(19) respectively into (8) gives for the input coupling factor with microphone cable impedances neglected

$$CF_i = 10 \log \frac{\left|1 + \frac{R_g}{Z_i}\right|^2}{\left|1 + \frac{Z_m}{Z_i}\right|^2} \frac{R_{mn}}{R_g}. \quad (9)$$

6.12 Output-Coupling Factor: The expression for the output-coupling factor may be obtained by substituting the proper expressions for W_{as} and W_l into (11). Figs. 3 and 4 will aid in following the derivations of these expressions.

W_l is defined as the power that the amplifier delivers to its rated load R_l, and is given by

$$W_l = \frac{E_l^2}{R_l}. \quad (20)$$

Since

$$E_l = \frac{E_0}{\left(1 + \frac{Z_0}{R_l}\right)}, \quad (21)$$

W_l may be written

$$W_l = \frac{E_0^2}{\left|1 + \frac{Z_0}{R_l}\right|^2 R_l}. \quad (22)$$

W_{as}, the available power input to the speaker, is defined as the maximum power available from a generator whose internal impedance is a resistance equal in magnitude to the loudspeaker nominal impedance R_{sn}. W_{as} therefore is given by

$$W_{as} = \frac{E_g'^2}{4R_{sn}}. \quad (23)$$

Since the voltage delivered by this source to the speaker terminals is

$$E_s = \frac{E_g'}{\left(1 + \frac{R_{sn}}{Z_s}\right)}, \quad (24)$$

(23) becomes

$$W_{as} = \frac{E_s^2}{4R_{sn}} \left|1 + \frac{R_{sn}}{Z_s}\right|^2. \quad (25)$$

In order to find the available power input to the speaker from any source, it is only necessary to substitute the value of E_s delivered by this source into (25).

The voltage delivered by the amplifier to the loudspeaker is

$$E_s = \frac{E_0}{\left(1 + \frac{Z_0}{Z_s}\right)}. \quad (26)$$

Substituting (26) into (25) gives for the available power input to the speaker from the amplifier

$$W_{as} = \frac{E_0^2}{4R_{sn}} \frac{\left|1 + \frac{R_{sn}}{Z_s}\right|^2}{\left|1 + \frac{Z_0}{Z_s}\right|^2}. \quad (27)$$

Using the expression for W_{as} given by (27) and the expression for W_l given by (22) in (11) gives for the output coupling factor

$$CF_0 = 10 \log \frac{\left|1 + \frac{R_{sn}}{Z_s}\right|^2 \times \left|1 + \frac{Z_0}{R_l}\right|^2}{\left|1 + \frac{Z_0}{Z_s}\right|^2} \cdot \frac{R_l}{4R_{sn}}. \quad (12)$$

Intermediate-Frequency Amplifiers for Frequency-Modulation Receivers*

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Summary—In order to obtain the maximum benefits from the frequency-modulation system of broadcasting, it is necessary to give special attention to the selectivity and symmetry of the receiver intermediate-frequency-amplifier channel. Voltage feedbacks must be reduced to a minimum in order to obtain good results in mass production without resorting to some sort of stagger tuning. Selectivity and stability formulas, stabilizing methods, and methods of aligning double-tuned transformers are discussed.

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THE ADVANTAGES of fidelity and noise rejection of the frequency-modulation receiver can be lost by poor design in the intermediate-frequency amplifier. It is desirable to have a channel with a broad nose and sharp skirts that is symmetrical and will remain so, retaining its bandwidth with time and changes in temperature, and with different components and sets of tubes. It is desirable to do this at low cost. Reducing voltage feedbacks to a practical minimum is a necessary feature in obtaining the best results.

In a high-frequency-amplifier design, it is desired to obtain the wanted results with "run-of-the-mill" tubes and components with tolerances that are not too tight. To do this, variable factors should be reduced to a minimum. Excessive grid and plate loadings and voltage feedbacks should be eliminated to as large an extent as is commensurate with cost. Usually, getting these factors down to a certain stage effectively eliminates trouble due to them.

