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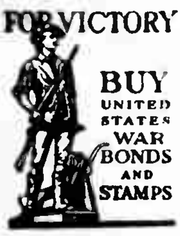


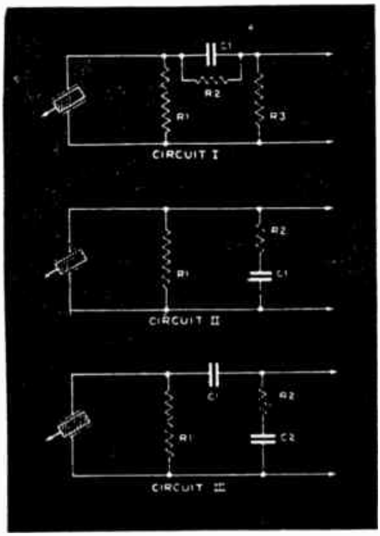
COMPENSATING TONE IN CRYSTAL PICKUPS*

Crystal phono pickups have a wide range of frequency response characteristics that are not always matched to the amplifier with which they are used. Certain compensation in the overall response can improve the performance.

In the accompanying diagrams three simple resistance-capacity compensating networks are shown. In circuit 1 the part values can be adjusted to change the response at both high and low frequencies. The shunt resistance R_1 controls the response at low frequencies and reducing its value will reduce the response. Since the crystal pickup is equivalent to a generator with an internal capacity reactance that increases as frequency increases, the voltage appearing across R_1 will be largest at low frequencies if the resistance is high. Usual values in this position are 250M to 1 meg or more. The capacitor C_1 paralleled by resistor R_2 and the resistor R_3 form a voltage divider for the output. The ratio of R_3 to $R_2 + R_3$ determines the output. The capacity of C_1 will determine the high frequency response. Making C_1 larger will improve the gain at high frequencies. R_2 can be about 100M to 500M, C_1 250 mmfd. to 1000 mmfd., R_3 1 to 5 megs. R_3 could conveniently be a potentiometer for volume control. Connect the arm and lower terminal to input of amplifier.

In circuit 2, increasing R_1 will increase the low frequency response, while increasing R_2 will increase high frequency response. The size of the capacity C_1 regulates the output as well as the high frequency response if R_2 is low.

In circuit 3 R_1 controls the low frequency response as in the other two circuits. Increasing R_2 increases the high frequency response, and increasing C_1 with respect to the sum of $C_1 + C_2$ will increase the output.



R-C networks described at left.

Any of the resistors may be made variable or several values of capacitors can be selected with a switch as a form of tone control. A control of the high frequencies is desirable in phono reproduction since it allows effective control of the scratch noise which is objectionable in some records.

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* *By courtesy "Radio Today."*

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SELL OR SWAP — Weston model 301 DC milliammeters, voltmeters, ammeters. Readrite model 430 tube tester, Weston model 506 DC milliammeters 0-100, 0-200. Want Riders manuals vol. 8 to 13; also condenser tester. C. F. McCracken, Hughes Park, Bridgeport, Pa.

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WANTED — Typewriter portable preferred; 12 or 16 ga. slide action shotgun; welding generator. For sale or trade — new Astatic DN-HZ mike and adjustable floor stand, new 12" Utah PM speaker, new signal transmitting key, stamp collection, battery charger, also 2 S127 cameras. Ralph Freyberger, Fleetwood, Pa.

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(Continued on page 13)

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Price consideration is based upon your net cost less reasonable depreciation for use, age, and condition of equipment. Inasmuch as all equipment is being purchased FOB Philadelphia, cost of packing and shipping can be shown separately so that an allowance for the costs can be made when material is accepted.

DESIGN OF BROAD-BAND AMPLIFIERS*

Simplified method for solving general problems dealing with amplifier response characteristics

There is a great deal of prior art on broad-band amplification. From a theoretical standpoint, practically every phase of this subject has been covered many times over. Most of the standard texts on radio engineering devote space to the analysis of this subject, which an engineer can utilize to solve a particular problem. There is some need, however, for a universal method of attack employing a unified and simplified form of mathematics. It is the purpose of this paper to present what is believed to be a useful method, from the engineering standpoint, for solving a large majority of broad-band problems.

