



# SPECIALIZED TELEVISION ENGINEERING

TELEVISION TECHNICAL ASSIGNMENT

RECEIVERS, RADIO FREQUENCY AMPLIFICATION

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740E

## RECEIVERS, RADIO FREQUENCY AMPLIFICATION

### FOREWORD

This assignment is a basic text on receivers. The analysis of r.f. amplifiers, local oscillators, mixer tubes, and i.f. amplifiers affords the student the fundamental knowledge required to design and adjust such equipment.

In addition, there is presented material on the tracking of ganged circuits and on alignment procedure, so that the theory, design, and adjustment of radio receivers are all developed. Finally, actual commercial receiver circuits are analyzed to afford the student an opportunity to see how the theory is applied in actual practice.

Although this assignment deals primarily with standard broadcast receivers, the material presented is of direct value in the study of f.m. receivers. This will be apparent when studying the assignment on f.m.; much of the material will be taken bodily from the present text. The reason, of course, is that such circuits as r.f. and i.f. amplifiers, mixers, etc., are the same for both types of receivers, although the frequencies involved are different. This brings out a fundamental fact of knowledge: a basic understanding is the main trunk from which the specialized facts branch out; if you have a basic understanding of an art, you can understand and appreciate the developments that arise in the course of time.

It is with this thought in mind that we offer you this text. Study it not only for the information it affords on standard broadcast receivers, but also for the groundwork it—and other previous assignments—lay for the proper understanding of subsequent assignments.

E. H. Rietzke,  
President.

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# RECEIVERS, RADIO FREQUENCY AMPLIFICATION

## GENERAL CONSIDERATIONS

Radio frequency amplification is used in radio receivers and transmitters, and is therefore a very important topic. It represents in general, amplification over a narrow range of frequencies, such as a broadcast carrier and its audio side bands, and hence involves a resonant circuit responsive to but a narrow frequency band, although this band may be adjustable to any part of the spectrum desired.

*RECEIVER CIRCUITS.*—The earliest broadcast receivers consisted of a tuned circuit connected to an antenna, and feeding a nonlinear element in the form of a crystal rectifier. The latter demodulated the incoming wave and produced an audio output that was used to feed a pair of headphones. In more elaborate sets a vacuum-tube audio amplifier was added to afford loud-speaker volume.

The next step was the regenerative detector, a rather unstable device in practice, which afforded, however, high gain in one stage, and also was capable of demodulating the incoming wave and thus furnishing audio output to a pair of headphones or to an audio amplifier and loud-speaker.

A subsequent development was the addition of tuned r.f. amplifier stages (t.r.f.) to an ordinary vacuum-tube detector (usually of the grid-leak type), and the use of neutralization to prevent regeneration when three or four r.f. stages were employed. The r.f. amplifier stages not only increased the sensitivity of the set, but its

selectivity as well, and thus permitted satisfactory reception as the broadcast spectrum began to fill up with stations.

However, in 1920 Armstrong had already introduced his superheterodyne principle, and some sets were produced quite early in the broadcast art. But it was not until the 1930's that the superheterodyne began to supplant the t.r.f. type of set, and today the superheterodyne type of receiver reigns practically supreme. Tuned r.f. stages may or may not be used in such a receiver, depending upon its price, but in general most of the knowledge and experience acquired from the inception of broadcasting until the present is of practical value today, and in particular, r.f. and i.f. amplifiers are an important item in any course on receivers.

*THE SUPERHETERODYNE PRINCIPLE.*—It will be of value to study first in block diagram form the superheterodyne principle. In Fig. 1 are shown the essential components of such a receiver. The incoming wave induces a voltage in the antenna, which is then amplified by one or two r.f. amplifier stages. (In cheaper sets this stage may be omitted, and the voltage applied directly to the mixer stage through selective circuits.)

A local oscillator in the set develops a relatively strong signal voltage at a frequency slightly different from and usually higher than the incoming signal. The two voltages produce, among other products, a voltage in the output of the mixer stage, whose frequency is the difference between that of the

local oscillator and that of the incoming signal. The output load

duction of the intermediate frequency voltage from the interaction

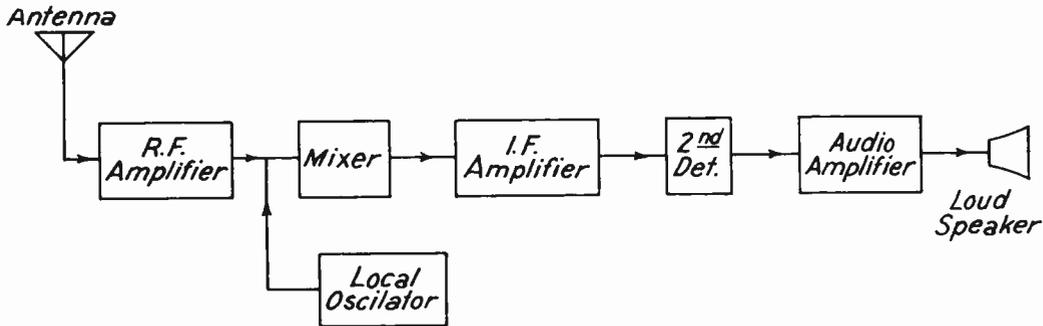


Fig 1.—Essential components of a superheterodyne receiver.

of the mixer is tuned to this difference frequency (intermediate frequency) and hence is responsive to it rather than to the other components developed in the mixer stage, so that the following i.f. stages amplify this i.f. voltage.

There has thus been made a frequency conversion of the initial carrier and its side bands to a lower frequency. The i.f. amplifier operates at this fixed frequency only, and can therefore be adjusted for optimum gain and selectivity. In order to select one station from another, only the r.f. and local oscillator stages are tuned so that the local oscillator is always the same frequency interval above (or below) the incoming wave, and the output of the mixer is always of the same intermediate frequency.

The output of the i.f. amplifier is then demodulated by a second detector, whose rectifying action is exactly similar to that of the detector in a t.r.f. receiver. The resulting audio output is then amplified by an audio amplifier and fed to a loudspeaker.

**MIXER OPERATION.**—The pro-

of the incoming signal and the local oscillator voltage in the mixer depends upon the nonlinear action of this tube. (Nonlinearity refers to the nonproportionality of the current to the voltage in the tube.) In many types of mixer tubes the  $G_m$  of the tube is variable and is a function of the applied grid voltages. Thus, if the local oscillator voltage is applied to one grid of the tube, it varies the  $G_m$  of the latter at its frequency, and this variation modifies the incoming signal in such manner as to produce new frequency components; in particular, a component of intermediate frequency.

Mixer stages may comprise a variety of nonlinear circuit elements. Thus, at very high frequencies, crystal detectors (rectifiers) are used for this purpose; diode rectifiers can be used at lower frequencies, and at frequencies up to two hundred megacycles or more, special multi-grid tubes, as well as the ordinary triode, will function satisfactorily. Particularly in the broadcast band (540 to 1,600 kc) a multi-grid tube may be employed in

such manner that the cathode in combination with two of the grids functions as a local oscillator, and the resulting oscillatory current enroute to the plate of the tube is further modified by the incoming signal impressed upon a third grid to produce so-called mixing in the electron stream with the production of an i.f. voltage in the output.

In order to understand mixing, it is also necessary to appreciate what is not mixing. Suppose two generators of somewhat different frequencies  $f_1$  and  $f_2$  are connected in series to the vertical deflection plates of a cathode ray oscilloscope. As shown in Fig. 2, the two waves combine on the screen so as to be

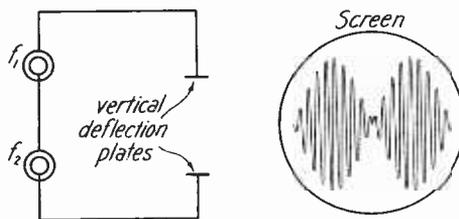


Fig. 2.—Simple superposition of two waves of frequencies  $f_1$  and  $f_2$  on an oscilloscope screen.

more or less additive in deflection for a certain number of cycles, then subtractive as the higher frequency wave shifts ahead of the other in phase, then additive once more as the continuing phase shift brings the first into phase once more with the second, and so on.

The variation in amplitude occurs at a rate equal to the difference in the two frequencies, or  $f_1 - f_2$  (assuming  $f_1 > f_2$ ). Thus it might seem that such simple superposition of the two waves produces a beat frequency ( $f_1 - f_2$ ). This, however, is not the case;

there is no energy present at this intermediate frequency; a wave analyzer would indicate the presence of  $f_1$  and  $f_2$  only, and would give zero reading when its dials were set at the frequency ( $f_1 - f_2$ ). The appearance of the wave of Fig. 2 is an optical illusion, there is no beat frequency in it.

On the other hand, if a rectifier element is placed in series with the two generators, then the wave shape shown in Fig. 3 is obtained. Observe that now a low-frequency variable component is

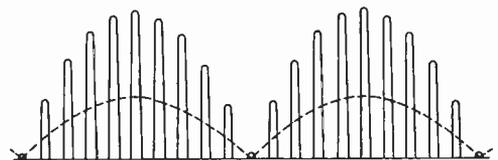


Fig. 3.—Action of rectifier on simple superposition of two waves.

present, as indicated by the dotted lines. If either frequency alone were impressed, so that the amplitude were constant, then a d.c. component would be produced by the rectifier.

If the amplitude is caused to vary such as by the interaction of the two waves, then the rectified d.c. becomes a variable wave; i.e., develops an a.c. component, as shown in Fig. 3, of a frequency equal to ( $f_1 - f_2$ ). This is called the *difference beat frequency*. At the same time the original two frequencies  $f_1$  and  $f_2$ , and harmonics thereof, are also present. However, what is not apparent from Fig. 3, and requires a mathematical proof to demonstrate its presence, is a component of frequency ( $f_1 + f_2$ ), and known as a *summation beat frequency*.

As will be shown in another assignment, the process of modulation of a carrier wave of frequency  $f_c$  by a modulating wave of frequency  $f_m$  consists in impressing the two waves upon a nonlinear element, whereupon there is obtained from it—among other things—waves of frequency  $f_c$ ,  $(f_c + f_m)$ , and  $(f_c - f_m)$ . These three represent the amplitude-modulated carrier; here the upper side band or summation beat frequency  $(f_c + f_m)$  is desired just as much as the lower side band or difference beat frequency  $(f_c - f_m)$ , or the carrier frequency  $f_c$ .

In the case of the superheterodyne receiver, the only component desired is the difference beat between the incoming signal and the oscillator output, or  $(f_o - f_s)$ . However, if the input signal is modulated, and contains side bands  $(f_s - f_m)$  and  $(f_s + f_m)$  as well as the carrier  $f_s$ , then the difference beat in the output of the mixer will consist of frequencies

$$(f_o - f_s) = f_1$$

i. f. carrier

$$[f_o - (f_s - f_m)] = f_1 + f_m$$

i. f. upper side band

$$[f_o - (f_s + f_m)] = f_1 - f_m$$

i. f. lower side band

In short, the output of the mixer consists of a wave modulated exactly like the incoming signal, but at the lower carrier frequency  $f_1$ .\* There will also be a large

\*It will be observed that the upper side band of the incoming signal becomes the lower side band of the i. f. carrier, and vice-versa; but this is not of

number of other components, such as  $f_o$ ,  $f_s$ ,  $(f_s - f_m)$ ,  $(f_s + f_m)$ ,  $(f_o + f_s)$ ,  $(f_o + f_s - f_m)$ ,  $(f_o + f_s + f_m)$ , etc. However, these are all at frequencies much higher and far removed from the three difference beat frequencies given above, so that if the output load circuit is made resonant and hence a high impedance to the difference beat frequencies, it will develop a high voltage only with regard to these frequencies, and effectively short circuit all the other components. The appreciable difference beat voltages can then be impressed on the grid of the following i. f. amplifier stage, and thus amplified for subsequent demodulation in the second detector.

The rectifier discussed above is rather an extreme case of non-linearity in that its conductivity changes abruptly from a value other than zero in one direction to zero in the opposite direction of current flow. The device is said to be a *unilateral* conductor. However, successful mixing and the production of difference beat components can be produced by less extreme types of nonlinear devices. For example, the ordinary multi-element vacuum tube, like the triode, usually exhibits a somewhat variable conductivity or amplification factor, or both, even in the conducting direction. Hence, even if the a. c. signals representing the incoming signal and that of the local oscillator are superimposed on the normal d. c. component of the tube's plate current of such magnitude that the total current at

any practical importance as regards the shape of the wave, owing to the symmetrical distribution of the side band on either side of the carrier.



the fundamental frequency, it is clear that the isolation afforded between the two circuits by the r.f. grounded grids of the 6L7-G mixer is a distinct advantage. It is also to be noted that "pulling" is further minimized by having a relatively high intermediate frequency; i.e., by having the oscillator frequency appreciably different from the signal frequency. There are certain other advantages as well as disadvantages in having a high intermediate frequency, and these will be discussed farther on.

**CONVERTER STAGE.**—It is also possible to combine the oscillator and mixer functions in one tube, with a resultant saving of one tube in the set. Such a stage is known as a *converter*, and is illustrated in Fig. 5. The particular tube shown, type 6K8-G, is particularly

structural details at the right in Fig. 5.

It will be observed that the oscillator and mixer plates are on opposite sides of the cathode, but that the oscillator grid acts on both sides of the cathode: on the left side as an oscillator control grid and on the right side as an injector grid. Note that the screen grids act as a shield between the injector and signal grids to decouple them.

In another design, a 6SA7 pentagrid converter tube was developed. The circuit is shown in Fig. 6. The oscillator portion consists of the cathode, grid #1 acting as the control grid, and grids #2 and 4 as the plate. However, the circuit is of the grounded plate type in that grids #2 and 4 are by-passed to ground by a 0.1- $\mu$ f

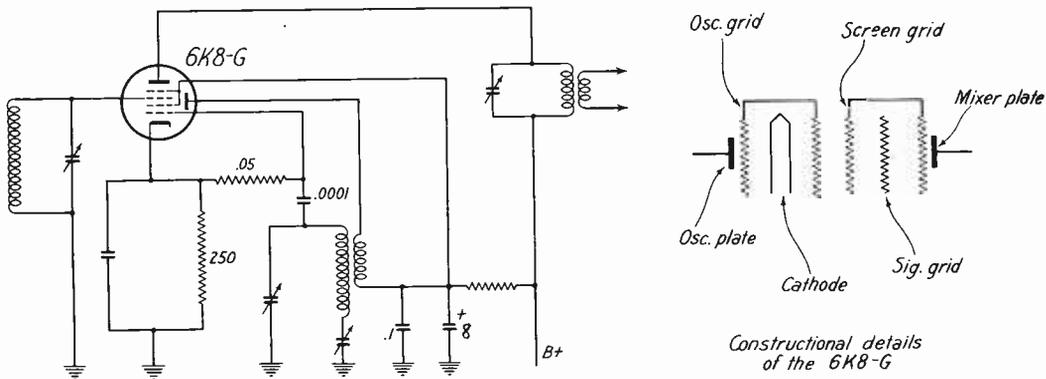


Fig. 5.—6K8-G converter, showing its constructional details and the associated circuit.

well adapted to operate not only at broadcast frequencies, but also at the higher frequencies of an all-wave receiver. It has greater stability with respect to "pulling" as compared to some of the older converter tubes owing to the use of two plates, as shown in the con-

structional details at the right in Fig. 5. It will be observed that the oscillator and mixer plates are on opposite sides of the cathode, but that the oscillator grid acts on both sides of the cathode: on the left side as an oscillator control grid and on the right side as an injector grid. Note that the screen grids act as a shield between the injector and signal grids to decouple them.

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signal grid. Thus no "pulling," even at the higher frequencies, is

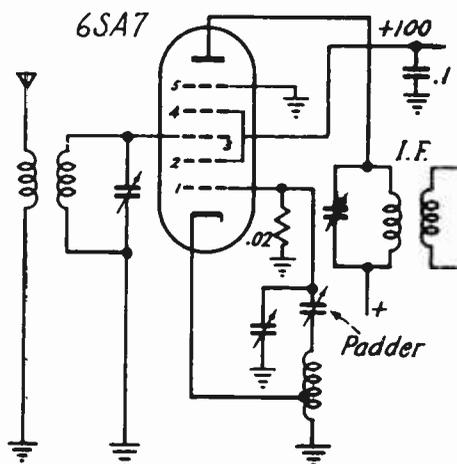


Fig. 6.—6SA7 pentagrid converter.

experienced, and the signal tube can therefore act as a combined oscillator and mixer stage.

**IMAGE FREQUENCY.**—The phenomenon of "pulling" has been cited as one reason for choosing as high an intermediate frequency as possible. Another reason for such choice is image frequency considerations. This will be made clear from the following simple example. Suppose the incoming signal frequency is 1,000 kc, and the oscillator frequency is 1,175 kc. Then the intermediate frequency is  $1,175 - 1,000 = 175$  kc. This is a rather low value, and will facilitate oscillator pulling.

However, suppose there is another station on the air whose frequency is 1,350 kc; that is, 175 kc above that of the oscillator, and therefore  $2 \times 175 = 350$  kc above that of the first station. If the signal from the second station comes through the r.f. selective circuits to the mixer, it also will beat with the oscillator to produce a 175-kc intermediate frequency, and hence

its signal will be amplified by the i.f. amplifier as well as that of the first-mentioned station. The result will be two signals in the loudspeaker.

The frequency, 1,350 kc, of the second station is called the image frequency for that particular oscillator setting. It represents a kind of reflection of the 1,000-kc station about the 1,175-kc oscillator frequency, and is—as mentioned above—twice the intermediate frequency above that of the first station. In order to minimize such unwanted response, the r.f. tuned circuits in the receiver should be as selective as possible, so that if they are tuned to the 1,000-kc station, they are greatly detuned to the 1,350-kc station, and a high "image rejection" is obtained.

However, the circuits must not be too selective, or they will appreciably cut the side bands of the wanted station and thus decrease the fidelity of reception. On the other hand, if the image station is close to the receiver, the signal input may be so great as to give appreciable output even though the circuits have the maximum permissible selectivity. Thus, receivers within the 250-millivolt per meter contour of a broadcast station are assumed to be within the so-called "blanket area" of the station, and reception without appreciable interference is not considered possible.