The most important factor to be considered in voltage stability is the grid-to-plate capacitance of the amplifier tube. The voltage feedback from this source usually cannot be effectively eliminated or bucked out. Also, it is not necessary to do so with present-day pentode tubes. However, this factor defines a maximum stage gain for a certain tube. In the extreme case, the tube will oscillate as a tuned-grid, tuned-plate oscillator, the grid-to-plate capacitance being the coupling means. At slightly lower gain the stage will be regenerative because the tube and its load reflects a negative impedance or negative loading into the grid circuit. Regeneration will give better gain and selectivity, but will give varying results between different tubes and components. Regeneration and degeneration (positive loading) should be avoided unless the feedback path is known and can be controlled. However, many receivers have been made with a large amount of regeneration in them and given satisfactory results. In frequency-modulation receivers, however, it is doubtful if satisfactory operations would be obtained with a highly regenerative receiver. Moreover, the good noise and fidelity characteristics, which are salient points in frequency modulation, will be much better with a symmetrical intermediate-frequency amplifier. Any regeneration will give an unsymmetrical curve. Symmetry can be obtained with a channel having some regeneration, but it is doubtful if it will remain symmetrical with time and temperature. All these factors make it necessary to design the channel to minimize the effect of the grid-to-plate capacitance.

To insure against instability due to this factor, it is necessary to choose amplifier tubes that have sufficient mutual conductance and small enough grid-to-plate capacitance to give the desired gains with stability. The maximum allowable inductance in double-tuned transformers is given by the following formula (derived using the circuit of Fig. 2(a)), assuming only inductive coupling:

$$L = \frac{A}{\omega Q}\sqrt{\frac{2}{g_m \omega C_{gp}}}$$

where L is the inductance at which oscillation will just occur; Q is the $\omega L/r$ of the tuned circuits; g_m is the mutual conductance of the tube used; C_{gp} is the grid-to-plate capacitance of the tube; and $f = \omega/2\pi$ is the operating frequency. A is a factor depending on the coupling factor K (K is unity at critical coupling). A is obtained from the circuit equations after a number of assumptions that are accurate in most cases.

When

$$K = 1, \quad A = 1.26$$

$$K = 0.9, \quad A = 1.22$$

$$K = 0.8, \quad A = 1.2$$

$$K = 0, \quad A = 1.0.$$

From this equation the maximum gains can be derived for a number of desired conditions.

1. When $K=0$ (single-tuned transformer),

$$\text{maximum gain} = \sqrt{\frac{g_m}{\pi f C_{gp}}}.$$

2. When $K=1$,

$$\text{maximum gain} = 0.63\sqrt{\frac{g_m}{\pi f C_{gp}}}.$$

3. When $K=1$ and it is desired to have no oscillation with the circuits in the plate and grid of the tube tuned but the circuits coupled to them detuned,

$$\text{maximum gain} = 0.5\sqrt{\frac{g_m}{\pi f C_{gp}}}.$$

Experience has shown that, if the gain is held within this latter figure, regeneration can be made negligible. It is probably desirable to allow something for variations in Q and inductance.

The next step is to get the above gain with stability, and with an arrangement that will give the least variation between tubes. One of the possible feedback paths in a single stage is coupling between plate and grid circuits in a cathode impedance. The mutual cathode impedance is not always negligible, even with the cathode pin grounded. At high frequencies the impedance of the cathode lead may be important. The effect of this can be minimized if the tube has two cathode connections and the grid circuit is returned to the ungrounded cathode pin. Also, in alternating-current–direct-current receivers with the cathode above ground, the cathode impedance becomes important. Calculation will show that, in a circuit represented by Fig. 1(a), Z will reflect a positive load if it is an inductance, and a negative load if it is a capacitance. The negative loading of the grid-to-plate capacitance can be bucked out by a cathode inductance of proper size at one frequency. This is difficult to hold, and also over-all stability is not helped by a cathode above ground.