This method is an approximate method. It involves the calculation of resonant circuit response on the basis of pure numbers. For such calculations, the concept of "relative frequency," as introduced by Wheeler,[1] replaces the concept of frequency; and the concept of power-factor is used. "Relative staggering" is shown to be synonymous with coupling, for staggered-stage calculations. Rule-of-thumb formulas are developed for engineering design, based on a family of universal response curves.[2]

The symbols R, L, c, f, etc., will refer to circuit parameters as is usual in the literature. Other symbols will be used to denote quantities, as follows:

A denotes amplification
p denotes power factor = 1/Q

k denotes coupling = coefficient of coupling

s denotes relative staggering (explained in Section III)

B denotes relative bandwidth

G is defined as the "gain-constant" of an amplifying stage

The subscript "zero" refers to center-frequency response (i.e., A_0 is center-frequency amplification)

A "primed" symbol refers to peak response (i.e., B' is relative peak-separation)

The subscript "—" refers to series circuits (i.e., $R_—$ is series resistance)

The subscript "||" refers to parallel circuits

d denotes differential frequency = $\pm(f - f_0)/f_0$ (on either side of resonance)

x and y denote relative frequency = 2d (refers to total differential frequency difference on both sides of resonance, i.e.: $+d - (-d)$).

An approximation developed from General Circuit Theory will be employed throughout the text:

$$(1) \quad p = R_{-}c\omega_0 = 1/R_{\|}c\omega_0 = 1/Q$$

The impedance of a series resonant circuit is:

* *By Madison Cawein in "Electronic Industries."*

$$(2) \quad Z_{-} = R_{-} + j\omega L + 1/j\omega C$$

$$(3) \quad \omega = 2\pi f = 2\pi(f - f_0 + f_0) = 2\pi f_0 (d+1) = \omega_0 (1 + x/2)$$

(3) expresses ω in terms of relative frequency. Substitute (3) in (2) and simplify, neglecting d wherever it appears in the expression $(1 + d)$.

(This assumes that d is small in comparison to unity. Whenever this approximation is used in numerator or denominator, the fact will be indicated by the symbols */ or /* following the equation.)

$$(5) \quad Z_p = (\omega M)^2 / Z_{-}$$

$$(6) \quad Z_p = k^2 \omega_0 L / (p + jx)$$

(6) is equation (5) simplified by means of (4) and substitution of $M = kL$.

Other equations relating to the equivalent circuit in Fig. 1 are:

$$(7) \quad i = g_m e_g = i_1 + i_2$$

$$(8) \quad i_{-} = e_{-} / Z_{-}$$

$$(9) \quad e_{-} = -j\omega M i_2$$

$$(10) \quad e = i_{-} / j\omega C$$

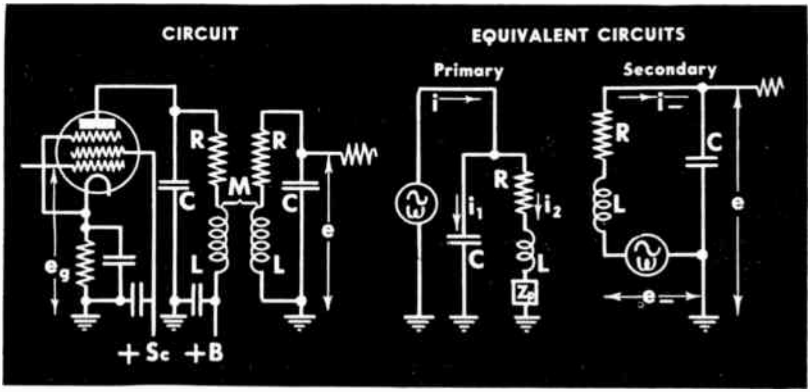

Fig. 1. Universal reference circuit and a simplified equivalent.

$$(4) \quad Z_{-} = \omega_0 L\,(R_{-}/\omega_0 L + jx) = \omega_0 L\,(p + jx) \quad */$$

Fig. 1 shows a typical double-tuned amplifier stage and its equivalent circuit. The impedance reflected in series with the primary of the transformer, from a resonant secondary of impedance Z_{-}, is:

The equivalent circuit of the primary is a constant current generator feeding two circuit branches in parallel. An impedance Z_p is reflected in series with the inductive branch and it may be proved easily that its value is as given in equation (5). Equation (7) is a statement of the approximate truth that in a pentode, considered as a con-