Outside of this blanket area, the amount of interference can be reduced to a lower level—for a given selectivity of the receiver—by using a higher intermediate frequency. The reason for this is illustrated in Fig. 7. For the given shape of overall selectivity curve shown, the greater the oscillator

frequency  $f_0$ , the greater will be the image frequency  $f_2$  (where  $f_2 - f_0 = f_0 - f_1$ ), and the smaller will be the image response  $E_2$  compared to the desired response  $E_1$ . Of course, if a flat-topped response, as shown by the dotted-line rectangle in the figure, could be obtained, then the difficulty of image rejection would not arise, but actual response

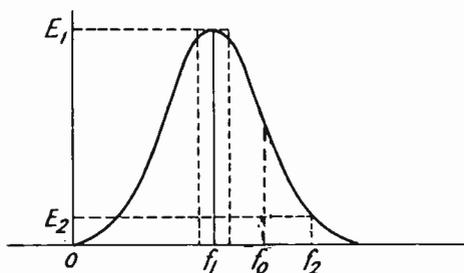


Fig. 7.—Relation of image response to desired response.

curves are of the more usual resonance shape, and increasing  $f_0$  then decreases the image response.

There are, of course, definite limits to the intermediate frequency  $f_0 - f_1$  that can be chosen. If  $f_0$  is made too high compared to  $f_1$ , the range of oscillator tuning will be markedly different from that of the r.f. circuits, and tracking of the two will be too difficult to achieve. (Tracking will be discussed farther on.) Even in the range of permissible values of the intermediate frequency there may be certain values that are undesirable. If a strong low-frequency station happens to be on that frequency, a sensitive receiver may pick up its signal directly on its i.f. amplifier.

To obviate such possibilities, present-day values for the intermediate frequency center around 455 kc, especially for all-wave receivers.

This frequency, and neighboring values, are free from the objections indicated above; for the normal selective circuits employed, image rejection is adequate, and yet the oscillator can be made to track satisfactorily with r.f. circuits.

**R.F. AMPLIFIERS**—The smaller and cheaper receivers generally do not employ an r.f. amplifier stage. The antenna is coupled to the mixer or converter through a tuned circuit. (Antenna coupling is discussed in the assignment on receiving antennas.) In the more expensive and larger receivers, an r.f. stage is employed. This not only permits an additional tuned circuit to be employed (in the plate circuit), so that image rejection is increased, but the unilateral (one-way) action of the r.f. amplifier tube prevents the oscillator energy from "leaking" into the antenna and producing reradiation with consequent interference to other receivers.

An r.f. amplifier stage is shown in Fig. 8. The primary  $L_p$  is

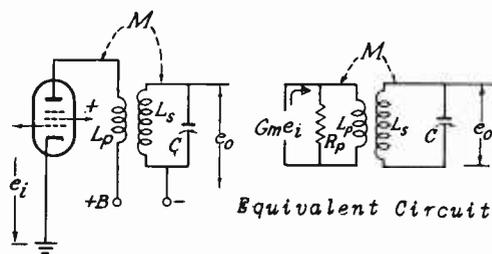


Fig. 8.—R.F. amplifier stage.

untuned; the secondary  $L_s$  is tuned by variable capacitor  $C$ . This circuit will be analyzed farther on; suffice it to state here that the introduction of the screen-grid tube for this type of amplifier stage not only minimized regenerative feedback through the grid-to-plate inter-

electrode capacity, but also increased the Q and selectivity of the circuit, owing to the higher plate resistance of the pentode tube. A higher plate resistance has less shunting effect upon the tuned transformer circuit, as will be shown subsequently.

**I. F. AMPLIFIERS.**—As explained previously, the i.f. amplifier is a fixed-frequency amplifier. It can

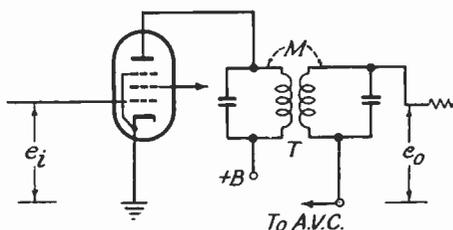


Fig. 9.—I.F. amplifier stage.

therefore consist of more complex tuned circuits which have higher selectivity together with more uniform amplification within the pass band.

In Fig. 9 is shown a typical i.f. amplifier stage. The i.f. transformer T usually consists of two identical coils wound on a bakelite or similar form, spaced so as to have a suitable amount of magnetic coupling between the two. Each coil is individually tuned to the intermediate frequency by the associated capacitor. In some designs fixed coils and variable (trimmer-type) capacitors are employed; in other designs C is fixed and L can be varied by means of an iron dust core slug that is adjusted by a screw feed to enter more or less into the bakelite cylinder on which the coil is wound. These are illustrated in Fig. 10.

The interaction of the two tuned circuits owing to the mutual inductance between them can produce two resonant peaks in the secondary output voltage  $e_o$  instead of one. This depends upon the degree of coupling between the coils: loose coupling produces but one resonant peak;

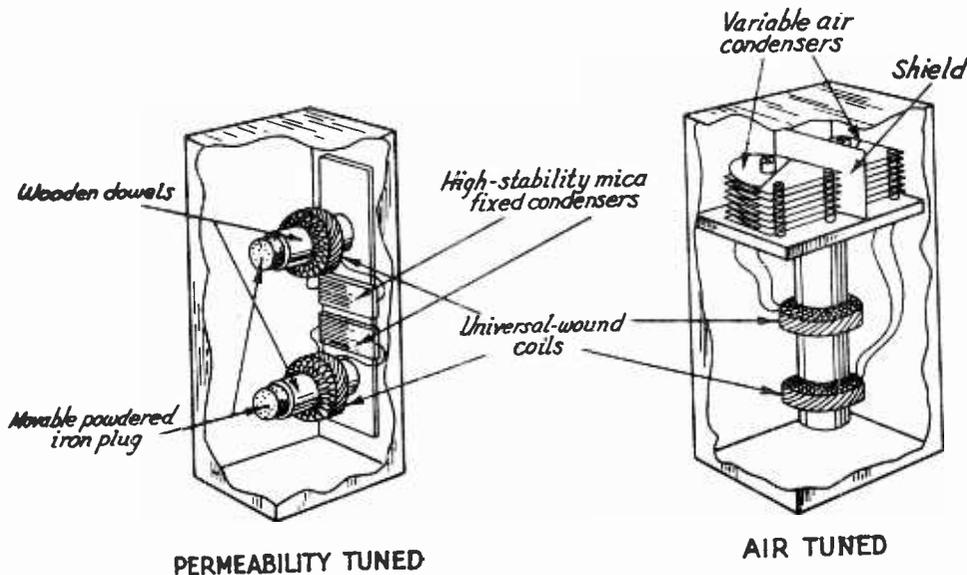


Fig. 10.—Constructional details of two types of i.f. transformers.

tight coupling produces two such peaks.

This is illustrated in Fig. 11.

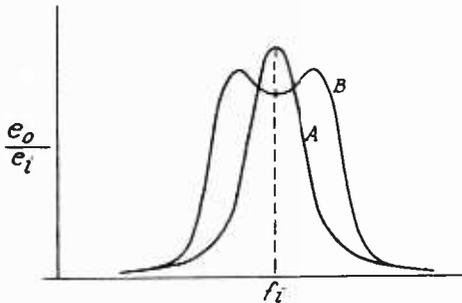


Fig. 11.—Effect of over coupling upon the response of an i.f. amplifier.

Curve A represents the single-peak response that is obtained when the coupling is loose; curve B, the double-peak response when the coupling is tight. Note that for curve B there is a more-or-less flat response over a range of frequencies centered on  $f_i$ , and then the response drops off sharply.

This approximates the ideal shape of response, mentioned previously, much closer than does curve A. Thus, the upper side bands are transmitted as well, if not better, than the i.f. carrier, but adjacent-channel signals are attenuated sharply. Most of the adjacent-channel selectivity is obtained in the i.f. amplifier, whereas image rejection is obtained in the tuned circuits of the antenna coupling unit and the r.f. amplifier (if used).

**AUTOMATIC VOLUME CONTROL.**—The a.v.c. (automatic volume control) is a variable negative bias voltage, obtained from the output of the second detector or associated rectifier, which varies the  $G_m$  of the tubes to which it is applied. The action is as follows: if the incoming signal

is strong, the rectified d.c. output of the a.v.c. diode is large and produces a large negative bias on the grids of the tubes it controls. This decreases the d.c. component of their plate current and their transconductance, which in turn reduces the amplification of the tubes with regard to the incoming signal.

Thus a strong incoming signal reduces the receiver gain and thereby limits the audio output. A weak incoming signal produces less a.v.c. negative bias, the  $G_m$  and gain of the amplifier stages are higher, and the audio output is therefore comparable to that for a strong signal. In this way the audio output of the receiver is more nearly uniform in level, and such a characteristic is particularly useful in areas where the received signal is subject to fading. With a.v.c. present, the only indication of fading is the increase in static and other background noises as the signal-to-noise level decreases; the increase in receiver gain compensates to a large extent for the drop in signal strength. The a.v.c. circuit will be discussed further in the next section.

**DETECTION.**—After the signal has been amplified to the desired degree in the i.f. amplifier, it is detected in exactly the same way as the amplified r.f. signal in the case of a t.r.f. receiver. Any non-linear circuit element can be employed, but the diode tube is preferred to a triode or pentode detector principally because of the ease with which an output voltage negative to ground can be obtained for a.v.c. purposes.

In Fig. 12 is shown a diode detector that also furnishes a.v.c. voltage. Capacitors  $C_p$  are used to

image rejection  
Adj. ch. select.

Drop Det

by-pass the i.f. currents to ground. The audio components owing to the interaction of the side bands and the carrier, as well as a d.c. component produced by the detector from the carrier, flow through  $R_1$  and  $R_2$  in series;  $R_1$  in conjunction with the lower  $C_f$  acts as an R-C filter for any i.f. current that is not by-passed by the upper  $C_f$ .

The audio components can pass through coupling capacitor  $C_2$  to produce an audio signal voltage across  $R_4$ , which is then amplified by the audio amplifier. At the same time,  $R_3$  and  $C_1$ , of high time constant, act as an R-C filter to eliminate the audio components, but do

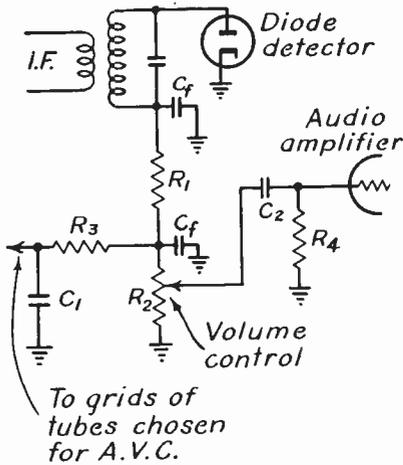


Fig. 12.—A.V.C. employing a single diode.

permit the d.c. component of voltage produced by rectification of the carrier across  $R_2$  to appear across  $C_1$ ; since this component varies but slowly if fading of the carrier takes place. There thus appears across  $C_1$  a negative voltage to ground which is employed for a.v.c. purposes.

An objection to this circuit is

that for audio frequencies,  $R_3 C_1$  as well as  $C_2 R_4$  shunt  $R_2$ . The shunting effect of  $C_2 R_4$  cannot be avoided since coupling to the grid of the audio amplifier is desired, but the shunting effect of  $R_3 C_1$  (essentially  $R_3$ , since  $C_1$  has a reactance at audio frequencies much less than the resistance of  $R_3$ ) can be eliminated if a separate diode element is employed for a.v.c. purposes.

The reason why the shunting effect on  $R_2$  is undesirable is that it occurs only for the audio-frequency components of the rectified output, and not for the d.c. component produced by rectification of the carrier alone. This means that the impedance to the audio component is lower than that to the d.c. component. At high levels of modulation approaching 100 per cent, clipping of the inward modulation peaks will occur, with consequent distortion of the audio output.

To avoid this, the circuit of Fig. 13 can be employed. Here a

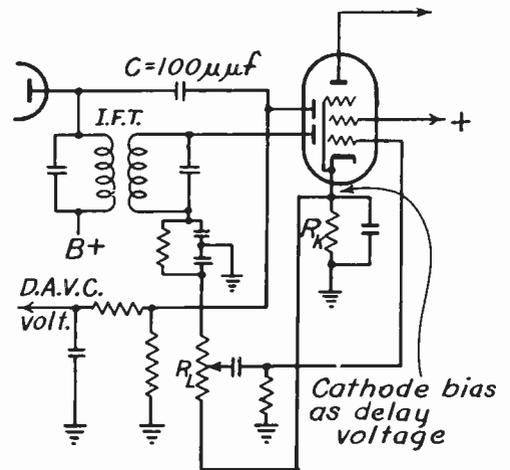


Fig. 13.—Delayed a.v.c. using two diodes.

double diode, such as the 6H6 is used. The upper plate rectifies the direct output of the i.f. amplifier tube for a.v.c. purposes. The lower plate is fed through the i.f. transformer and furnishes the audio component to  $R_L$  and the pentode portion of the duo-diode-pentode tube combination. In this way the shunting effect of the a.v.c. network on audio load resistor  $R_L$  is avoided, and distortion at high modulation levels is decreased.

**DELAYED A. V. C.**—An interesting feature of this circuit is the so-called delayed a.v.c. feature. Delay here does not refer to time, but to level of signal. It will be observed that the cathode of the tube is positive to ground by the voltage drop produced by the plate current in  $R_K$ . It is further to be noted that  $R_L$  is connected directly to the cathode instead of to ground.

As a result, there is no d.c. bias between  $R_L$  and the cathode, so that the smallest voltage produced in the i.f. circuit will be rectified by the lower diode plate and will produce audio output across  $R_L$ . On the other hand, the a.v.c. network is connected to ground, and is therefore negative to the cathode by the voltage drop in  $R_K$ . Consequently, unless the signal output of the i.f. amplifier is sufficient to overcome this bias, no rectification will occur in the upper diode plate, and no a.v.c. voltage will be produced.

The net result is that the a.v.c. is inoperative until the incoming signal attains a certain magnitude, and the receiver operates at fixed gain and variable output up to this value. The a.v.c. action is said to be "delayed" until a certain level of input signal is reached.

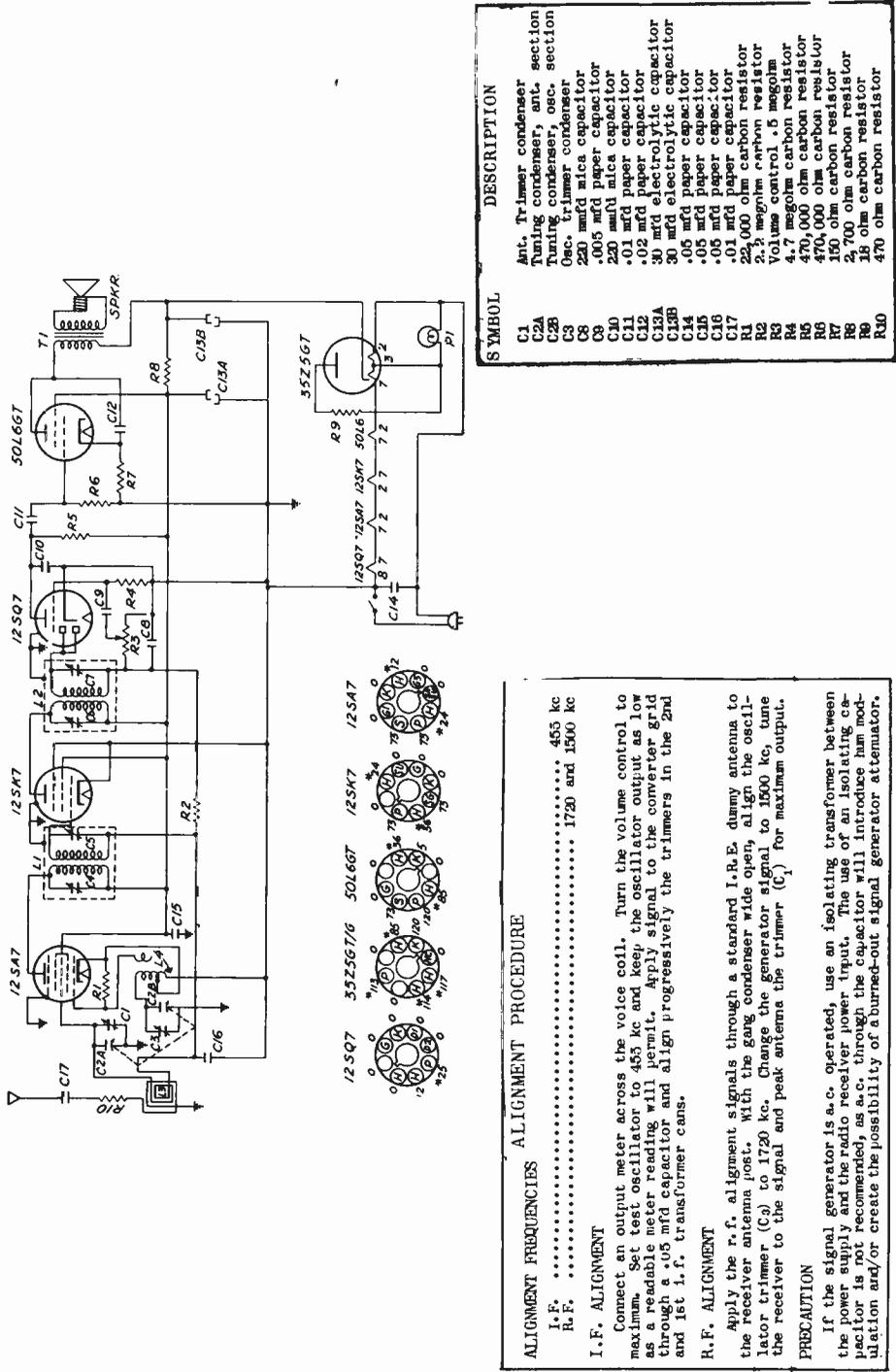
When the signal reaches this level, which is of a magnitude sufficient to furnish adequate output at a satisfactory signal-to-noise ratio, the a.v.c. action takes hold, and the gain is varied thereafter to maintain essentially constant output even though the carrier level may increase in variable amounts from this initial value.