In Fig. 1(b) the tuned circuits are returned to cathode. C_1 and C_2 are capacitances from grid and plate to ground. By a $Y-\Delta$ transformation, these capacitances, together with Z, form an impedance in one of the legs across the grid-to-plate capacitance. If Z is a capacitance, this will be a capacitance which will increase the regeneration. This circuit appears often in alternating-current–direct-current receivers and can cause trouble, even though Z is a large bypass. The capacitances C_1 and C_2 can become large, due to the fact that the cans shielding the transformers are grounded. This type of regeneration can operate over more than one stage. The circuit of Fig. 1(c) is a combination of Figs. 1(a) and (b), but is usually better than Fig. 1(b) since C_2 is within the tube and is small.

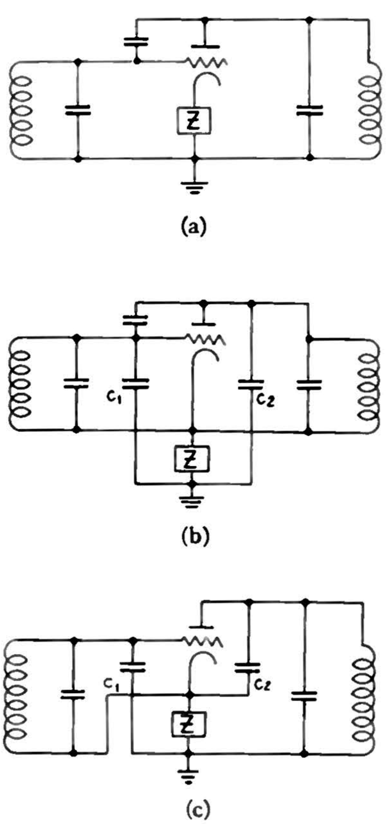


Fig. 1—Feedback paths in a single stage with the cathode ungrounded.

fact that they largely eliminate spurious circuits tuned at or near the channel center frequency.

It was found that a circuit like that of Fig. 2(b) gave better results than that of Fig. 2(a). Capacitive coupling is indicated in Fig. 2(a) to show that it has not been effectively eliminated. Fig. 2(c) shows a method of grounding the tuned circuit and also eliminating cathode-lead inductance, if the frequency or component spacing or both make this feedback path effective. This has been tried at 100 megacycles with excellent results.

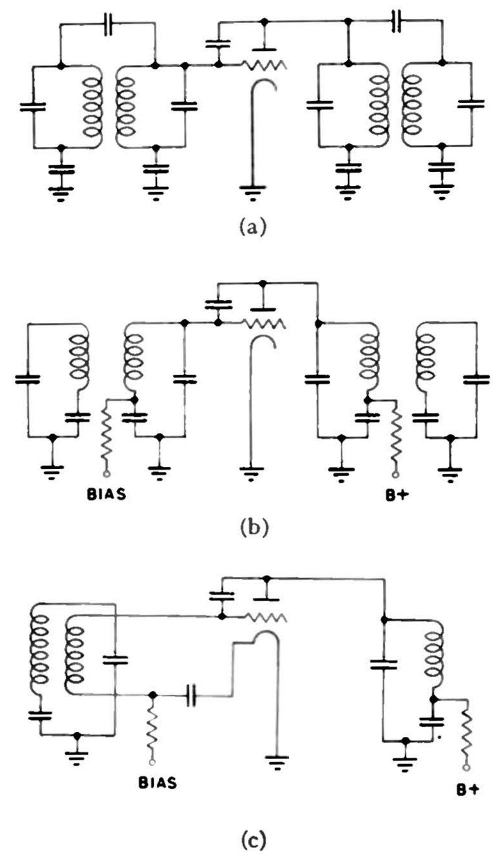


Fig. 2—Methods of improving stability at higher frequencies.

Over-all regeneration in alternating-current–direct-current receivers is often largely due to the common cathodes by-passed to ground by one capacitor. An inductance in series with the by-pass to tune it to the channel frequency has been used with success. At the higher frequencies it is better to isolate the cathodes and have separate by-passes. More by-passes can be used without exceeding the maximum safe value, since they can be small.