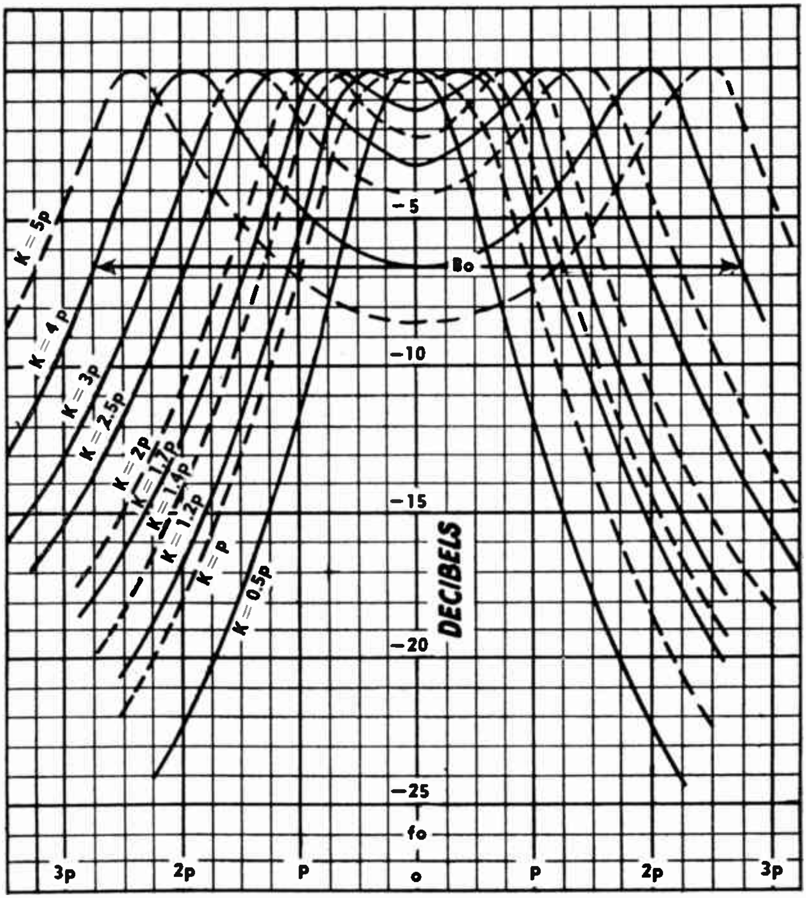

Fig. 2. Here the scale of abscissae is pure number: that is, units of p=1/Q. To convert to frequency, multiply by fo. For example, if the Q of the coils is 20, and the center-frequency is fo=10 mc, then each division (p) represents pfo cycles=0.05 x 10 mc=500 kc.

stant current generator, the current is independent of the load and is proportional to the grid voltage. The factor of proportionality is the mutual conductance.

The equivalent circuit of the secondary is a constant-voltage generator feeding a series circuit. Equation (9) states that the generator voltage is in negative quadrature with the current in the inductive branch of the primary circuit and is equal in magnitude to the product of the mutual reactance and this current. Equation (10) states that the secondary grid-voltage is the product of the secondary current and the terminating capacitive reactance, which is in parallel with this grid.

Equations (2) to (10) are merely mathematical representations of the experimental laws of electric circuits.

The voltage amplification is:

$$(11)\quad A = e/e_g = i_-/j\omega c e_g = e_-/j\omega c e_g Z_-$$
$$= -\frac{j\omega M i_2}{j\omega c e_g Z_-} \text{ or } A = -\frac{kL}{c e_g}\left|\frac{i_2}{Z_-}\right|$$

Simple calculation of i_2 from the laws of parallel circuits, and substitution from equations (3) to (10) show that:

$$(12)\quad i_2 = \frac{g_m e_g (p + jx)}{(p + jx)(jp - x) + jk^2}$$

$$(13)\quad A = \frac{k\, g_m}{\omega_0 c \sqrt{(p^2+k^2-x^2)^2+4p^2x^2}}$$

obtained by substituting (4) and (12) in (11). It shows a symmetrical function of x. A is here expressed in terms[1] of the relative frequency x, and the constant parameters p and k. Since x, p and k are pure numbers, the graph of the function A is a family of universal curves. These are plotted in Fig. 2.