It is to be observed, in passing, that when a.v.c. is employed the variation in receiver gain is produced by contrary variation in carrier level. On the other hand, the depth of modulation determines the audio output; this is not affected by the a.v.c. action, and thus loud and quiet passages in music, for example, will be reproduced as such by the receiver, whereas irrelevant variations in signal strength owing to fading and the like, will be minimized by the a.v.c. action.

## REPRESENTATIVE RECEIVER CIRCUITS

Before proceeding to a detailed analysis of r.f. and i.f. amplifiers, it will be of value to study some typical receiver circuits in order to note the application of some of the principles mentioned in the foregoing sections. In this way a picture of the overall circuit will be obtained.

**TABLE-MODEL RECEIVER.**—In Fig. 14 is shown a representative midget or table-model receiver. This type of receiver is in the low-price field, and does not have all the features of the more expensive console models. Observe that the antenna couples directly into the 12SA7 converter tube, the tuned input circuit consisting of the loop



SYMBOL	DESCRIPTION
C1	Ant. trimmer condenser
C2A	Tuning condenser, ant. section
C2B	Tuning condenser, osc. section
C3	Osc. trimmer condenser
C8	220 mfd mica capacitor
C9	.005 mfd paper capacitor
C10	230 mfd mica capacitor
C11	.01 mfd paper capacitor
C12	.02 mfd paper capacitor
C13A	30 mfd electrolytic capacitor
C13B	30 mfd electrolytic capacitor
C14	.05 mfd paper capacitor
C15	.05 mfd paper capacitor
C16	.05 mfd paper capacitor
C17	.01 mfd paper capacitor
R1	224,000 ohm carbon resistor
R2	2.3 megohm carbon resistor
R3	4.7 megohm carbon resistor
R4	470,000 ohm carbon resistor
R5	470,000 ohm carbon resistor
R6	150 ohm carbon resistor
R7	2,700 ohm carbon resistor
R8	15 ohm carbon resistor
R9	15 ohm carbon resistor
R30	470 ohm carbon resistor

(Courtesy Service)

Fig. 14.—G. E. YRB 82-1, 67-1 and 67-2 a-c/d-c five-tube receiver with loop primary and external antenna containing a .01-mfd blocking capacitor and 470-ohm damping resistor to attenuate any resonant peaks in antenna.

**ALIGNMENT PROCEDURE**

**ALIGNMENT FREQUENCIES**

I. F. .... 455 kc  
 R. F. .... 1720 and 1500 kc

**I. F. ALIGNMENT**

Connect an output meter across the voice coil. Turn the volume control to maximum. Set test oscillator to 455 kc and keep the oscillator output as low as a readable meter reading will permit. Apply signal to the converter grid through a .05 mfd capacitor and align progressively the trimmers in the 2nd and 1st I. F. transformer cans.

**R. F. ALIGNMENT**

Apply the r. f. alignment signals through a standard I. R. E. dummy antenna to the receiver antenna post. With the gang condenser wide open, align the oscillator trimmer (C3) to 1720 kc. Change the generator signal to 1500 kc, tune the receiver to the signal and peak antenna the trimmer (C1) for maximum output.

**PRECAUTION**

If the signal generator is a.c. operated, use an isolating transformer between the power supply and the radio receiver power input. The use of an isolating capacitor is not recommended, as a.c. through the capacitor will introduce hum modulation and/or create the possibility of a burned-out signal generator attenuator.

and tuning condenser C2A and trimmer condenser C1. Provisions are made for the connection of an external antenna. To prevent any resonant peaks in the antenna pickup from causing undue response at certain frequencies, a damping resistor R10 of 470 ohms is employed in conjunction with a .01- $\mu$ f blocking capacitor C17.

The oscillator section of the 12SA7 converter has some interesting features. Tank coil L4 is tuned by C2B and trimmer C3. Grids 2 and 4 are at a.c. ground owing to .05  $\mu$ f by-pass capacitor C15, and are therefore essentially connected to the bottom end of L4, which is grounded. The cathode is tapped up on L4, and the oscillator grid (#1) could therefore be connected to the top end of L4 for regenerative feedback. Instead, the grid is connected to an auxiliary coil which acts as a capacity winding to couple it to the top of the tank coil. This capacity acts in conjunction with grid leak R1 of 22,000 ohms to establish grid bias as well.

The plate of the 12SA7 is coupled to i.f. transformer L1, which in turn feeds the single stage i.f. amplifier employing the 12SK7 tube and i.f. transformer L2. The latter in turn is coupled to the two diode plates (connected in parallel) of the 12SQ7 tube. Thus, although the tube has two diode elements, they are used as one, and furnish both audio output from R3 through C9 to R4, and a.v.c. through R2 and C16. Note that a.v.c. is furnished to the signal grid of the converter tube as well as to the grid of the i.f. tube.

The grid of the 12SQ7 is connected to the top end of R4, and thus receives the audio signal,

which is then amplified and fed to the 50L6GT power output tube. Observe that the heaters of all tubes are in series across the line, and that the line voltage is directly rectified by the 35Z5GT tube to furnish the "B" voltage. The plate of the 50L6GT receives very little filtering—that of C13B—, but the use of C12 in conjunction with R7 acts as a feedback loop to inject hum voltage in the cathode circuit of the 50L6GT in opposition to that injected into the plate circuit. This is sufficient to decrease the hum output to a satisfactorily low level, particularly in view of the fact that this is a high-level stage, the loudspeaker has very little low-frequency response, and finally, in a screen-grid tube, minor variations in the plate voltage have very little effect upon the plate current. In addition, the capacitor prevents the pentode from accentuating the high-frequency response owing to its by-passing action.

The rest of the circuits, including the screen of the 50L6GT, receive filtered current through the R-C filter C13B, R8, and C13A. This filter is sufficient to provide satisfactory plate voltage to the rest of the circuits. In passing, it is to be noted that often some of the best engineering effort goes into the design of a small receiver of this type, and its simplicity, cheapness, and apparent crudeness of components actually represent great skill and ingenuity on the part of the designer in obtaining the results he does at such low cost.

*CONSOLE-TYPE RECEIVER.*—In Fig. 15 is shown the circuit diagram for an RCA Victor console-type receiver—Models 59AV1 and 59V1. This is an a.c. operated combination radio and



automatic phonograph, and incorporates a superheterodyne radio with a self-contained loop antenna and automatic tuning, as well as an automatic record changer with a record capacity of fourteen 10-inch or twelve 12-inch records.

It employs nine tubes, including an r.f. amplifier, converter, one i.f. amplifier, a second detector and first audio stage, a phase inverter, and a push-pull power output stage. In addition it has an audio pre-amplifier stage for the phonograph pickup, and a transformer-type power supply. There are three bands:

("A")	540 - 1,600 kc	(Broadcast)
("B")	2.3 - 6.3 mc	(Short Wave)
("C")	9.4 - 15.4 mc	

The push buttons are employed for the broadcast band only; this band, as well as the other two, may be tuned manually. In order to enable the circuit to be analyzed more readily, it has been redrawn for the broadcast range, with the switches eliminated, as shown in Fig. 16. Only the stages up to and including the second detector have been included; the rest of the circuit will be analyzed farther on.

Refer first to the r.f. stage. The loop antenna is connected to a parallel-tuned circuit, which is in turn resistance-capacity coupled to the grid of the 6SG7 tube. This is the only selectivity furnished for image rejection. The plate circuit is coupled to the signal grid of the 6SA7 converter tube through what is known in television as a series-peaking circuit. Peaking coil 73 forms a pi-configuration in conjunction with the output capacitance of the 6SG7 and the input capacitance of the 6SA7 (both shown in dotted lines). This pi-configuration acts as a low-pass filter whose pass band, how-

ever, extends up to 15.4 mc! Thus the plate circuit of the r.f. stage acts as a broad-band amplifier stage that requires no tuning and covers the entire range of radio frequencies handled by the receiver.

Such a broad-band stage has, of course, no image selectivity and serves merely as an amplifier. Since amplification is not necessary in the broadcast range, it is clear that the main function of this stage is to amplify the input signal for the short-wave and "c" bands, since in this range the set noise—particularly that of the mixer stage—begins to be appreciable, and pre-amplification of the input signal helps to increase the signal-to-noise ratio. However, the r.f. stage is of value even in the broadcast range to obviate reradiation of local-oscillator energy via the loop antenna.

The converter stage has features similar to that of the table model discussed previously. Thus, a grounded-plate oscillator is employed so that the oscillator plate (grids #2 and #4 of the 6SA7 tube) act as a shield between the signal and oscillator grids and thus prevent "pulling." The oscillator grid is coupled to the oscillator tank circuit through a 56- $\mu$ f capacitor and 56,000-ohm grid leak.

The tank circuit has some interesting features. The tank inductance is the iron-cored coil labelled B11. To the top end of this is coupled the oscillator grid; to the bottom (ground) end is connected the oscillator plate through a 15- $\mu$ f electrolytic condenser. For regenerative feedback and oscillation, the cathode should be connected electrically somewhere along the tank coil between the grid and plate.



that the main diode load is the 270,000-ohm resistor paralleled by the 180- $\mu$ f i.f. by-pass condenser. Shunting this combination is an R-C network which furnishes a.v.c., and audio output to the grid of the triode portion of the 6SQ7 tube.

grid of the top 6F6 tube direct, and also feeds the grid of the inverter 6SQ7 tube through a resistance voltage divider of 2.2 megohms in series with 39,000 ohms. The .005- $\mu$ f blocking condenser enables this grid to receive a lower bias (voltage drop

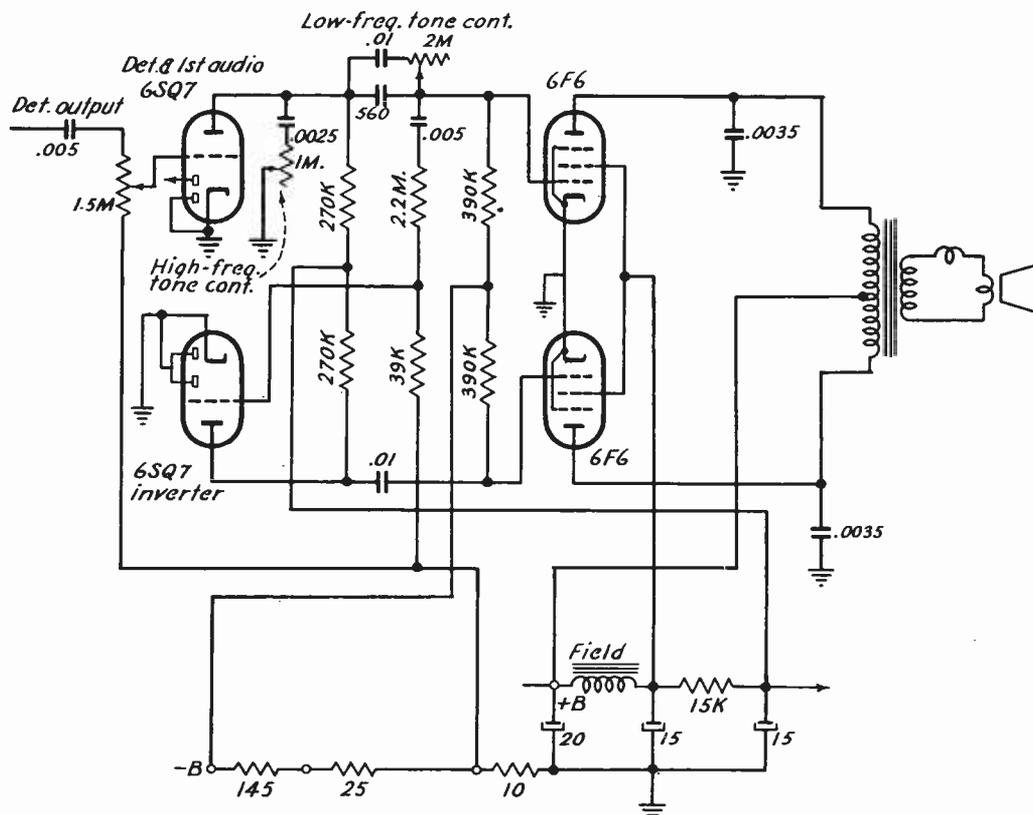


Fig. 17.—Audio section of RCA Victor Model 59AV1, 59V1 receivers.

The 1.5-megohm potentiometer acts as a volume control, but has two taps to which are connected series R-C networks to modify the audio response. These act as fixed tone controls in addition to two variable tone controls (shown in Figs. 15 and 17) for adjusting independently the high and low audio frequencies.

The audio circuit is redrawn in Fig. 17. The plate of the detector and first audio 6SQ7 tube feeds the

across the 10-ohm resistor) than the grid of the top 6F6 tube receives, in that the latter is connected to -B, where maximum negative bias to ground is obtained.

Owing to the above-mentioned voltage divider, the audio voltage applied to the grid of the inverter tube is  $39,000 / (39,000 + 2,200,000) = 1/57.3$  of that applied to the grid of the top 6F6 tube. However, the voltage gain of the inverter tube is

approximately 57.3, so that the inverter output to the grid of the lower 6F6 is  $57.3/57.3 = 1$  or equal to that to the top 6F6, but 180 degrees out of phase. Hence the two 6F6 tubes are capable of operating in push-pull to feed the loudspeaker load.

Finally, some additional features to note are the use of 0.0035- $\mu$ f capacitors from each 6F6 plate to ground, to prevent the plate-to-plate impedance from building up to too high a value owing to leakage reactance of the output transformer

RECEIVER CALCULATIONS

*R.F. AMPLIFICATION.*—The r.f. amplifier stage is illustrated once again in Fig. 18. In (A) is shown the actual vacuum tube circuit. By the equivalent plate-circuit theorem, the vacuum tube can be replaced by an equivalent generator whose generated voltage is  $\mu e_{g1}$ , and whose internal resistance is  $R_p$ , as shown at (B): to this circuit can be applied Thevenin's theorem.\* As viewed from the secondary terminals,

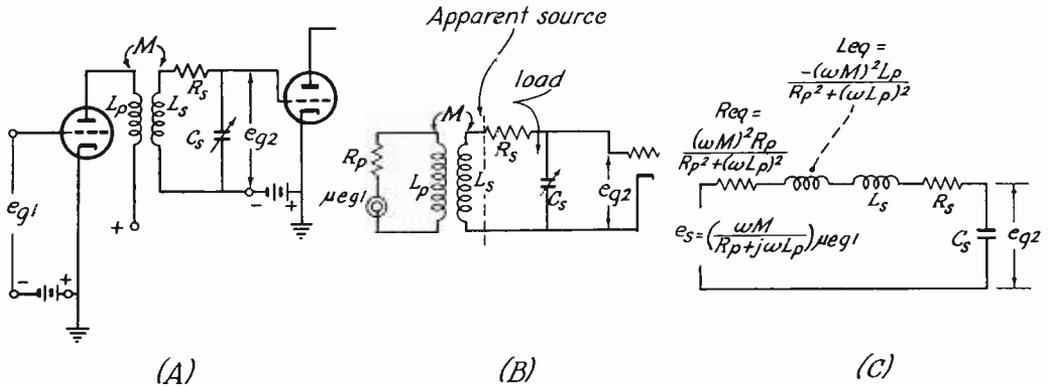


Fig. 18.—R.F. amplifier stage, showing actual circuit, equivalent plate circuit, and equivalent secondary circuit.

and inductance of the loudspeaker voice coil, since this tends for distortion and shrill reproduction from pentode tubes; the use of an additional R-C filter of 15,000 ohms and 15  $\mu$ f (see Fig. 15) for the plate circuits of the two 6SQ7 tubes; and the use of maximum and not-too-well filtered voltage from a point ahead of the filter choke (loudspeaker field) to supply the push-pull output stage, since any hum practically cancels in a push-pull stage.

with the secondary load  $C_s$  removed, the impedance of the apparent source—as measured to the left of the dotted line in (B)—is,

\*Thevenin's theorem states that any circuit can be divided into any two desired parts, one of which represents an apparent generator, and the other the apparent load. With the circuit broken apart, the apparent generator has an apparent generated voltage equal to that appearing across the break, and an apparent internal impedance equal to that measured or calculated "looking" into the two terminals of the break.

as mentioned in a previous assignment,

$$Z'_p = \frac{(\omega M)^2}{Z_p} = \frac{(\omega M)^2}{R_p + j\omega L_p} \quad (1)$$

Multiply numerator and denominator by the conjugate ( $R_p - j\omega L_p$ ) and obtain

$$\begin{aligned} Z'_p &= \frac{(\omega M)^2 (R_p - j\omega L_p)}{R_p^2 + (\omega L_p)^2} \\ &= \frac{(\omega M)^2 R_p}{R_p^2 + (\omega L_p)^2} - \frac{j(\omega M)^2 \omega L_p}{R_p^2 + (\omega L_p)^2} \quad (2) \end{aligned}$$

The quantity  $(\omega M)^2 R_p / R_p^2 + (\omega L_p)^2$  is free of  $j$ , and represents a resistance; the quantity

$$\frac{-j(\omega M)^2 \omega L_p}{R_p^2 + (\omega L_p)^2}$$

represents the reactance of a negative inductance of magnitude

$$\frac{-(\omega M)^2 \omega L_p}{R_p^2 + (\omega L_p)^2}$$

Both of these quantities are coupled into the secondary circuit, as is shown in (C).

The apparent generated voltage  $e_s$  is that induced into the secondary circuit when its load  $C_s$  is disconnected, so that the secondary is essentially open-circuited. In this case the primary current is simply

$$i_p = \frac{\mu e_{g1}}{R_p + j\omega L_p}$$

This current induces a voltage in the secondary of magnitude

$$e_s = \omega M i_p = \left( \frac{\omega M}{R_p + j\omega L_p} \right) \mu e_{g1} \quad (3)$$

Thus, finally, the circuit, as viewed from the secondary side,

appears as shown in Fig. 18(C).