At the higher frequencies it has been found by experiment that better results can be obtained if the capacitive coupling in the transformers is kept to a minimum. Grounding the tuned circuit also helps stability. This can be done by putting the by-pass in the tuned circuit, as shown in Fig. 2(b). It is best to keep the path into which the tuned circuit current flows small. The effectiveness of these methods is probably mainly due to the

Fig. 3 shows schematically the connections for a dual channel for broadcast (455 kilocycles) and frequency modulation (8.3 megacycles).

There are other precautions necessary for stability. It may be necessary to isolate high-voltage and bias leads at the transformers. It is necessary to be sure that supposedly cold leads (high-voltage, automatic-volume-control, heaters, etc.) do not pass near hot points at the front and rear end of the channel. Isolation resistors and capacitors should be right at the point to be isolated in order to be most effective. Automatic-volume-control leads should be isolated at the detector end of the channel. It is best not to lay these supposedly cold leads together in a cable. Cathode bias should be used as little as possible, since it increases the common cathode-lead inductance.

In frequency-modulation channels it is very desirable to have as flat a nose as possible, and also good skirt selectivity. Any regeneration will hurt the ratio of nose to skirt. This makes nearly perfect voltage stability more necessary than in the broadcast band. The ratio of bandwidth at 1000 times down to that at 2 times down depends only on the number of double-tuned transformers for a given coupling factor, and not on the channel center frequency, except to a minute degree.

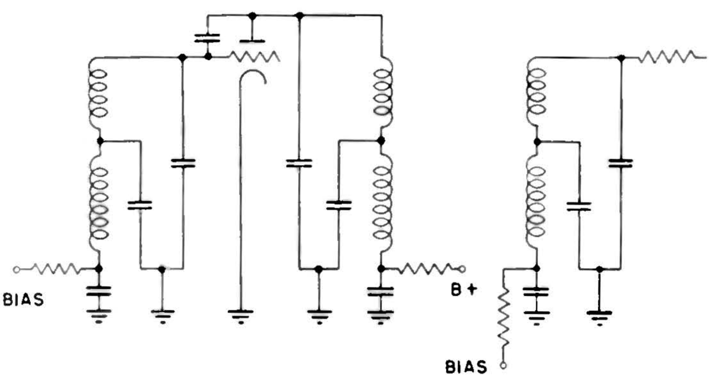


Fig. 3—Basic circuit of dual intermediate-frequency transformer.

Circuit Q's and channel center frequency will determine the actual values of bandwidth at 2 and 1000 times down. The relation between bandwidth and times down is[1]

$$Y = \left[\frac{(4x^2 + 1 - K^2)^2 + 4K^2}{(1 + K^2)^2}\right]^{n/2}$$

where

Y = number of times down
$X = Q(f_d/f_0)$
$f_d = \frac{1}{2}$ bandwidth
f_0 = center frequency
K = the coupling factor
n = the number of double-tuned circuits.

From this equation, by solving for $X = Q(f_d/f_0)$ we can find the ratio between two values of f_d(f_a and f_b):

$$\frac{f_a}{f_b} = \sqrt{\frac{\sqrt{a^{2/n}(1 + K^2)^2 - 4K^2} - (1 - K^2)}{\sqrt{b^{2/n}(1 + K^2)^2 - 4K^2} - (1 - K^2)}}$$

where f_a and f_b are half-bandwidths at "a" and "b" times down. For three transformers, if $K = 1$ (critical coupling),

$$\frac{f_{1000}}{f_2} = 3.6.$$

For $K = 0.8$ with three transformers,

$$\frac{f_{1000}}{f_2} = 4.1.$$

A feature that would be desirable is the ability to trim the circuits with an unmodulated signal and always obtain the same results, a flat, symmetrical curve. Unfortunately, this cannot be done with double-tuned circuits. The curve will be lopsided on one side or the other, depending on whether the circuits are originally below resonance or above. In a circuit like that of Fig. 2(b), with critical coupling, a random tuning will give the required flat nose but the curve will be unsymmetrical. A complicated procedure of starting some trimmers in and some out will give a practically symmetrical curve, but this method is not foolproof. This necessitates alignment with a sweep oscillator and an oscilloscope. Best results will be obtained if each stage is successively tuned on the oscilloscope for symmetry.

