" series) — are valves specifically designed for those industrial and commercial applications requiring at least 10,000 hours life, exceptional uniformity and stability of characteristics, rigidity of construction to resist shock and vibration. Their counterparts in the receiving valve line are the types 6SL7-GT, 6SN7-GT and 6SJ7 respectively.



DISCONTINUED R.C.A. TYPES.

Radiotron type 2051 — (gas tetrode) can be replaced in all applications by type 2050.

Radiotron type 1948 — miniature ionization gauge tube for use in electron microscopes.

Short-Circuit Input Admittance Data.

In Radiotronics 126 data were published giving short-circuit input admittances for a number of valve types. As further data become available for other types this will be published so that this information will be as complete as possible. The following is extracted from R.C.A. Application Note AN-127.

Short-Circuit Input Admittance Data at 100 Mc/s for type 6BJ6.

Operating Conditions:

Plate Voltage	250 volts
Screen Voltage	100 volts
Control-Grid Voltage	–1 volt
Transconductance	3800 μmhos

Short-Circuit Input Capacitance:*

Tube Operating	8.2 μμf
Tube Cutoff	6.6 μμf
Tube Cold	6.4 μμf
Capacitance Increase (cold to cutoff)	0.2 μμf
Capacitance Increase (cutoff to operating)	1.6 μμf

Short-Circuit Input Conductance:*

Tube Operating	275 μmhos
Tube Cutoff	24 μmhos
Tube Cold	18 μmhos
Conductance Increase (cold to cutoff)	6 μmhos
Conductance Increase (cutoff to operating)	251 μmhos
Grid-to-Cathode Capacitance (measured at low frequency with tube cold)	2.6 μμf

* Data for tube and socket as measured on admittance meter; socket capacitance, 0.8 μμf; socket conductance, 2.3 μmhos.

CORRECTION.

Calculation of Second Harmonic Distortion.

The formula given in Radiotronics 122 page 118 had the brackets in the denominator inadvertently omitted. The equation should read:

$$H_2 = \frac{EQ/QG - 1}{2(EQ/QG + 1)} \times 100$$

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