The advantage of the circuit shown in (C) as compared to those shown in (A) or (B) is that all the circuit elements appear in a simple series connection. Hence, to find the output voltage  $e_{g2}$ , simply multiply  $e_s$  by the ratio of the reactance of  $C_s$  to the total impedance of the circuit. In formula form this is:

$$e_{g2} = e_s \frac{(1/j\omega C_s)}{R_{e_{q.}} + R_s + j\omega(L_s + L_{e_{q.}}) + (1/j\omega C_s)} \quad (4)$$

If now the values for  $e_s$ ,  $R_{e_{q.}}$ , and  $L_{e_{q.}}$  are substituted in Eq. (4), the value of  $e_{g2}$  can be found in terms of the fundamental circuit constants  $\omega$ ,  $L_p$ ,  $L_s$ ,  $R_p$ ,  $R_s$ ,  $C_s$ ,  $\mu$ , and the input voltage  $e_{g1}$ . The actual substitution results in a very complicated expression. This can be simplified by making suitable further substitutions, and by ignoring certain terms that are negligibly small in normal practice.

In general,  $R_p$  is much greater than the reactance  $\omega L_p$ , so that the latter may be ignored in comparison with the former in such expressions as  $[R_p^2 + (\omega L_p)^2]$ ; i.e.,

$$R_p^2 + (\omega L_p)^2 \approx R_p^2 \quad (5)$$

Thus, consider the coupled-in resistance

$$R_{e_{q.}} = \frac{(\omega M)^2 R_p}{R_p^2 + (\omega L_p)^2}$$

If  $(\omega L_p)^2$  is ignored, this quantity reduces to

$$R_{e_{q.}} \approx \frac{(\omega M)^2 R_p}{R_p^2} = \frac{(\omega M)^2}{R_p} \quad (6)$$

This indicates that the higher the plate resistance  $R_p$  of the tube, the

lower is  $R_{e_q}$ , and the lower is the damping in the circuit of Fig. 18(C), so that higher gain and selectivity may be obtained. Since a pentode inherently has a higher  $R_p$  than a triode, the former may be expected to give greater selectivity, and that is one reason why it is preferred to a triode.

The coupled-in reactance becomes—in view of Eq. (5)—

$$j\omega L_{e_q} = \frac{-j\omega L_p (\omega M)^2}{R_p^2} \quad (7)$$

Here the coupled-in reactance increases directly with  $L_p$  and inversely as the square of  $R_p$ ; the negative sign indicates that this reactance cancels—in part—some of the secondary reactance. An important point to note is that both the resistance  $R_{e_q} = (\omega M)^2/R_p$  and the negative inductance  $L_{e_q} = -(\omega M)^2 L_p/R_p^2$  are not constants, but functions of the frequency owing to the factor  $(\omega M)^2$ . Thus they do not behave quite like an ordinary resistance or inductance, which are independent of frequency.

If these simplified values are inserted in Eq. (4), there results

$$e_{g2} \approx \mu e_{g1} \left( \frac{\omega M}{R_p} \right) \frac{1/j\omega C_s}{\left[ \frac{(\omega M)^2}{R_p} + R_s \right] + j\omega \left[ L_s - \frac{(\omega M)^2 L_p}{R_p^2} \right] + \left[ 1/j\omega C_s \right]} \quad (8)$$

Note that  $\mu/R_p = G_m$ , the transconductance of the tube, so that this may be inserted in Eq. (8) to yield:

$$e_{g2} \approx e_{g1} \frac{\omega M G_m (1/j\omega C_s)}{\left[ \frac{(\omega M)^2}{R_p} + R_s \right] + j\omega \left[ L_s - \frac{(\omega M)^2 L_p}{R_p^2} \right] + \left[ 1/j\omega C_s \right]} \quad (9)$$

This is a general equation that holds for triode as well as pentode tubes, and involves but one approxi-

mation, namely—that  $R_p$  is much greater than (at least three or four times)  $\omega L_p$ . In the case of a pentode, which is today used practically universally in broadcast receiver practice,  $R_p$  is so large that the coupled-in primary inductance  $(\omega M)^2 L_p/R_p^2$  and the coupled-in primary resistance are practically negligible in comparison with  $L_s$  and  $R_s$ , respectively, so that Eq. (9) can be further reduced (after also dividing through by  $e_{g1}$  to obtain the gain or amplification) to:

$$e_{g2}/e_{g1} = \alpha \approx \frac{\omega M G_m (1/j\omega C_s)}{R_s + j\omega L_s + 1/j\omega C_s} \quad (10)$$

The steps required in simplifying an expression down to a workable form depend upon the physical conditions in the problem, and a proper recognition of such factors enables the engineer to determine when *not* to use as well as when to use a particular formula. Hence, such important steps in the derivation of the gain formula are presented here.

Eq. (10) can be simplified in form to a very compact expression by making additional substitutions.

Thus, the resistance and reactance terms can be simplified by the use of the concept of the  $Q$  of a circuit. It will be recalled that the

$Q$  of a circuit is the ratio of the inductive (or capacitive) reactance to the resistance in the

circuit; i.e.,

$$Q = \frac{X_L}{R} = \frac{X_C}{R}$$

From Fig. 18(C) it is clear that the Q of the circuit is the ratio of the reactance of  $L_s$  (ignoring  $L_{e_q}$ ) to the resistance  $R_s$  (ignoring  $R_{e_q}$ ). Since the reactance of  $L_s$  varies with frequency, Q will vary in similar manner. The circuit is in resonance at a frequency  $f_r = \omega_r/2\pi$  such that

$$\omega_r^2 = \frac{1}{L_s C_s} \quad (11)$$

At this frequency Q will be

$$Q_r = \frac{\omega_r L_s}{R_s} \quad (12)$$

Another factor that appears in the simplifying of Eq. (10) is  $\omega^2 L_s C_s$ . From Eq. (11) it is seen that

$$\omega^2 L_s C_s = \omega^2 / \omega_r^2 \quad (13)$$

Let  $\gamma = \omega/\omega_r$ , so that  $\gamma^2 = \omega^2/\omega_r^2$ . Thus, the single symbol  $\gamma$  can be substituted for the double symbol  $\omega/\omega_r$ . Finally, if desired, M may be expressed in terms of  $L_s$  and  $L_p$  and the coefficient of coupling k as follows:

$$M = k \sqrt{L_p L_s} \quad (14)$$

If all the above substitutions are made and a certain amount of algebraic manipulation performed, the following expression is obtained:

$$\alpha = k G_m \sqrt{L_p/C_s} \frac{\gamma}{\sqrt{(1-\gamma^2)^2 + \gamma^2/Q_r^2}} \quad (15)$$

At resonance,  $\gamma = 1$ , and the gain is equal to

$$\alpha_r = k G_m Q_r \sqrt{L_p/C_s} \quad (16)$$

(This can also be rearranged to give  $\alpha_r = G_m \alpha M Q_r$ )

This is the fundamental gain formula for use in amplifier calculations. Off resonance the gain varies in a manner indicated by Eq. (15) and illustrated in Fig. 19. An important relationship exists between the Q of the circuit and the frequencies (or frequency-ratios) at which the gain drops to .707 of its value  $\alpha_r$  at resonance. Thus, the fractional departure below resonance is

$$\gamma_a = 1 - \frac{1}{2Q_r} \quad (17a)$$

and the fractional departure above resonance is

$$\gamma_b = 1 + \frac{1}{2Q_r} \quad (17b)$$

These two points are shown in Fig. 19. Eqs. (17a) and (17b) can

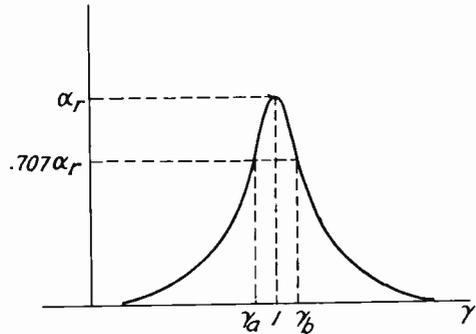


Fig. 19.—Resonance curve representing gain of a tuned r.f. amplifier.

readily be translated into frequency and band widths. Thus, in Fig. 20 is presented the same information as in Fig. 19, but on a frequency rather than a frequency-ratio basis. The circuit is tuned to resonance at the signal frequency  $f_s$ . For frequency deviations  $\Delta f$  to either side of  $f_s$ , the response falls to .707 of its value  $\alpha_r$

*Gain Form*

at resonance. The relationship in terms of the circuit Q is

$$\frac{\Delta f}{f_s} = \frac{1}{2Q} \quad (18)$$

$\Delta f$  is generally, and somewhat arbitrarily, defined as the half-band width. It is the band over which the response is reasonably flat; that is, the range of frequencies

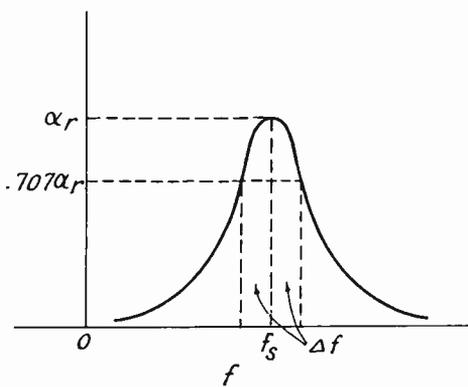


Fig. 20.—Resonance curve of Fig. 19 replotted on a frequency basis.

over which the voltage output ratio does not drop to less than .707 of its mid-band value, nor the power drop to  $(.707)^2 = 1/2$  of its mid-band value. One may assume that if the carrier (of frequency  $f_s$ ) and its side bands fall within the range  $2\Delta f$ , reasonably flat transmission will be had.

As an example, suppose the band width desired is 10 kc =  $2\Delta f$ , and the carrier frequency is 1,000 kc =  $f_s$ . Then, by Eq. (18), the required circuit Q will be

$$Q = f_s / 2\Delta f \quad (19)$$

or

$$Q = 1,000 / 10 = 100$$

It is possible to obtain such a value with a high Q coil and a pentode tube.

**IMAGE REJECTION.**—Although an r.f. stage should provide gain and reasonable uniformity of transmission, its main function in the broadcast range is to afford image rejection. This can be calculated on the basis of the Q (or band width) and the intermediate frequency chosen. An analysis yields the following formula:

$$\begin{aligned} & \frac{\text{Response at image frequency}}{\text{Response at signal frequency}} \\ &= \left( \frac{\Delta f}{2f_1} \right) \left( \frac{f_s + 2f_1}{f_s + f_1} \right) \\ &= \left( \frac{f_s}{4Qf_1} \right) \left( \frac{f_s + 2f_1}{f_s + f_1} \right) \quad (20) \end{aligned}$$

It will be of value to work through an example showing the use of this formula and related ones in calculating the design of an r.f. stage. First find the Q required to give the necessary band width at the lowest frequency in the band, say 540 kc. Assume a band width of 10 kc is desired. Then, by Eq. (19)

$$Q = 540 / 10 = 54$$

This is a rather low Q, hence assume 8 kc is adequate, since transmission takes place beyond 4 kc on each side in the case of an ordinary resonant circuit.

$$Q = 540 / 8 = 67.5$$

The image rejection will be, by Eq. (20), and assuming  $f_1$  (intermediate frequency) = 450 kc:

$\frac{\text{Response at image frequency}}{\text{Response at signal frequency}}$

$$= \frac{540}{4 \times 67.5 \times 450} \left( \frac{540 + 2 \times 450}{540 + 450} \right)$$

$$= 0.00647$$

On a db basis this is:

$$20 \log (0.00647) = -20 \log \frac{1}{0.00647}$$

$$= -20 \log 154.9$$

$$= -20 (2.1901)$$

$$= -43.8 \text{ db}$$

where the significance of the minus sign is that it represents attenuation.

Now calculate the band width and image rejection at the top end of the band, namely—1,600 kc. The variation in Q follows a rather complicated law. As will be discussed farther on, the Q varies over the band for a normal coil so as to have a broad maximum somewhere near the center of the band. As a first approximation, it may be assumed to have the same value at both ends of the band, namely 67.5.

The half-band width, by Eq. (19), will be:

$$\Delta f = (1,600)/(2 \times 67.5) = 11.85 \text{ kc}$$

This half-band width is more than adequate, and shows how an r.f. stage tunes much broader at the higher than at the lower frequencies. In the case of a superheterodyne receiver, however, this is of no concern, since the channel band width is fixed by the i.f. amplifier, providing the band width of the r.f. amplifier is greater in value.

The image rejection, by Eq. (20), is now:

$\frac{\text{Response at image frequency}}{\text{Response at signal frequency}}$

$$= \frac{1600}{4 \times 67.5 \times 450} \left( \frac{1600 + 2 \times 450}{1600 + 450} \right)$$

$$= 0.01606$$

In db this is

$$-20 \log \left( \frac{1}{.01606} \right) = -20 \log 62.2$$

$$= (-20) (1.7938)$$

$$= -35.9 \text{ db}$$

Note that although this is still a fairly high value, it is 7.9 db less than the attenuation at 640 kc, and indicates that the image rejection decreases as the frequency increases. Since the antenna coupling circuit to the grid of the tube is also a tuned circuit, additional image rejection can be expected at this point.

The antenna primary coil is ordinarily weakly coupled to the tuned secondary feeding the grid, in order that different antennas will not couple appreciable reactance (and resistance) into the tuned circuit and thus upset its tracking with the other circuits. The antenna secondary coil itself can be treated just as the plate secondary, and the band width or Q, and the image rejection calculated. It is thus evident that by the use of two such tuned circuits, image rejection on the order of 70 db can be obtained.

For the value of Q just calcu-

3)  $75 \angle 60^\circ + 15 \angle -35^\circ$   
 $= \frac{75}{15} \cdot \frac{\angle 60^\circ - (-35^\circ)}{15} = 5 \angle 95^\circ$

$148 \angle 128^\circ + 22 \angle 316^\circ =$   
 $\frac{148}{22} \frac{\angle 128^\circ - 516^\circ}{128} = 6.73 \angle -188^\circ$

$81 \frac{729}{729} \angle -45^\circ$

$$X_C = 2\pi fL = 6.28 \times 9 \times 10^5 \times 80 \times 10^{-6}$$

$$= 6.28 \times 9 \times 8 = 452.$$

$$X_C = \frac{1}{2\pi fC} = \frac{10^6}{6.28 \times .0015 \times 10^5} = \frac{10000}{.00942}$$

$$= 1060$$

parallel.  $Z_1 = 25 + j452$

$$Z_2 = 10 - j1060$$

$$Z = \frac{(25 + j452)(10 - j1060)}{35 - j608}$$

---


$$= 300 \text{ W.}$$

~~$$60 + 240 =$$~~

$$20 \times 3 + (40 \times 6)$$

$$= 880 \cos 70^\circ$$

$$= 880 \times .342 = 302$$

$$= 220 \times 4 \cos(170 - 100)$$

$$P = E I \cos \theta$$

lated, the gain can be computed by Eq. (16) if additional values are specified. Thus, assume a 6SG7 tube with a  $G_m = 4,100 \mu\text{mhos}$ ,  $L_p = 480 \mu\text{h}$ ,  $L_s = 225 \mu\text{h}$ , and  $k = 6.5$  per cent. In order to tune to 1,600 kc, the tuning capacitor  $C_s$  will have to have a value of

$$\begin{aligned} C_s &= \frac{1}{\omega^2 L_s} \\ &= \frac{1}{(2\pi \times 1.6 \times 10^6)^2 (225 \times 10^{-6})} \\ &= 44 \mu\mu\text{f} \end{aligned}$$

Then, from Eq. (16)

$$\begin{aligned} \alpha_r &= (.065) (.0041) (67.5) \sqrt{\frac{480 \times 10^{-6}}{44 \times 10^{-12}}} \\ &= 59.2 \end{aligned}$$

A greater gain can be obtained by increasing the coefficient of coupling,  $k$ , but gain is not necessary in the broadcast band; the main functions of the r.f. stage are to increase the image rejection and to prevent reradiation. However, 59.2 is a sizeable amount of gain.

At 540 kc, the value of  $C_s$  is equal to:

$$\begin{aligned} C_s &= \frac{1}{(2\pi \times .54 \times 10^6)^2 (225 \times 10^{-6})} \\ &= 386 \mu\mu\text{f} \end{aligned}$$

Then

$$\begin{aligned} \alpha_r &= (.065) (.0041) (67.5) \sqrt{\frac{480 \times 10^{-6}}{386 \times 10^{-12}}} \\ &= 20.1 \end{aligned}$$

This is roughly one-third of the gain at 1,600 kc, and exhibits rather vividly the large variation

in gain with frequency in the case of an r.f. amplifier. This variation can be compensated for by an a.v.c. circuit for signals of normal level, but if a t.r.f. receiver were used, compensation for the usual three r.f. stages employed would be quite difficult, and recourse would be had to special coupling circuits. These computations strikingly illustrate the superior performance of a superheterodyne receiver, both as to constancy of overall gain, and constancy of adjacent channel selectivity.

**TUNING CAPACITOR RANGE.**—A receiver designed for broadcast use will necessarily place the major emphasis on the broadcast spectrum. This band will be taken to cover the frequencies from 540 to 1,600 kc/s. The capacity required to tune this band will be a function of the frequency ratio and the minimum fixed capacity of the circuit. The frequency ratio is 1,600/540 or 2.96 to 1. If one allows some overlap on each end then the band may be designed to cover from 530 to 1,750 kc/s. Other overlaps may be taken and the virtue of a somewhat greater overlap at the high-frequency end will be seen. The frequency ratio becomes 1,750/530 or 3.3 to 1. With a fixed inductance, dictated by the minimum circuit capacity, the minimum capacity will be determined. This minimum capacity will be made up of the distributed capacity of the coil, the stray circuit capacity (due to wiring, socket, etc.), the tube input capacity and the minimum value of the tuning capacitor capacity. The stray capacity must be estimated and will probably be in the vicinity of from 12 to 15  $\mu\mu\text{f}$  with careful arrangement of parts, the tube input capacity will approxi-

mate 5  $\mu\text{f}$  and will be somewhat higher for mixer tubes, the distributed capacity of the average broadcast coil will run another 5  $\mu\text{f}$ , while the minimum of the average condenser gang will approximate 15  $\mu\text{f}$ . The total minimum capacity (which will run about the same on all bands) will be the sum of the shunting capacities as shown in Fig. 21. This sum is the fixed minimum circuit capacity. It is given by

$$C_{\text{min}} = C_d + C_s + C_i + C_{t_m} \quad (21)$$

where

- $C_{\text{min}}$  = total fixed circuit minimum capacity.
- $C_d$  = distributed capacity of the coil.
- $C_s$  = stray circuit capacity.
- $C_i$  = input capacity of the tube.
- $C_{t_m}$  = minimum capacity of the tuning capacitor.