The use of single-tuned transformers would do away with this difficulty, but the ratio of f_{1000}/f_2 for three single-tuned transformers is 13, which is not very good. Another way to achieve practically symmetrical trimming with an unmodulated signal is to undercouple the double-tuned transformers somewhat and begin alignment of each stage with one coil detuned as far as possible. The curves of Fig. 4 show why this is so. These curves are based on the following factor, which is the denominator of the equation for the ratio of output voltage to input voltage of one stage:

$$Z = K^4 + 2K^2(1 - 4x_1x_2) + (1 + 4x_1^2)(1 + 4x_2^2).$$

This can be used in finding maxima and minima, since the numerator of the equation is constant.

$$K = \text{coupling factor}$$

$$x_1 = \frac{Qf_1}{f_0}, \qquad x_2 = \frac{Qf_2}{f_0}$$

where f_1 and f_2 are off-resonance frequencies of primary and secondary circuits.

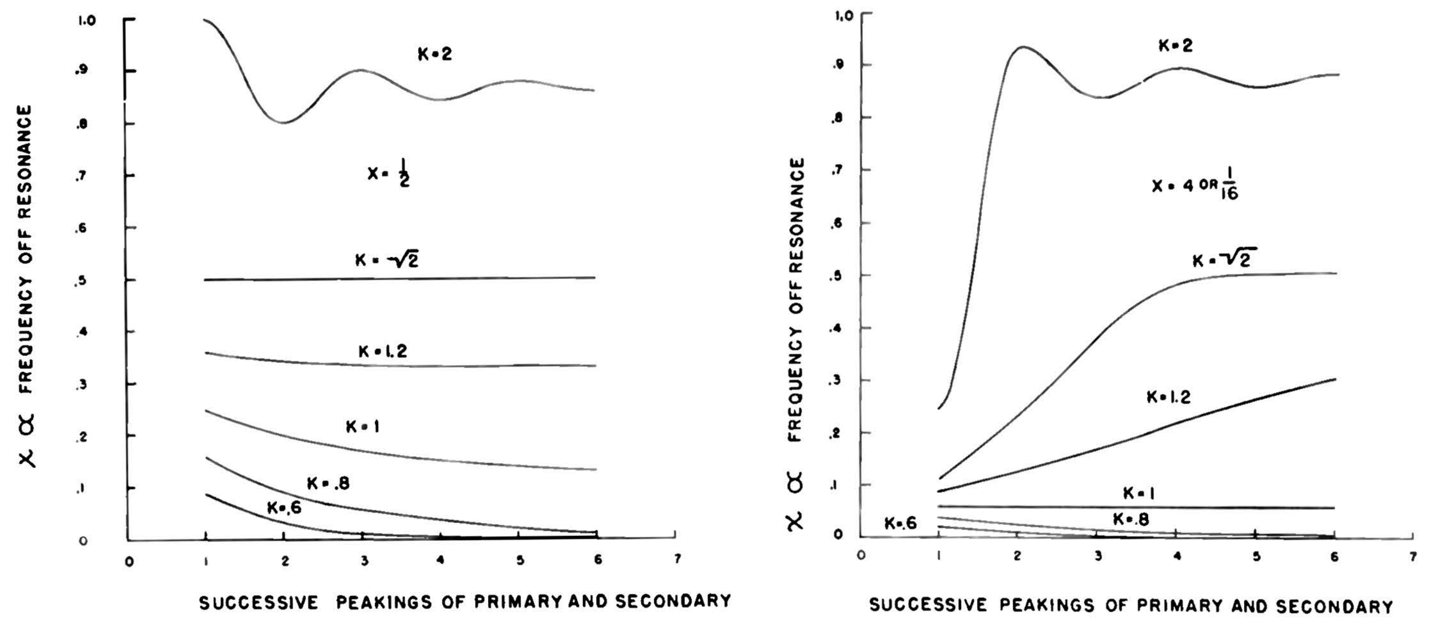
This factor is symmetrical in x_1 and x_2, so that any analysis starting with the primary would apply to a series of operations starting with the secondary. It is desired to find the minimum of this factor (or the maximum transfer) as the primary and secondary are successively peaked. To do this, first assign a value of X to x_2 (this is the amount the secondary is off resonance before the primary is peaked). Then evaluate the derivative of Z with respect to x.