There are three forms of equation (13), obtained by algebraic manipulation:

$$(13)\ (a)\quad A = kG/\sqrt{(p^2+k^2-x^2)^2+4p^2x^2}$$
$$(b)\quad A = kG/\sqrt{(p^2+k^2+x^2)^2-4k^2x^2}$$
$$(c)\quad A = kG/\sqrt{(p^2-k^2+x^2)^2+4p^2k^2}$$

$G = g_m/\omega_0 c$ is the gain-constant of the stage, and defines the absolute level of amplification. It would seem at first glance that this level is, then, inversely proportional to the frequency: this is true only because as f_0 is increased (Fig. 2) the relative bandwidth, which depends on x, increases proportionally; unless the scale of x is changed by modifying the power-factor, p. This will be clarified later.

Differentiation of (13) shows that the maximum value of A occurs at (or, can be determined by an examination of equation (13c))

$$(14)\quad p^2+x^2=k^2 \text{ or } x = B' = \sqrt{k^2-p^2}$$

which is a well-known equation defining the relative peak-separation.

The gain at the peaks

$$(x = \sqrt{k^2-p^2}) \text{ is } (15)\quad A' = G/2p$$

The gain at the center frequency $(x = 0)$ is:

$$(16)\quad A_0 = kG/(p^2+k^2)$$

The dip-to-peak ratio is:

$$(17)\quad R_0 = A_0/A' = 2pk/(p^2+k^2)$$

The simultaneous solution of (14) and (17) gives two very useful relations:

$$(18)\quad p^2 = \frac{(B')^2}{2}\,\frac{1-\sqrt{1-R_0^2}}{\sqrt{1-R_0^2}} = \frac{(B')^2 D}{2} = \frac{B_0^2\, D}{4}$$

$$(19)\quad k^2 = \frac{(B')^2}{2}\,\frac{1+\sqrt{1-R_0^2}}{\sqrt{1-R_0^2}} = \frac{(B')^2 D'}{2} = \frac{B_0^2\, D'}{4}$$

D and D′ will be called the dip-function and the conjugate dip-function, respectively. These are related by the equation:

$$(20)\quad D' = D + 2$$

regardless of the value of R_0. Thus, for over-coupled stages (R_0 is im-

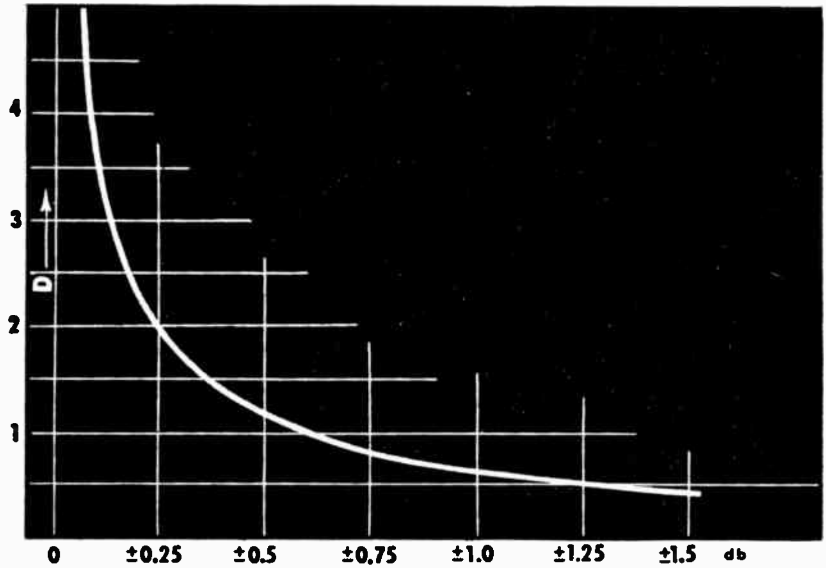

Fig. 3. Values of the Dip Function, D, as a pure number for various departures from flatness, in decibels

aginary unless k is equal to or greater than p) the following holds:

$$(21) \quad k^2/p^2 = D'/D$$

The graph of D is shown in Fig. 3. In Fig. 3 the scale of abscissae has been plotted in decibels of departure-from-flatness (± db from mean level between peaks and valley) for the convenience of those engineers who prefer to work with db-gain rather than absolute gain.