For the case stated above, the following values will be assigned:

$$C_{\text{min}} = 5 + 15 + 5 + 15 = 40 \mu\text{f} \quad (22)$$

Since the maximum frequency for the broadcast band (usually denoted as band B), is 1,750 kc/s and the minimum capacity in the circuit is 40  $\mu\text{f}$ , the inductance required is given by the expression

$$L = \frac{25300}{f^2 \times C} \quad (23)$$

where

- $L$  = the inductance in microhenries.
  - $f$  = the frequency in megacycles.
  - $C$  = the capacity in micromicrofarads.
- For the values of frequency and capacity in the information above,  $L$  becomes

$$L = \frac{25300}{1.75^2 \times 40} = 206 \text{ microhenries} \quad (24)$$

This fixes the value of the tuning inductances for the mixer input and r.f. input stages (if r.f. stages are used). It may be necessary in the experimental model to make slight changes in the coil inductance but the changes will not be considerable. It must be pointed out at this point that this value of 206  $\mu\text{h}$  is the value of the coil in the shield, due allowance having been taken for this fact in the design of the coil. This will be explained farther on.

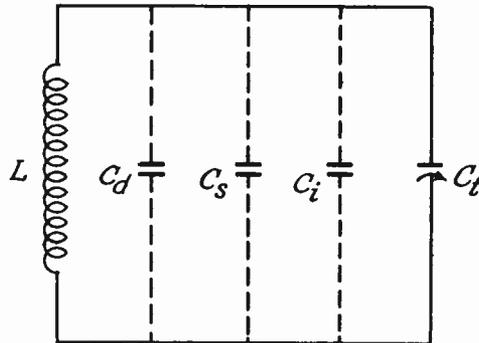


Fig. 21.—Circuit capacities associated with a tuned r.f. stage.

The maximum value of total required capacity may be determined as follows:

$$f_{1750} = \frac{1}{2\pi \sqrt{LC_{\text{min}}}} \quad (25)$$

$$f_{530} = \frac{1}{2\pi \sqrt{LC_{\text{max}}}} \quad (26)$$

In the above expressions, everything is known except  $C_{\text{max}}$ . Remember that the denominator of both right-side expressions contain the two constants,  $2\pi$  and  $L$ , since  $L$  is a fixed inductance in a given case, where the tuning is by varying the capacity. This means that either denominator may be written

L with the Max C

as  $2\pi\sqrt{L} \times \sqrt{C}$ . Divide Eq. 25 by Eq. 26 and obtain

$$\frac{f_{1750}}{f_{530}} = \frac{\sqrt{C_{\max}}}{\sqrt{C_{\min}}} \quad (27)$$

Square both sides and solve for  $C_{\max}$  and obtain

$$C_{\max} = C_{\min} \left[ \frac{f_{1750}}{f_{530}} \right]^2 \quad (28)$$

Substitute the values of  $C_{\min} = 40$ ,  $f_{1750} = 1,750$  and  $f_{530} = 530$  and obtain

$$\begin{aligned} C_{\max} &= 40 \times [1750/530]^2 \\ &= 436 \mu\text{f} \end{aligned} \quad (29)$$

This is the maximum capacity required to tune 206  $\mu\text{h}$  to 530 kc/s but not all of it is in the tuning condenser. Of the total minimum, 25  $\mu\text{f}$  belongs to the circuit, and this 25  $\mu\text{f}$  is still effective. Therefore the tuning capacitor maximum capacity will be  $436 - 25 = 411 \mu\text{f}$ . Such a capacitor is readily obtainable to cover this range. The expression (28) can be generalized to cover any case and may be written

$$\frac{C_{\max}}{C_{\min}} = \left[ \frac{f_{\max}}{f_{\min}} \right]^2 \quad (30)$$

When any three of the four variables are known the fourth can be solved for in terms of the other three. Thus, if the frequency ratio and the minimum capacity for any range are known, the inductance and the maximum capacity may readily be determined as shown.

In the average all-wave receiver, the same tuning capacitor assembly is used for all bands. This dictates the range of the other bands in terms of the maximum and minimum capacities. Thus, the next higher band will start at 1,750 kc/s

and will necessarily have the same frequency ratio, namely 3.3 to 1. So this band will run from 1,750 kc/s to  $1,750 \times 3.3$  or 5,780 kc/s. The third band will start at 5,780 kc/s and run from 5,780 kc/s to  $5,780 \times 3.3$  or 19,200 kc/s. Similarly, the inductances for these bands will be in the same ratio as the ratio of the square of the frequency ratio. That is, the inductance for band C will be  $206/(3.3)^2$  or 19  $\mu\text{h}$ , and the inductance for band D will be  $19/(3.3)^2$  or 1.72  $\mu\text{h}$ . These values of inductance will be used in the r.f. and mixer input tuned circuits exactly as for the broadcast band and are subject to the same design considerations. Due to unpredictable variations in receiver layout which may affect the distributed capacity and other strays, there will be some variation, but at least the designer knows approximately where to start and thus is not groping entirely in the dark.

**COIL DESIGN.**—Coils used in standard broadcast receivers are usually either of the single-layer solenoid form or of the short solenoid (pancake) type. The latter are usually universal wound, which involves a type of winding in which successive turns in series are not directly adjacent to one another, thereby reducing the distributed capacitance.

In Fig. 22 is shown a type of transformer used in an r.f. amplifier stage. The secondary may have an inductance of 225  $\mu\text{h}$  and a Q of 90; the primary may have an inductance of 450 to 500  $\mu\text{h}$  and a Q as low as 15. The greater inductance of the primary enables it to present an appreciable load to the pentode tube feeding it, and its Q is not important because the

plate current flowing through it is determined primarily by the pentode tube, whose high plate resistance

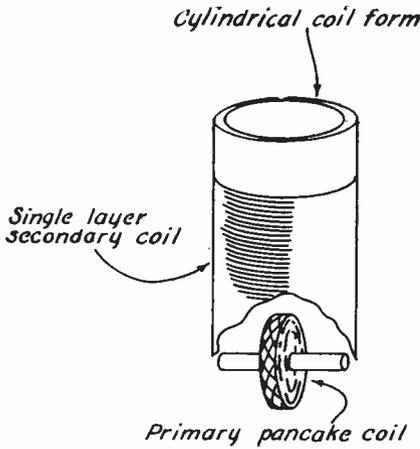


Fig. 22.—R.F. transformer.

causes it to act as a constant current generator.

The secondary inductance enables it to operate satisfactorily with the usual variable tuning capacitors available, and its high Q assures suitable image rejection, as discussed previously. The coefficient of coupling is low—usually not over 10 per cent—and is limited primarily by the amplifier gain that it is safe to develop.

Inductance Formula.—Although the inductance of each coil can be calculated fairly accurately by means of elaborate formulas, relatively simple formulas are available that are accurate to within one per cent.\* For a single-layer solenoid, this is

$$L = \frac{r^2 n^2}{9r + 10l} \text{ microhenries} \quad (31)$$

where r is the radius of the coil and l is its length in inches. It

\*H.A. Wheeler, "Simple Inductance Formulas for Radio Coils," Proc. I.R.E., Oct. 1928.

is assumed that l is greater than 0.8r.

As an example, suppose a coil of 113 turns of #32 wire is wound on a one-inch cylindrical form. What is its inductance? No. 32 wire has a diameter of 0.00795 inches. Assume that the diameter is 0.009" including the insulation. The length of the coil will therefore be  $113 \times 0.009 \approx 1.0$  inches = l. By Eq. (31)

$$L = \frac{(0.5)^2 (113)^2}{9 \times .5 + 10 \times 1.0} = 220 \mu\text{h}$$

For a short multiple-layer solenoid,

$$L = \frac{0.8r^2 n^2}{6r + 9l + 10h} \quad (32)$$

*Multi-Layer*

where the letters indicate the dimensions in inches, shown in Fig. 23. As an example, suppose a coil of this type is wound with

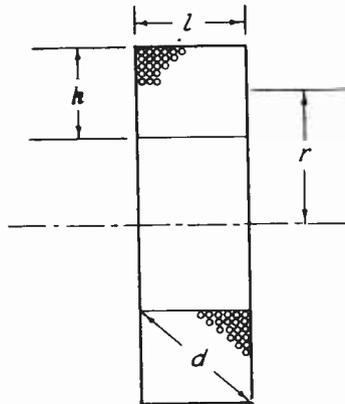


Fig. 23.—Short multiple-layer solenoid, showing significant dimensions.

#22 wire with .002" insulation. The coil is wound on a 1/4-inch wood dowel, and has 12 turns per layer, and 17 layers. No. 22 wire has a diameter of .025". Add  $2 \times .002$ "

for insulation to get an overall diameter of 0.029". For 12 turns, this gives a length of

$$l = 12 \times .029 = .348" \cong 0.35 \text{ inches}$$

The height will be

$$h = .029 \times 17 = 0.493"$$

Then

$$r = \frac{h}{2} + \frac{1/4"}{2} = .372"$$

There are  $12 \times 17 = 204$  turns =  $n$ .  
Then, from Eq. (32),

$$L = \frac{0.8 (.372)^2 (204)^2}{(6 \times .372) + (9 \times .35) + (10 \times .493)}$$

$$= 444 \mu\text{h}$$

These two coils represent a possible design for the r.f. coil shown in Fig. 22. There are formulas for the mutual inductance between two coils, but these are exceedingly involved. It is in general simpler to adjust the coupling experimentally to the desired value.

Coil Losses and Q.—The most difficult factor to determine is the equivalent series resistance of a coil at the operating frequency. This resistance is greater than the d.c. resistance because of modifications in the current distribution through the cross section of the wire owing to skin effect, proximity of neighboring conductors and shield cans, dielectric losses in the coil form, and core losses, if an iron dust core is employed.

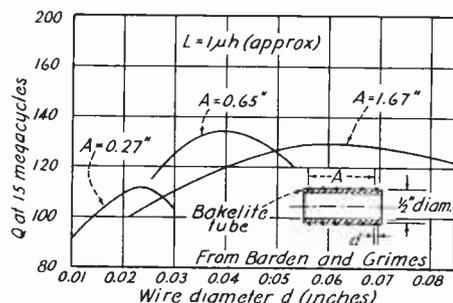
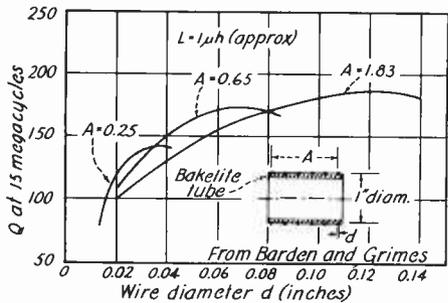
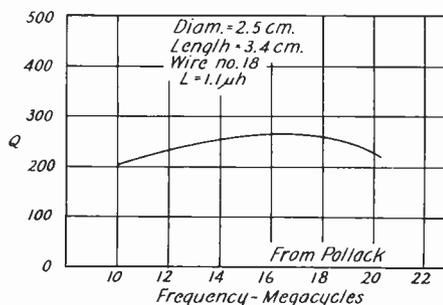
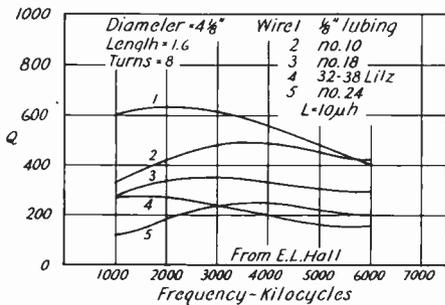
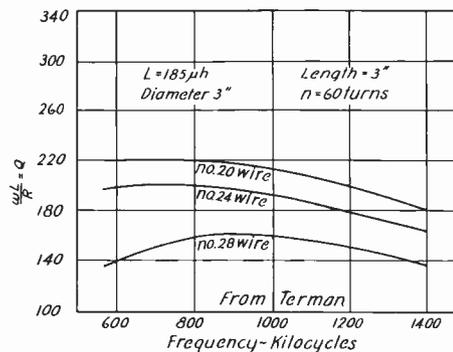
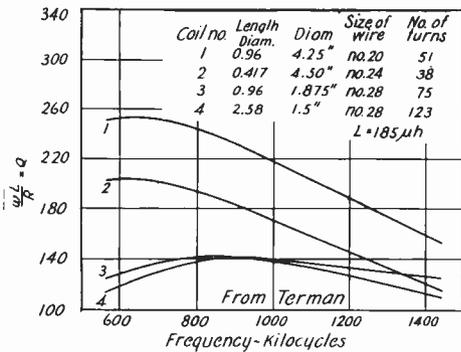
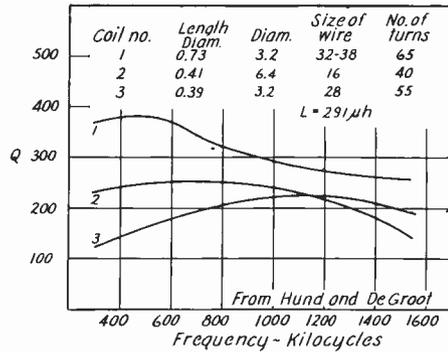
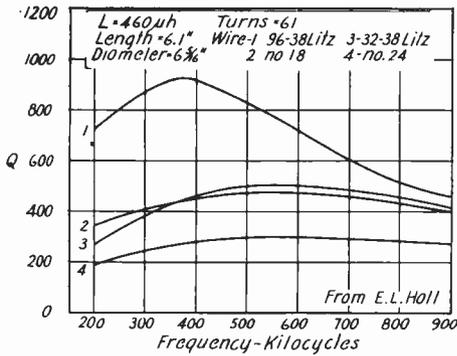
Owing to skin effect, the losses, and hence the resistance should increase as the square root of the frequency. Since the inductive reactance increases directly

with the frequency, their ratio, or the Q of the coil, should increase as  $f/\sqrt{f} = \sqrt{f}$ . Actually, owing to the other factors mentioned, the Q of an average coil increases with frequency up to a certain point and then starts to decrease—once more, so that in general a broad maximum is observed in the operating range.

Variations in Q of from 50 to 100 per cent can be expected from the minimum to the maximum value. Dielectric losses are in general small in the standard broadcast band, unless unimpregnated insulation and coil forms, such as paper are employed, since these absorb moisture. The fact that the high-frequency resistance increases more rapidly for large conductors than for small ones makes the size of wire used relatively unimportant, although the larger conductor in general has the higher Q. Also, larger coils have higher Q's, but since a Q much greater than 100 is in general not required, fairly small coils today have supplanted the earlier large and bulky coils used in the original t.r.f. receivers.

At intermediate and even at broadcast frequencies, Litz (Litzendraht) wire affords lower losses than solid wires. Litz wire consists of several strands of fine insulated wire, twisted so that all strands occupy in practically equal lengths all parts of the cross section of the built-up conductor. Thus, 3 #40 wires may be used. Since no one strand occupies exclusively the center of the conductor, the current is forced to flow equally in all strands, whereas in a solid wire of the same effective cross section, skin effect

RECEIVERS, RADIO FREQUENCY AMPLIFICATION



(Courtesy Radio Engineers' Handbook, by F.E. Terman.)

Fig. 24.—Examples of Q-variation with frequency and with size of conductor.

forces the current to the surface, so that the interior of the conductor carries practically no current and hence does not contribute to its conductance.

However, at higher frequencies Litz wire fails to show any advantage over solid wire, and even at broadcast and intermediate frequencies its advantage is not so great as to warrant its use in all cases. Probably more r.f. coils are wound with solid wire than with Litz wire owing to lower cost and ease in soldering.

Formulas are available to calculate the ratio of a.c. to d.c. resistance of a coil as a function of its dimensions and the frequency.\* These, however, are not too accurate, and serve mainly to compare coils of various shapes, rather than to give the losses of an actual coil. They are mainly of interest to the coil designer. In Fig. 24 are given curves based on actual measurements of coils, and serve to give the reader some idea of the variation of the Q with frequency and with wire size.

Shielding.—In order to prevent magnetic or electrostatic coupling between stages, the parts must be shielded; i.e., enclosed in metallic cans. This is particularly true of coils. All electric field lines emanating from the turns of a coil that are "hot" to ground will terminate on the shield can and fail to penetrate it. Thus there can be no capacitive coupling to other coils or circuit components.

Any alternating magnetic flux from the coil, in attempting to pass through the shield, will induce eddy currents in it just like in any short-circuited secondary of a

transformer. These eddy currents set up a counter magneto-motive force that repels the flux back into the can. Thus few lines can penetrate the shield, particularly if the eddy currents can be set up with ease; i.e., if the shield has a high conductivity, such as copper, or even aluminum.

The action of a shorted secondary has been shown previously to produce the effect of a negative inductance in the primary, so that in the case of a coil, the shield tends to reduce its self inductance. This is the major effect, although the losses in the shield, even though small, do tend to lower the Q of the coil.

In Fig. 25 is given a family of curves, taken from the R.C.A. Application Note #48, which enables the effect of the shield to be calculated.

Note that the scales are logarithmic. The meanings of the symbols are clearly shown in the lower right-hand corner. The symbol  $k$  is the reduction factor and may also be considered as the percentage reduction if  $k$  is multiplied by 100. For example, if the inductance of a coil is 100  $\mu\text{h}$  and a certain shield indicates that  $k$  is .25, the inductance of the coil in this shield will be 25 per cent less than the inductance of the coil without the shield, i.e., 100 - 25 or 75  $\mu\text{h}$ . This relationship may be expressed mathematically. Let  $L_a$  = the inductance of the coil in air, let  $L_s$  = the inductance of the coil in the shield, and  $k$  = the reduction factor. Then  $L_s = L_a (1 - k)$ . For the example just cited, by substitution of values

$$L_s = 100(1 - .25) = 75 \mu\text{h}$$

\*See, for example, Terman's "Radio Engineers' Handbook."