$$\frac{dz}{dx_1} = -8K^2X + 8x_1(1 + 4X^2).$$

Set this equal to zero and solve for x_1:

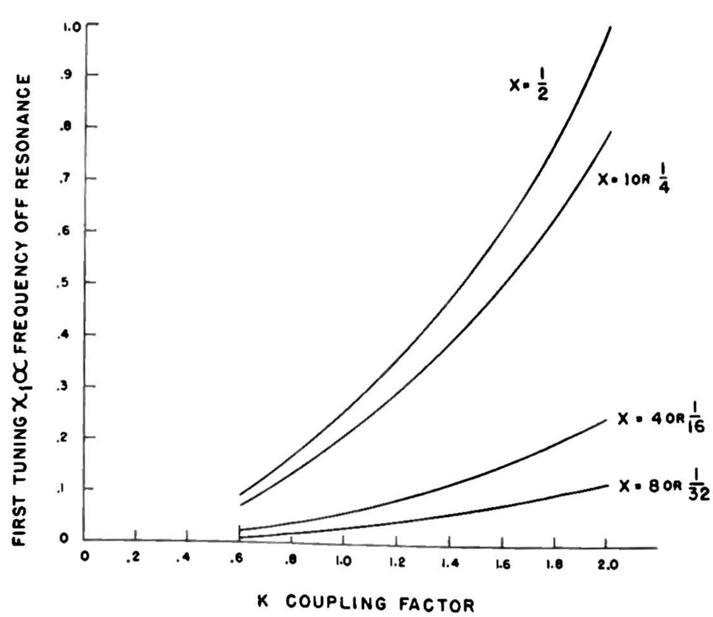
$$x_1 = \frac{K^2X}{1 + 4X^2}.$$

[1] J. J. Adams, "Undercoupling in tuned coupled circuits to realize optimum gain and selectivity," Proc. I.R.E., vol. 29, pp. 277–279; May, 1941.



(a) (b)

This is the amount off resonance the primary will be after it is peaked once. From the equation it can be seen that, if the secondary was on resonance ($X=0$), the primary will trim to resonance ($x_1=0$). The primary will be farthest off resonance when trimmed once, if the secondary is originally at $X=\pm\frac{1}{2}$. If X is larger than one-half, the primary will trim closer to resonance. The larger X is, the closer the primary will tune to resonance. If now the primary is left at the value x_1 and the secondary is tuned, x_2 becomes the variable with x_1 having the value from the above equation. The minimum is found at a new value for x_2. This is then repeated for successive tunings of the primary and secondary. The curves of Fig. 4 show the results. Notice that the convergence for $K=1$ is very slow; so slow, in fact, that resonance would never be reached in practice. For overcoupled transformers, successive trimmings usually give a divergence from resonance. However, if the secondary is set far off resonance (X large), the first peaking of the primary will bring it near resonance. The indicated procedure, which is well known, is to set the secondary far off resonance, then peak the primary and then the secondary. Neither trimmer should be touched after this.



(c)

Fig. 4—Curves showing the amount the coils are off resonance after successive peakings for difference values of K and X.

These methods of obtaining the desired selectivity and gain characteristics are not the only ones available but appear to be the most straightforward. Stability, of course, can be obtained by the simple expedient of having lower sensitivity. However, sensitivity, up to a certain point, is worth the effort to obtain it.