Thus, in designing an over-coupled stage of amplification, the only necessary data required is the determination of the bandwidth, B_o, and the desired departure from flatness over this band. k and p may then be calculated from (18) and (19).

The center-frequency amplification is, from (16):

$$(22) \quad A_n = G\sqrt{D'}/B_n (D+1)$$

from which the amplification is seen to be independent of center-frequency, but is inversely proportional to the bandwidth in cycles (the factors G and B_o each contain $1/f_o$, which cancels out of numerator and denominator).

Let B_o = relative bandwidth of resonance curve at the gain-level of the dip.

Then the relative bandwidth, B_o, across the valley is

$$(23) \quad B_0 = \sqrt{2}\, p'$$

This is shown in Fig. 2, curve $K = 4p$, and can be proved easily by calculating the value of x which makes $A = A_o$. This bandwidth is of some significance, as will be discussed later.

The analysis given in this section has been symmetrical, even as regards circuit components. Actually, if it is desired to get the maximum gain from a broad-band amplifier, it is usual to design the coils to resonate with the distributed capacitance on each side. These capacitances are in general slightly different. High-g_m amplifiers, such as the 6AC7, together with circuit components have a realizable minimum plate circuit capacitance of about $\mu\mu$f, and a realizable minimum grid circuit capacitance of about 16 $\mu\mu$f. The actual dissymmetry of the peaks (which did not show up in the mathematical analysis due to neglections indicated by */, /*) can be equalized by detuning the plate and/or grid circuits slightly from resonance at f_o.

It is usual to omit the plate-side damping resistor shown in Fig. 1, and to introduce all the damping in the grid side. This is allowable because of the fact that power factors are additive. As has been shown by Mountjoy,[3] the use of a grid damping-resistor only will increase the gain by several db per stage.

The design formulas are obtained in such a case by the methods outlined in this section, using different values of L, c, and R on each side of the transformer. Let p be the resulting power-factor of the grid circuit, and p_1 that of the plate circuit. Then, it can be shown that equation (13) becomes:

$$(13.1)\quad A = kG/\sqrt{(pp_1 + k^2 - x^2)^2 + x^2(p+p_1)^2}$$

$$(13.2)\quad A = kG/\sqrt{(k^2 - x^2)^2 + x^2p^2}$$

obtained when p_1 is zero, which is double peaked, quite flat, and very selective for $k > p$. Since the value $p_1 = o$ cannot be realized, the equation for a value of $p_1 = np$ will be of more practical use:

$$(13.3)\quad A = kG/\sqrt{(np^2 + k^2 - x^2)^2 + x^2p^2(n+1)^2}$$

It is possible to realize a value of $n = 0.1$. The amplification calculated from equation (13.3) will be found to be about 6 db higher than that from equation (13), for this value of n. That is, a higher gain per stage is realized by using grid damping instead of grid and plate damping of the double-tuned transformer.

The peaks occur at a value of x obtained from differentiation of (13.3):

$$(13.4)\quad x' = 0.7\sqrt{2k^2 - p^2[(n+1)^2 - 2n]}$$

A flat response is obtained by making $k = p$ in this case, which gives an overcoupled response having a departure from flatness corresponding approximately to a value of $D = 2$. The value of optimum coupling is obtained by making (13.4) equal to zero, and solving for k_o:

$$(13.5)\quad k_o = 0.7\,p\sqrt{(n+1)^2 - 2n} \doteq p/2$$

for small values of n. The relative peak separation and the relative bandwidth are, respectively:

$$(13.6)\quad B' = 0.7\,p \qquad \text{(for } k = p\text{)}$$

$$(13.7)\quad B_o = \sqrt{2}\,B' = p$$

Thus, the relation between parameters is $k = p = B_0$ for flat design when using single-sided damping.

It is of interest to the experimental engineer that the formulas for k and p given by equations (18) and (19) depend upon quantities which can be checked with a signal generator and a vacuum-tube voltmeter, i.e., on bandwidth B_0, and on a function of valley-peak gain as expressed by D.

REFERENCES

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2. F. E. Terman, "Radio Engineering," page 56, McGraw-Hill, 1937.
3. Garrard Mountjoy, "RCA Review," January, 1940, page 299.

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(Continued from page 5)

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