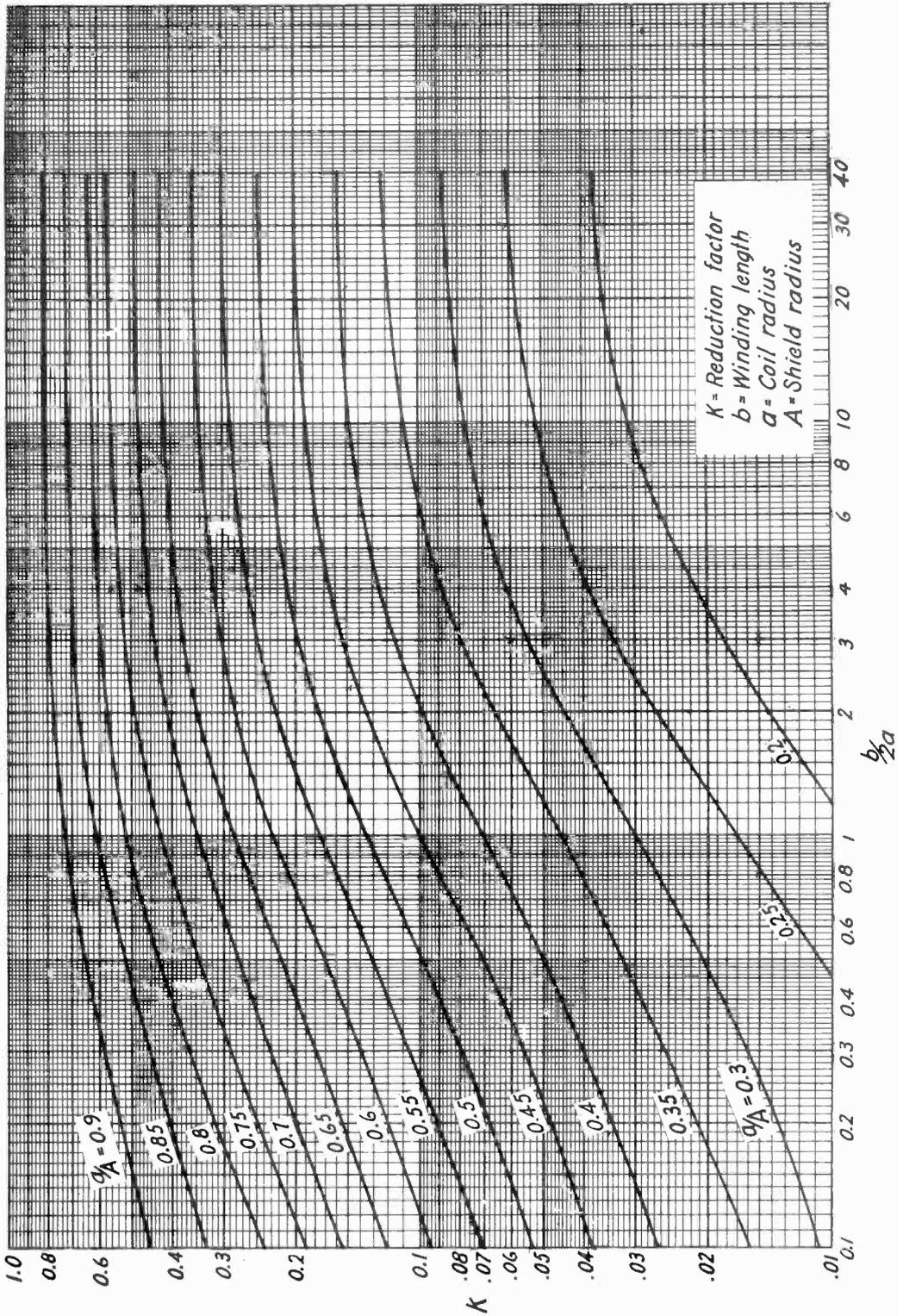


Fig. 25.---Curves for determination of decrease in inductance produced by a coil shield.

Before citing examples illustrating the uses of the chart, there are certain restrictions for its use. If the shield is round, the symbols are used without modification. If the coil shield is square, the value of the shield radius  $A$  is taken as  $.6S$  where  $S$  is the dimension of the side of the square shield. It is further understood that the coil shield must extend above the coil winding by a distance at least equal to the coil radius. The relations between the coil and shield are shown in cross section in Fig. 26. The coil should also be above the chassis the same distance as the shield is spaced above the coil, unless a large hole is used under the coil in the chassis.

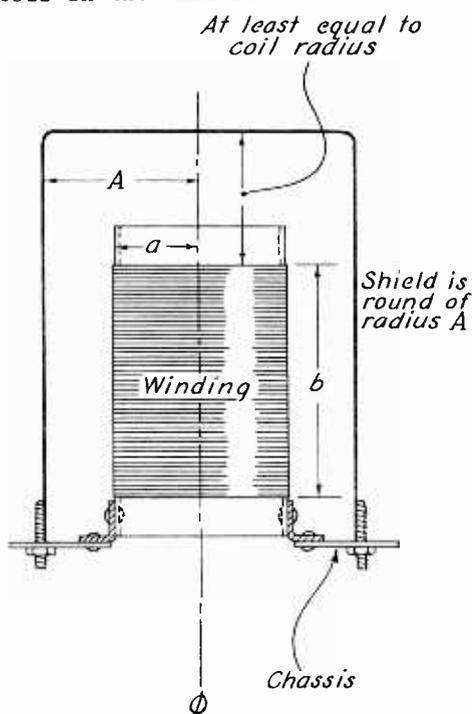


Fig. 26.—Pertinent coil and shield dimensions.

Assume that a certain coil has a winding length  $b$  of 2 inches, a radius  $a$  of  $.5$  inches, a shield

radius  $A$  of 1 inch and an inductance in air of  $280 \mu\text{h}$ . Substituting the values in the two expressions, there obtains

$$b/2a = \frac{2}{2 \times .5} = 2$$

and

$$a/A = \frac{.5}{1} = .5$$

Locating the abscissa value  $b/2a$  of 2 and the curve labeled  $a/A$  of  $.5$ , one notes the intersection of these curves and from this intersection determines the value of the ordinate for  $k$ . For the values cited,  $k = .175$ . This shield will accordingly reduce the inductance by the factor  $.175$  or by 17.5 per cent. Substituting the values in the previous expression for  $L_s$ ,

$$L_s = 280(1 - .175) = 231 \mu\text{h}$$

This example illustrates the fact that a coil must be designed for the proper inductance in the shield since shielding a coil with ordinary shields always results in a reduction in the effective inductance.

Suppose that calculations, such as were discussed in another assignment, show that a coil should have an inductance of  $250 \mu\text{h}$ , effective inductance, in a certain receiver and that this coil will be used in a shield. Assume further that this coil is to have the dimensions of that of the previous example, similarly for the coil shield. The problem now is to determine the value of the inductance for the coil without the shield since the formulas and charts for coil design do not take into account the shielding effect but rather the value of the inductance in air.

By exactly the same procedure the value of  $k$  is found to be .175.  $L_a$  is known to be 250  $\mu$ h. Solve the expression for  $L_a$ , from  $L_a = L_a (1 - .175)$ .

$$250 = L_a (1 - .175)$$

or

$$L_a = \frac{250}{(1 - .175)} \text{ or } 303 \mu\text{h}$$

A third example is that of determining the shield dimension for a stated coil but with the restriction that the effect of the shield shall not exceed a certain arbitrary value. Suppose that for the coil dimensions as stated, the effect of the shield shall not reduce the value of  $L_a$  by more than 10 per cent. For this problem,  $k$  becomes in the limit .10 and  $b/2a = 2$  as before. From the curves locate the intersection of  $b/2a$  and  $k$ , and note that it is also crossed by the curve  $a/A$  of .4. Since  $a$  is known and is .5,  $A$  may be determined from the relation  $a/A = .4$ . Substitute and obtain  $A = 1.25$  inches. This is the least shield radius that can be used. Values of  $A$  greater than this will result in a smaller value of  $k$  and will represent correspondingly less effect on the reduction of inductance.

If the values do not lie directly on curves drawn in Fig. 25 one must resort to interpolation, as is usual in the use of curves. It may also be mentioned that these curves apply only to single-layer coils. Study of the curves reveals that for a given shield and coil radius the use of various values of  $b/2a$ , and  $k$  increases up to a certain point fairly rapidly but after a certain value of  $b/2a$  is reached, the effect of shield increases very slowly. It is also observed that

the smaller the ratio  $a/A$  the less the variation in  $k$  for a given change in  $A$  or  $a$ .

TRACKING.—In the early receivers, the usual three t.r.f. stages were individually tuned; each variable capacitor had a separate dial for the purpose. Even though the capacitors and coils were made as nearly alike as possible, unavoidable small variations in these components, as well as variations in tube and stray wiring capacities precluded mounting the variable capacitors on one shaft so that one tuning dial would suffice.

However, in spite of the sharpness of tuning and consequent accuracy of adjustment required,

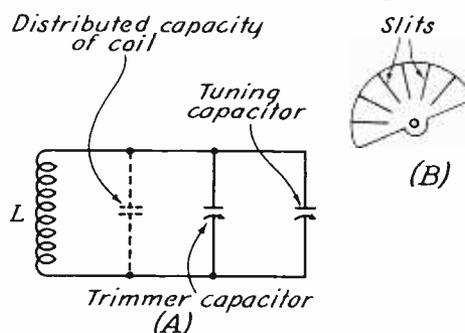


Fig. 27.—Capacities associated with an r.f. coil.

means were found to obtain this desired result. Each variable capacitor had a small adjustable "trimmer" capacitor connected in parallel with it, as shown in Fig. 27(A). At the low end of the band, the plates of the main tuning capacitor are fully meshed and the capacity is a maximum. In this case, the small trimmer has very little effect upon the tuning capacitance.

At the high-frequency end of the band, however, the tuning capacitor is at a minimum, and the

parallel trimmer therefore has maximum effect. It can therefore be used at this end of the band to provide an adjustment to make the tuning of the various stages the same.

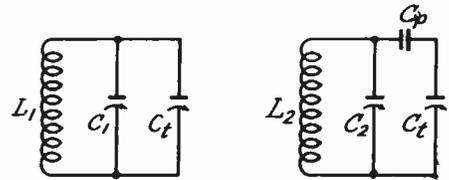
At the low end of the band, any discrepancies between stages can be taken care of, if necessary, by the use of a radially slotted plate on the tuning capacitor, as shown in (B). The individual sectors can then be bent more or less into proximity with the corresponding stator plate, and the capacitance adjusted. This adjustment is clearly not effective at the high end of the band because the sectors are all unmeshed with the stator plates.

The advent of the superheterodyne receiver introduced a somewhat different problem. In this type of receiver, only one—or at most two—tuned r.f. circuits are employed, and they are satisfactorily ganged by the use of trimmer capacitors. But the local oscillator tank circuit must be tuned so as to be always at the *fixed* intermediate frequency above the r.f. circuits. This means that L or C or both must be less for the oscillator tank.

In practice it is desirable to use identical variable capacitor sections for the r.f. and oscillator stages. However, by putting a relatively large "padding" capacitor in series with the variable capacitor, the effective capacity of the oscillator tank is reduced. In addition, the oscillator coil is suitably reduced, and in this way a higher oscillator frequency at any dial setting is obtained.

The circuits are shown in Fig. 28. Here  $C_1$  and  $C_2$  are trimmer

capacitors\*;  $C_t$  is the tuning capacitor; and  $C_p$  is the series padder



R.F. tuned circuit    Oscillator tank circuit

Fig. 28.—Tuned circuit for r.f. amplifier and local oscillator stages.

capacitor. Also  $L_1$  is greater than  $L_2$ . By the proper choice of  $L_2$  relative to  $L_1$ , of  $C_p$ , and of  $C_2$  relative to  $C_1$ , it is possible to make the oscillator circuit tune to a frequency  $f_1$  above that of the tuned circuit (where  $f_1$  is the intermediate frequency) for three settings of the tuning dial. Note that the choice of three factors as given above enables three points of the dial tuning to be satisfied. At other points of the dial the oscillator frequency may be somewhat too high or too low compared to the correct value.

However, the maximum deviation from perfect tracking is not serious if the series and shunt condensers are properly chosen. For the broadcast band covering the range of frequencies outlined above, these points might be taken as 1.550 mc/s, 1.075 mc/s, and .600 mc/s. The points of perfect tracking and the deviations are somewhat as shown in Fig. 29. The same idea will

\*These are normally small mica-compression type variable capacitors with a range from 3-30  $\mu\text{mf}$ .

also be applied to the other bands. However, with the use of the same

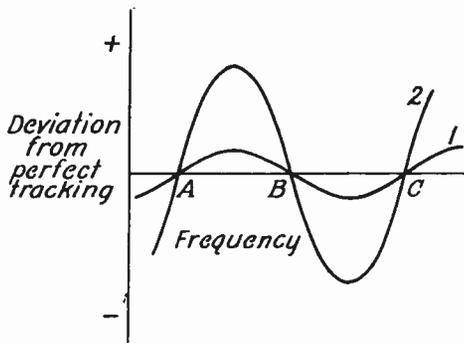


Fig. 29.—Tracking error of local oscillator.

intermediate frequency, the problem of alignment becomes easier as the frequency range increases as will be seen.

If perfect tracking could be obtained, the deviation would be zero at all frequencies over the tuning ranges. With the use of the circuit for oscillator tuning as shown in Fig. 28, there is some deviation but with a properly designed oscillator circuit the amount of deviation is small enough not to interfere with proper operation. Curve 1 shows the deviation where a low intermediate frequency is used, as for example 175 kc/s, while curve 2 shows the deviation with a higher intermediate frequency. Measured values of deviation for an i.f. of 175 kc/s showed a maximum deviation of 2.1 kc/s, for an i.f. of 3.000 mc/s a maximum deviation of 6.5 kc/s. Points A, B and C represent the frequencies where perfect tracking obtains, which for the example cited would be .600, 1.075, and 1.550 mc/s. If the circuits are properly designed, the points of maximum deviation will be approximately mid-way

between the points where perfect tracking is obtained. It is thus seen that the deviation is a function of the i.f. employed; the width of frequency band to be covered and the care with which the circuit is designed.

There are various methods for determining the size of the oscillator components. Since the final answers depend in a sense on the difference between two large and nearly equal quantities, the accuracy obtained is often not too good when an ordinary slide rule is used; however, in actual practice the final adjustments are made experimentally, and hence the calculations that are to follow have sufficient accuracy to indicate the range of the adjustable components—such as the trimmer capacitor—that are to be ordered.

The circuits commonly employed for the r.f. and oscillator stages are shown in Fig. 30 (A) and (B), respectively. Suppose the three

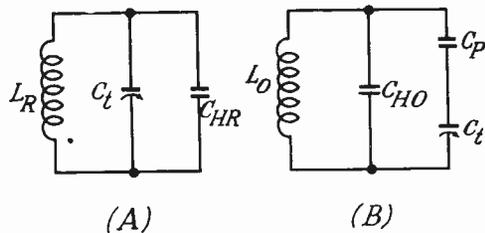


Fig. 30.—Representative tuned circuits for the r.f. and oscillator stages of a receiver.

signal frequencies at which perfect tracking is to take place are  $f_{1r}$ ,  $f_{2r}$ , and  $f_{3r}$ . Representative values are those previously given, namely, 1.550, 1.075, and .600 mc.

If the intermediate frequency

is  $f_1$ , then the corresponding three oscillator frequencies are (assuming the oscillator is above the signal frequency):

$$f_{10} = f_{1r} + f_1$$

$$f_{20} = f_{2r} + f_1$$

$$f_{30} = f_{3r} + f_1$$

For example, if  $f_1 = .455$  mc, then for the values just given for the signal frequencies,  $f_{10} = 1.550 + .455 = 2.005$  mc,  $f_{20} = 1.530$  mc, and  $f_{30} = 1.055$  mc.

The setting of the tuning capacitor  $C_t$  will be assumed to be the same for the r.f. and oscillator stages (although this is not absolutely necessary), and it can be readily found by simple computations from the values first calculated for the r.f. stage. Suppose at signal frequency  $f_{1r}$ ,  $C_t$  has the value  $C_1$ ; at  $f_{2r}$ , the value  $C_2$ ; and at  $f_{3r}$ , the value  $C_3$ . Then

where the frequencies are in megacycles, the capacitors in micro-microfarads, and the inductance is in microhenries.

As an illustration, consider the values of  $L_R$  and  $C_t$  found previously in the computations for the tuning capacitor range. They were  $L_R = 206 \mu h$ , and  $C_t$  (max.) =  $411 \mu \mu f$  at .530 mc, with  $25 \mu \mu f$  additional in stray wiring capacity, etc. Assume that this also includes  $C_{HR}$ , the trimmer capacitance. Hence, at .530 mc, the r.f. stage is resonant when  $L_R = 206 \mu h$ , and the total capacitance is  $411 + 25 = 436 \mu \mu f$ , as can be readily checked by the resonance formula.

At 1.550 mc, the total capacitance required is

$$C_1 = 436 \left( \frac{.530}{1550} \right)^2 = 51.1 \mu \mu f$$

of which  $51.1 - 25 = 26.1 \mu \mu f$  is in tuning capacitor.

Similarly, at 1.075 mc,

$$C_p = \frac{\left[ \frac{1 - f_{10}^2/f_{20}^2}{1 - f_{10}^2/f_{30}^2} \right] \left[ \frac{C_3 - C_1}{C_2 - C_1} \right] C_2 - C_3}{1 - \left[ \frac{C_3 - C_1}{C_2 - C_1} \right] \left[ \frac{1 - f_{10}^2/f_{20}^2}{1 - f_{10}^2/f_{30}^2} \right]} \quad (\mu \mu f)$$

$$C_{H0} = \frac{\left[ \frac{C_1 C_p}{C_1 + C_p} \right] \frac{f_{10}^2}{f_{20}^2} - \left[ \frac{C_2 C_p}{C_2 + C_p} \right]}{\left[ 1 - f_{10}^2/f_{20}^2 \right]} \quad (\mu \mu f)$$

$$L_o = \frac{1}{4\pi^2 f_{10}^2 \left[ C_{H0} + \frac{C_1 C_p}{C_1 + C_p} \right]} \quad (\text{Henry})$$

(33)

$C_2$  is found to be

$$436 \left( \frac{.530}{1.075} \right)^2 - 25 = 81.1 \mu\mu\text{f}$$

and at .600 mc,

$$C_3 = 436 \left( \frac{.530}{.600} \right)^2 - 25 = 315 \mu\mu\text{f}$$

It is now possible to calculate  $C_p$ ,  $C_{HO}$ , and  $L_o$  in succession by means of Eq. (33). Thus

$$C_p = \frac{\left( \left[ \frac{1 - (2.005/1.530)^2}{1 - (2.005/1.055)^2} \right] \left[ \frac{315 - 26.1}{81.1 - 26.1} \right] 81.1 \right) - 315}{1 - \left[ \frac{315 - 26.1}{81.1 - 26.1} \right] \left[ \frac{1 - (2.005/1.530)^2}{1 - (2.005/1.055)^2} \right]} = 440 \mu\mu\text{f}$$

$$C_{HO} = \frac{\left( \frac{26.1 \times 440}{26.1 + 440} \right) (1.717) - \left( \frac{81.1 \times 440}{81.1 + 440} \right)}{(1 - 1.717)} = 36.4 \mu\mu\text{f}$$

The value of  $C_{HO}$  is the oscillator trimmer shown in Fig. 30(B) and includes the stray capacity of the circuit. Now to find  $L_o$ ,

$$L_o = \frac{1}{4\pi^2(2.005)^2 \left[ 36.4 + \frac{26.1 \times 440}{26.1 + 440} \right]}$$

$$L_o = 103.6 \mu \text{ henries}$$

This is  $103.6/206 = .502$  or 50 per cent of  $L_R$ , the inductance required in the r.f. stage.

The same procedure can readily be applied to the other bands, such as from 1.750 mc to 5.780 mc. If the same tuning capacitor is employed, then the inductance required in the r.f. stage will be less in

the proportion of  $(.530/1.750)^2 = .0917 = (1.750/5.780)^2$  or  $(206) (.0917) = 18.89 \mu\text{h}$ . Three signal frequencies at which perfect tracking is obtained must then be chosen; these can be, if desired, exactly three times the values chosen for the standard broadcast band, in which case the same capacitor settings,  $C_1$ ,  $C_2$ , and  $C_3$  will apply. Then, assuming the same i.f. amplifier is employed (as

is normally the case), the corresponding three oscillator frequencies can be found.

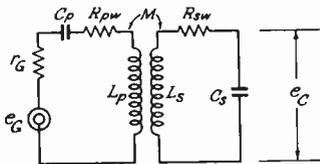
The values, when substituted in Eq. (33), will then give the values of  $C_p$ ,  $C_{HO}$ , and  $L_o$  corresponding to this band. The most striking result will be that the padder capacitor  $C_p$  will come out greater than the value  $440 \mu\mu\text{f}$  required for the standard broadcast band. In short, the greater the ratio of the signal frequency to the intermediate frequency, the larger  $C_p$  comes out to be, which in turn means the less effect it has when in series with  $C_t$ . Hence tracking is most difficult in the standard broadcast band, and for that reason the previous problem has been worked out in detail for this band.

*I.F. AMPLIFICATION.*—As indicated previously, the single res-

onant circuit employed ordinarily in r.f. amplification is too broad in its frequency response for i.f. amplification. The resonance curve does not have a sufficiently broad flat top, and does not drop off sufficiently rapidly beyond the pass band to afford adequate adjacent-channel selectivity.

A response curve that approaches more closely the ideal rectangular shape is obtained by employing a double-tuned circuit. Since it operates at a fixed (intermediate) frequency, no tuning complications are encountered, and the increased circuit complexity is entirely feasible.

*Analysis.*—The analysis is based on the circuit shown in Fig. 31. This is somewhat easier to analyze than the double-tuned circuit actually employed, in which  $C_p$  parallels  $L_p$  (Fig. 32). Moreover, the latter type of circuit can be readily transformed into the one shown in Fig. 31,



RPW and RSW are the coil resistance values of  $L_p$  and  $L_s$ , respectively Fig. 31.—Double-resonant circuit employed in the analysis.

so that no loss in generality occurs in employing the latter circuit. To show this, break the circuit in Fig. 32 at the dotted line, and

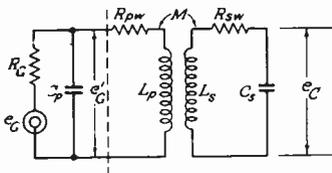


Fig. 32.—Actual double-tuned i.f. circuit.

consider the part to the left, the generator (by Thevenin's theorem), and the part to the right, the load. The impedance of the apparent generator is  $C_p$  and  $R_G$  in parallel, and the apparent generated voltage is the fraction of  $e_G$  that appears across  $C_p$  with the load disconnected. The apparent internal impedance is

$$Z_G = \frac{(R_G)(1/j\omega C_p)}{R_G + 1/j\omega C_p} = \frac{R_G}{1 + j\omega C_p R_G}$$

Multiplying numerator and denominator by the conjugate  $1 - j\omega C_p R_G$ , there is obtained:

$$Z_G = \frac{R_G - j\omega C_p R_G^2}{1 + \omega^2 C_p^2 R_G^2} \quad (34)$$

If  $R_G$  represents the plate resistance of a pentode tube, as is almost universally the case, then

$$R_G \gg 1/\omega C_p$$

or, multiplying both sides of this inequality by  $\omega C_p$ , there is obtained:

$$\omega C_p R_G \gg 1$$

so that

$$\omega^2 C_p^2 R_G^2 \gg 1$$

Hence, the denominator of Eq. (34) can be written simply as  $\omega^2 C_p^2 R_G^2$ , whereupon the equation reduces to

$$Z_G \approx \frac{1}{\omega^2 C_p^2 R_G} - \frac{j}{\omega C_p} \quad (35)$$

This means that the generator impedance of Fig. 32 appears to be a resistance  $1/\omega^2 C_p^2 R_G$  in series with a capacitor of value  $C_p$ , and hence is identical to Fig. 31 except that the generator resistance varies with  $1/\omega^2$ , whereas  $R_G$  is constant. However, over the narrow i.f. range  $1/\omega^2 C_p^2 R_G$  may be regarded as substantially constant.

It will be recalled that in the case of an r. f. amplifier, the coupled-in resistance of the  $R_p$  of the tube was negligibly small and did not appreciably lower the  $Q$  of the secondary coil. In the case of the i. f. transformer, the effect of the  $R_p$  of the tube (here denoted in the general case as  $R_G$ ), is comparable to the primary coil resistance  $R_{pw}$ ; i. e., the term  $1/\omega^2 C_p^2 R_G$  cannot be neglected.

Thus, if  $R_G = 645,000$  ohms, and  $\omega = 2\pi \times 455,000$ , and  $C_p = 75.5 \mu\mu\text{f}$ , then

$$1/\omega^2 C_p^2 R_G = 1 \div [(2\pi \times 455,000)^2 \times (75.5 \times 10^{-12})^2 (645,000)] = 33.3 \Omega$$

This compares with a value of  $R_{pw}$  that is somewhere around 38.6 ohms, for a value of  $L_p = 1.62 \times 10^{-3}$  h, and a coil  $Q = 120$ .

The apparent generated voltage in Fig. 32 is

$$e'_G = \frac{e_g(1/\omega C_p)}{R_G} = \frac{e_g}{\omega C_p} \frac{(1 + j\omega C_p R_G)}{R_G} \\ = \frac{e_g + j\omega C_p R_G e_g}{R_G \omega C_p} \quad (36)$$

If the source is a pentode tube, then the actual generated voltage  $e_g$  is really  $\mu e_g$ , where  $e_g$  is the signal input to the grid of the tube, and  $R_G = R_p$ , the plate resistance of the tube. Hence, in this case Eq.

(36) becomes

$$e'_G = \frac{\mu e_g + j\omega C_p R_p}{R_p \omega C_p} = \frac{G_m e_g}{\omega C_p} + \frac{j\omega C_p R_p}{\omega C_p R_p} \\ \approx \frac{G_m e_g}{\omega C_p} \quad (37)$$

Thus, in analyzing the circuit of Fig. 31, these substitutions must be made when the solution of that circuit is applied to the actual circuit of Fig. 32. An analysis of Fig. 31 is very involved, even when simplifying assumptions are made, because of the large number of circuit elements present. Thus, there are two circuit  $Q$ 's involved: that of the primary =  $Q_p$ , and that of the secondary =  $Q_s$ . In addition, the value of the coefficient of coupling  $k$ , has a profound effect upon the circuit behavior, particularly the band width, and must be determined. About the only important simplifying assumption that can be made is that the two coils are individually tuned to the same i. f. carrier, or

$$L_p C_p = L_s C_s \quad (38)$$

In practice,  $L_s$  and  $L_p$  are usually also identical coils, but the formulas and graphs to be presented here do not require the latter stipulation.

Following Terman, we can write as the basic equation for Fig. 31:

$$\frac{\text{Output voltage across } C_s}{\text{Input generated voltage}} = \frac{e_c}{e_g}$$

$$= \frac{-1}{\gamma^2} \sqrt{\frac{L_s}{L_p}} \frac{k}{\left[ k^2 + \frac{1}{Q_p Q_s} - \left(1 - \frac{1}{\gamma^2}\right)^2 + j \left(1 - \frac{1}{\gamma^2}\right) \left(\frac{1}{Q_p} + \frac{1}{Q_s}\right) \right]} \quad (39)$$

where  $\gamma$  = the ratio of the actual to the resonant frequency.

$e_o$  = the voltage impressed in series with the primary.

$Q_p$  =  $\omega L_p / R_{pt}$  for the primary circuit, where  $R_{pt}$  is the total series resistance of the primary circuit.

$Q_s$  =  $\omega L_s / R_{st}$  for the secondary circuit, where  $R_{st}$  is the total series resistance of the secondary circuit.

$L_p$  = total inductance of the primary circuit.

$L_s$  = total inductance of the secondary circuit.

This equation can be employed to calculate the response, if desired, point by point for various values of  $\gamma$ , the frequency ratio. This is an arduous task, and is unnecessary for points in the pass band, as will be shown. For values of  $\gamma$  outside of the pass band, the reactances so greatly overbalance the resistances that the Q's may be assumed infinite, and  $1/Q_p = 1/Q_s = 0$ , so that Eq. (39) reduces to:

$$\frac{e_o}{e_g} \approx \frac{-1}{\gamma^2} \sqrt{\frac{L_s}{L_p}} \frac{k}{\left[ k^2 - \left( 1 - \frac{1}{\gamma^2} \right)^2 \right]} \quad (40)$$

This equation is of value in calculating the attenuation of the circuit to some particular frequency remote from the pass band.

At resonance (intermediate carrier frequency),  $\gamma = 1$ , and Eq. (39) reduces to

$$\frac{e_c}{e_o} \text{ (at resonance) } = (-1) \sqrt{\frac{L_s}{L_p}} \frac{k}{\left[ k^2 + \frac{1}{Q_p Q_s} \right]} \quad (41)$$

This formula will be useful in calculating the gain after the various circuit constants have been determined from the specifications initially laid down for the stage, as will now be shown.

If the coupling  $k$  is increased from a low value, a critical value  $k_c$  is reached, where

$$k_c = \frac{1}{\sqrt{Q_s Q_p}} \quad (42)$$

and the response curve changes from a single-peak to a double-peak form, provided  $Q_s = Q_p$ . If the two Q's are not equal,  $k$  will have to exceed the critical value  $k_c$  by

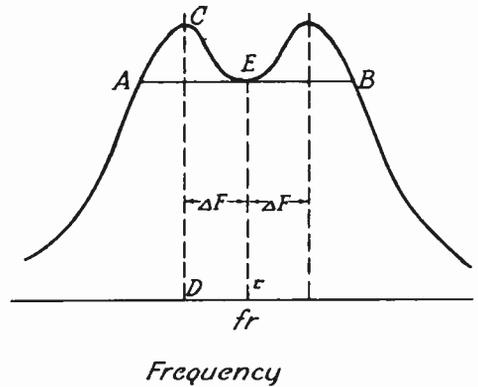


Fig. 33.—Band width relationships for double-peaked resonance curve.

an amount depending upon the ratio of  $Q_s$  to  $Q_p$ , and the two peaks will differ somewhat in height.

In Fig. 33 is shown a typical double-peaked resonance curve. The peaks occur at a frequency  $f_r + \Delta f$  and  $f_r - \Delta f$ , where  $f_r$  is the resonant

frequency to which each coil is individually tuned. The response, however, may be considered to be relatively flat over a greater band width, namely—from A to B.

It has been shown by Aiken\* that the band width AB is very simply expressed in terms of  $\Delta f$ , namely,

$$AB = \sqrt{2} (2\Delta f) \quad (43)$$

Hence, if the overall band width AB is specified, the quantity  $\Delta f$ , which represents the frequency separation of either peak from the resonant frequency, can readily be determined. This in turn permits the other factors in the design to be determined.

The problem to be solved may be one of the following:

1. Given the band width desired, and the variation in gain permitted over this band width, to determine the primary and secondary circuit Q, and then in turn the coil Q's. For narrow-band i.f. amplifiers, such as used in broadcast receivers, the circuit Q's may come out so high as to require commercially impractical coil Q's, since the coil Q must of necessity be higher than the circuit Q because the latter takes into account the additional damping of the tube plate resistance  $R_p$  or the diode load resistance  $R_d$ . In addition, it is necessary to know or assume a value for  $Q_p/Q_s$ .

2. Given the band width desired (for adequate adjacent channel selectivity), and a reasonable primary and secondary coil Q, as well as the additional circuit

damping, to find the variation in gain over the pass band; specifically, the ratio of gain at either peak to the gain at the center resonant frequency.

It is felt that the latter is the more practical problem, and accordingly formulas and graphs have been developed to provide a solution. The method and equations are best described and utilized in conjunction with an actual problem, and for this purpose an i.f. transformer of the following characteristics will be assumed:

$$L_p = L_s = 1.6 \text{ mh}$$

$$Q_p = Q_s = 100$$

For an intermediate frequency of 450 kc, the tuning capacitors  $C_p$  and  $C_s$  are

$$C_p = C_s = \frac{1}{(2\pi \times 450,000)^2 (1.6 \times 10^{-3})}$$

$$= 78.2 \text{ } \mu\text{f}$$

For a Q of 100, the equivalent series resistance of each coil is

$$R_{p_w} = R_{s_w} = \frac{2\pi f L}{Q}$$

$$R_{p_w} = R_{s_w} = \frac{2\pi \times 450,000 \times 1.6 \times 10^{-3}}{100}$$

$$= 45.3 \text{ ohms}$$

Now assume this stage employs a 6K7 i.f. amplifier tube, whose  $R_p$  is 700,000 ohms (this depends on the a.v.c. bias), and  $G_m = 2,000 \text{ } \mu\text{mhos}$ . By Eq. (35), the coupled-in effect of  $R_p$  is

$$\frac{1}{\omega^2 C_p^2 R_p}$$

\*Aiken, C. B., "Two-mesh Tuned Coupled Circuit Filters," *Proc. I.R.E.*, Feb. and June, 1937.

$$= \frac{1}{(2\pi \times 450,000 \times 78.2 \times 10^{-12})^2 \times 7 \times 10^5} = 29.2 \text{ ohms}$$

Hence the total resistance in series with the primary is

$$R_{p t} = 45.3 + 29.2 = 74.5 \text{ ohms}$$

and the primary circuit Q is

$$Q_p = \frac{\omega L}{R_{p t}}$$

$$= \frac{2\pi \times 450,000 \times 1.6 \times 10^{-3}}{74.5} = 60.8$$

Next suppose that the secondary coil feeds a diode load. The effect that the actual load resistance  $r_D$  in the diode circuit has upon the

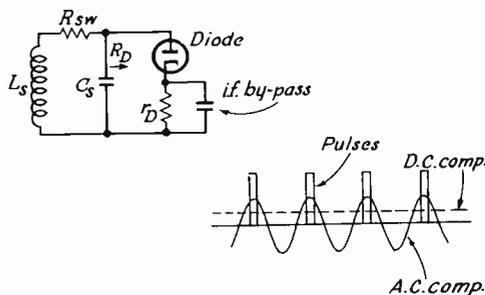


Fig. 34.—Diode circuit and current relationships.

tuned circuit feeding it is as follows: Ignoring the small plate resistance of the diode itself, the diode current flows in a series of short pulses through the bypassed load resistance  $r_D$ , as is shown in Fig. 34. (The circuit is essentially a rectifier with a capacitor-

input filter.

The current pulses can be resolved into a Fourier series. For very narrow pulses, it can be shown that the a.c. fundamental component peak is very nearly twice the d.c. component. The significance of this is that if, for example, an unmodulated i.f. carrier is impressed by the tuned circuit upon the diode circuit, then for every milliampere d.c. flowing through  $r_D$ , 2 ma peak i.f. is drawn from the tuned circuit.

At the same time, the d.c. voltage across  $r_D$  is practically equal to the peak a.c. voltage developed across the tuned circuit, or  $E_{d.c.} = E_p$ . By Ohm's law, the d.c. component is

$$i_{d.c.} = E_p / r_D \tag{44}$$

The corresponding i.f. component drawn from the tuned circuit is

$$i_{a.c.} = 2 i_{d.c.} = 2 E_p / r_D$$

$$= E_p / (r_D / 2) \tag{45}$$

This means that an actual diode load  $r_D$  appears to the tuned circuit as *half* of this value; in short, the effect of  $r_D$  is to act as a resistance  $R_D = r_D / 2$  across  $C_s$ . This lowers the Q of the secondary in exactly the same way in which  $R_p$  lowers the Q of the primary.

Suppose the diode load is 1 megohm. Then it appears as  $R_D = 0.5$  megohm across the secondary coil. By Eq. (35) the equivalent resistance in series with the secondary coil is

$$\frac{1}{\omega^2 C_p^2 R_p} = \frac{1}{(2\pi \times 450,000 \times 78.2 \times 10^{-12})^2 (5 \times 10^5)} = 40.9 \text{ ohms}$$

The  $Q$  of the secondary circuit is therefore

$$Q_s = \frac{2\pi \times 450,000 \times 1.6 \times 10^{-3}}{45.3 + 40.9} = 52.6$$

Then

$$Q_p/Q_s = \frac{60.8}{52.6} = 1.156$$

Note that even if the two coils have identical  $Q$ , the loading of the  $R_p$  of the tube, and  $R_p$  of a diode will alter the circuit  $Q$ 's considerably from either coil  $Q$ . In the case of a mixer tube feeding the grid of an i.f. tube, the primary of the i.f. transformer will be loaded with the  $R_p$  of the mixer tube, whereas the secondary will not be appreciably loaded by the grid of the i.f. tube, and  $Q_p$  will be appreciably less than  $Q_s$ . In the above case of the i.f. stage feeding a diode load, the two  $Q$ 's are reduced more nearly equally.

*Determination of Circuit Behavior.*—Now suppose that a total band width of 10 kc is desired; this is the normal standard (a.m.) broadcast channel width. By Eq. (43), the separation between either peak and the resonant frequency is ( $AB = 10$  kc):

$$\Delta f = 10/2 \sqrt{2} = 3.535 \text{ or } 3.54 \text{ kc}$$

and

$$\frac{\Delta f}{f} = \frac{3.54}{450} = 0.00787$$

The corresponding value of  $\gamma$ , call it  $\gamma_p$ , is

$$\gamma_p = 1 - \frac{\Delta f}{f}$$

$$= 1 - .00787 = .99213 \text{ or } 0.992$$

The required coupling coefficient  $k$  can now be calculated from the following formula:

$$k = \sqrt{\left(\frac{1}{\gamma_p^2} - 1\right)^2 + \frac{1}{2} \left(\frac{1}{Q_s^2} + \frac{1}{Q_p^2}\right)} \quad (46)$$

Substitute the numerical values for the various quantities and obtain:

$$\begin{aligned} k &= \sqrt{\left[\frac{1}{(.992)^2} - 1\right]^2 + \frac{1}{2} \left[\frac{1}{(52.6)^2} + \frac{1}{(60.8)^2}\right]} \\ &= \sqrt{(.0002644) + \frac{1}{2} (.000361 + .000269)} \\ &= \sqrt{.0005794} = .02404 \end{aligned}$$

The next step is to calculate the maximum variation in gain in the pass band for this value of  $k$ . Let  $\alpha_p$  be the average of the two maximum gains occurring at the two resonant peaks.\* Let  $\alpha_r$  be the gain at the center resonant frequency. Then the ratio  $\alpha_p/\alpha_r$  represents the fractional excess in gain at the peaks as compared to the center resonant frequency. The following formula applies:

\*If  $Q_p \neq Q_s$ , the two peaks will be of different heights. The difference is usually small, but Eq. (47) refers to the average of the the gains,  $\alpha_p$ .

$$\frac{\alpha_p}{\alpha_r} = \frac{(k_c/k) + (k/k_c)}{\gamma_p^2} \cdot \frac{1}{\left[ \sqrt{\frac{Q_p}{Q_s}} + \sqrt{\frac{Q_s}{Q_p}} \right] \left[ 1 - \frac{k_c^2}{4k^2} \left( \sqrt{\frac{Q_s}{Q_p}} - \sqrt{\frac{Q_p}{Q_s}} \right)^2 \right]^{1/2}}$$

Note that this equation involves the ratios of certain quantities, such as  $Q_p/Q_s$ . To find  $k_c/k$  or  $k/k_c$ ,  $k_c$  itself must first be determined. This is readily done by means of Eq. (42).

But  $e_c$  is the voltage developed across  $C_s$ , and  $e_g$  is the voltage in series with the primary, and can be expressed in terms of Eq. (37) as

$$e_g = \frac{G_m e_g}{\omega C_p}$$

$$k_c = \frac{1}{\sqrt{Q_p Q_s}} = \frac{1}{\sqrt{60.8 \times 52.6}} = 0.0177$$

Then

$$k/k_c = \frac{.024}{.0174} = 1.38$$

It was also found previously that  $Q_p/Q_s = 1.156$ . If these values are all substituted in Eq. (47) there is obtained

where  $e_g$  is the input signal to the grid of the tube. If this value is properly inserted in Eq. (41) there is obtained

$$\alpha_r = \frac{e_c}{e_g} = \frac{G_m}{\omega_r C_p} \sqrt{\frac{L_s}{L_p}} \frac{(-k)^*}{k^2 + \frac{1}{Q_p Q_s}} = \frac{G_m \sqrt{L_s/C_p} (-k)}{k^2 + k_c^2} \quad (48)$$

$$\frac{\alpha_p}{\alpha_r} = \frac{(1.38) + (1/1.38)}{(.992)^2 \left[ \sqrt{1.156} + \sqrt{1/1.156} \right] \left[ 1 - \frac{1}{4(1.38)^2} (\sqrt{1.156} - \sqrt{1/1.156})^2 \right]^{1/2}} = \frac{(2.104)}{.984} \frac{1}{(2.004) [1 - .131(.0222)]^{1/2}} = 1.067$$

In short, for the circuit  $Q$ 's chosen, and for an overall band width of 10 kc, a coefficient of coupling of 0.024 is required, and the gain will be flat over this band to within 6.7 per cent. This is a very satisfactory performance.

It remains to calculate the gain at resonance, namely  $\alpha_r$ . Using Eq. (41),

$$\frac{e_c}{e_g} \text{ (at resonance)} = \frac{-k \sqrt{\frac{L_s}{L_p}}}{k^2 + \frac{1}{Q_p Q_s}} \quad (41)$$

after some simple algebraic manipulation. Substitute the given numerical values and obtain

$$\alpha_r = \frac{.002 \sqrt{1.6 \times 10^{-3} / 78.2 \times 10^{-12}} (.024)}{(.024)^2 + (0.0174)^2}$$

$$= 246$$

This is a reasonable value of gain, yet not so high as to engender the possibility of oscillation. Indeed, a value of  $\alpha_r = 200$  is not unusual.

\*(-k) means a reversal in phase only.

In the case of a converter tube, the primary side will be loaded with the  $R_p$  of the tube, whereas the secondary will not, since it is normally connected to the grid of the first (and usually only) i.f. stage. Hence,  $Q_p$  will be smaller than  $Q_s$ —possibly half as great. Also, in calculating the gain of the converter stage, the ratio involved is *i.f.* output voltage divided by *r.f.* input voltage. The tube operates not only as an amplifier, but as a frequency converter as well. As a result, the gain formula Eq. (48) involves what is known as *conversion conductance*  $G_c$  instead of *transconductance*  $G_m$ , and  $G_c$  is usually between 1/3 and 1/4 of  $G_m$ . Hence, the gain of a converter tube is appreciably less than that of a straight i.f. amplifier tube.

Typical values are those for a 6SA7 converter tube. For 250 volts on the plate, and 100 volts on the oscillator anode (grids #2 and 4), and approximately zero volts on the control or signal grid (#3), the conversion conductance is  $G_c = 450 \mu\text{mhos}$ , and  $R_p$  is about 1 megohm. If  $G_c$  is substituted in Eq. (48) in place of  $G_m$ , the value of  $\alpha$  is just as readily computed.

*Overall Gain Characteristics.*—The formulas for the gain of an r.f., a mixer, and an i.f. stage have been developed, so that it is now possible to calculate the overall gain of the receiver up to the first audio stage. (The gain of an audio stage has been developed in a previous assignment.)

Usually about 1 volt output is expected from the second detector. The gain from the antenna

to the first grid (of the r.f. or converter tube, as the case may be), is in the neighborhood of 6. The gain of the converter tube and first i.f. amplifier can readily be calculated on the same basis as the i.f. stage analyzed previously; a representative gain is about 70. The example of an i.f. amplifier discussed in this assignment had a gain of 246. Assume the r.f. stage analyzed previously; the gain at 540 kc was computed as 20.2.

The overall gain is the product of the individual gains, or

$$\begin{aligned}\alpha_t &= 6 \times 20.2 \times 70 \times 246 \\ &= 2,084,000\end{aligned}$$

Thus, for 1 volt output from the the second detector, an input voltage from the antenna of

$$\frac{1 \times 10^6}{2,084,000} = .479 \mu\text{volts}$$

is required. This is about one-hundredth of the minimum considered acceptable for entertainment purposes in the broadcast band, and indicates that the receiver has more than sufficient gain to handle the average broadcast signal, so that the average signal will actuate the a.v.c. circuit and thus enable the receiver to function properly.

It is also to be noted that the gain will be higher at the top end of the band, owing to the rise in gain in the r.f. stage, as has been discussed previously. Such variations in gain can also be easily handled by the a.v.c. circuit.

## RECEIVER ADJUSTMENTS

*RECEIVER ALIGNMENT.*—This

discussion will apply to the aligning of a receiver just as it comes off the assembly line or to the alignment of a receiver which has been in service for some time and needs realignment. It is assumed that all circuits are electrically correct and capable of functioning in the normal manner when they are properly adjusted.

In the alignment of the receiver it is ordinarily best to work from the demodulator or second detector toward the antenna. The following equipment is recommended even though it is possible to do a very good aligning job with less than these items: A specially made insulated screw driver or aligning tool, a test oscillator or signal generator which is reasonably accurate as to frequency calibration and which is capable of being modulated from 30 to 40 per cent at 400 cycles or so, and an output meter. The use of the cathode ray oscilloscope will not be discussed in this assignment.

The receiver must be aligned at the proper frequencies otherwise the receiver tuning dial indicator will not coincide with the frequency calibrations on the dial. This indicates the need for a test oscillator whose output frequency is dependable. It is also desirable to remove the chassis from the cabinet. The chassis and test oscillator should then be allowed to warm up to normal operating temperature. This will reduce the possible effect of oscillator drift and its effect on alignment. The warm-up period will usually be from 15 to 30 minutes. A chassis from a closely confined cabinet may not reach the same temperature on the test bench as in the cabinet,

but the effect of drift will be minimized.

The tuned circuits immediately preceding the demodulator in the superheterodyne will be aligned at the intermediate frequency. The test oscillator should be set accurately to this frequency. The high-frequency oscillator section of the receiver should be made inoperative to avoid the production of any beat in the oscillator-mixer section of the receiver. This may be done in a variety of ways. If the oscillator is a separate tube, it may be removed from the socket provided its removal does not upset the proper voltage and current values in other circuits of the receiver. Or the tuning condenser associated with the oscillator may be shorted out by a clip lead or other device. Make certain, however, that no high-frequency oscillation in the oscillator is produced which might interfere with the alignment. Study of the schematic diagram of the receiver will suggest other methods equally effective. In the case of a pentagrid mixer tube, the oscillator section cannot be removed from the circuit physically, and the second suggested method may be employed. In general, shorting out the oscillator tuning condenser will suffice.

The output meter is normally a rectifier type high sensitivity a.c. voltmeter. A vacuum tube voltmeter may also be used if desired. In any event, the meter used must be an a.c. device. The meter is connected in the audio circuit of the receiver, a convenient connection being across the voice coil. Since the voice coil is usually of low impedance, the 0-10 volt a.c. scale is normally satisfactory.

It may be necessary to resort to either a lower or higher range scale during the aligning process but just which scale to use will depend on the audio output of the receiver. The magnitude of the deflection on the output meter for any scale will readily indicate whether or not the scale chosen is proper.

If the intensity of the modulated signal from the loudspeaker is annoying, the voice coil winding may be unsoldered (if convenient) and a resistance equal to the voice coil impedance connected in the circuit in its place. The output meter is then connected across the resistance. However, the audible signal from the speaker may be of value in the earlier stages of alignment in indicating just how the alignment is proceeding.

The output meter may also be connected across the primary of the output transformer and a high-range voltage scale employed. It is desirable to keep the d.c. out of the meter so a 1- $\mu$ f condenser should be connected in series with the meter. The idea is to measure the a.c. voltage appearing across any output load in the a.f. section and to use this value as an indication of proper resonance conditions in the r.f. or i.f. sections of the receiver.

If the receiver is a test model, the designer will be familiar with any special information relating to alignment. If the receiver is being checked for proper alignment and the person conducting the alignment procedure is not thoroughly familiar with it, he is urged to read the manufacturer's instructions noting any special instructions. The proper alignment of the receiver is very vital

to its proper operation in service and no deviations from the stated aligning frequencies should be made. In general, it will be impossible to duplicate the frequency dial settings for each band unless the proper and exact aligning frequencies are employed.

Practically all modern receivers employ a.v.c. The normal a.v.c. action of the receiver is such as to "drag" the tuning across the aligning frequency and to distort the familiar resonance curve of tuned circuits. Reflection on the operation of a.v.c., namely that of keeping the output fairly constant for changing values of input, will show why this is so.

As the receiver is detuned by a small amount from the mid-band frequency, the a.v.c. will operate to keep the output constant by keeping the input to the demodulator constant. This means that for proper identification of the mid-band frequency it is necessary that the a.v.c. be made inoperative during alignment. The rendering of the a.v.c. circuits inoperative may be done in several ways.\* The simplest scheme, and one which is applicable to practically every existing receiver model of recent years, is that of grounding the main a.v.c. filtering condenser. To avoid mistakes, it is recommended that this condenser be identified and that the ground be made at the condenser by connecting a clip lead directly across it. The proper method is shown in Fig. 35.

\*In receivers using d.a.v.c. with a fixed minimum bias, the alignment can be done by keeping the input signal very low so a.v.c. action does not occur.

If the receiver employs d.a.v.c., the receiver may be aligned without disturbing the a.v.c. circuit by employing a signal from the test oscillator of such magnitude that it does not operate the a.v.c. circuit. If the receiver employs a separate tube for a.v.c. and the

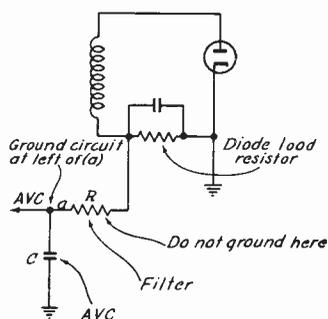


Fig. 35.—Schematic of a.v.c. circuit showing where to ground it out.

removal of this tube will not affect the normal operation of the receiver, the tube may be taken from its socket.

If the receiver employs a resonance indicator, in certain cases this may be used to indicate alignment. In general, the use of an output meter is recommended. The resonance indicator will indicate a peak of resonance condition exactly the same as will the output meter since a maximum indication on this device operates exactly the same way that the output meter does except that the resonance indicator operates on the rectified or average value of the carrier while the output meter operates on the audio modulation voltage which is a maximum for resonance.

**ALIGNMENT OF THE I.F. SECTION.**—Assume that an output meter is being employed and that it is connected across the voice coil of the speaker; that the a.v.c. system

is inoperative or ineffective; that the proper i.f. has been identified; that the test oscillator is properly set to this i.f. and that both it and the receiver have warmed up to the proper operating temperature; and finally that the receiver to be aligned has two i.f. stages.

The output attenuator on the test oscillator is set at maximum output, and the ground connection on the test oscillator is connected to the receiver ground, which is normally the chassis. The "high" side of the test oscillator is now connected to the input circuit of the last i.f. stage. This may be done variously. The principal point to remember is that the tube in the last i.f. stage should operate normally which means that provision to give this tube its proper minimum bias must be insured. In most modern test oscillators this is easily provided since there is a d.c. path through the attenuator on the output of the test oscillator. This means that the d.c. bias voltage is applied in series with the r.f. driving voltage through this attenuator.

See Fig. 36(A). In this figure it is observed that the bias is obtained from a fixed resistor in the cathode circuit and that the negative end of this resistor is connected to ground. Now if the input circuit to the last i.f. tube is opened at x, which is conveniently done by lifting the grid clip, and the "high" output terminal is connected to the grid of the tube, the d.c. path is provided through the attenuator resistance. If the tube has a shield, the shield should be in normal position. If the tube is metal with a grid cap, shielding

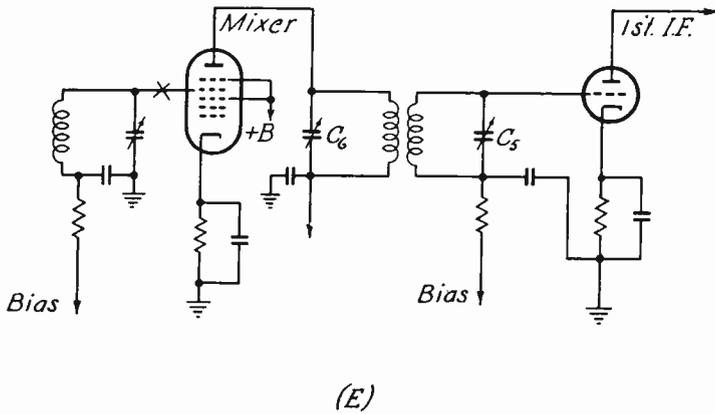
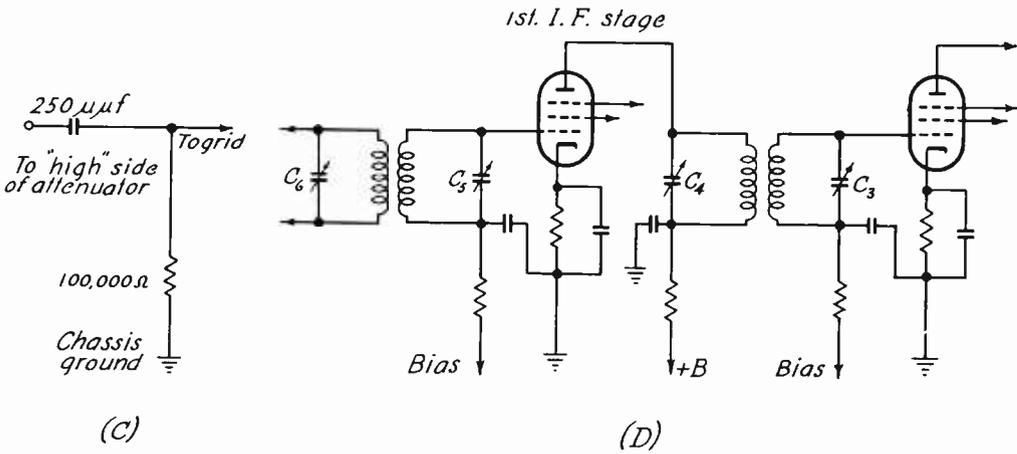
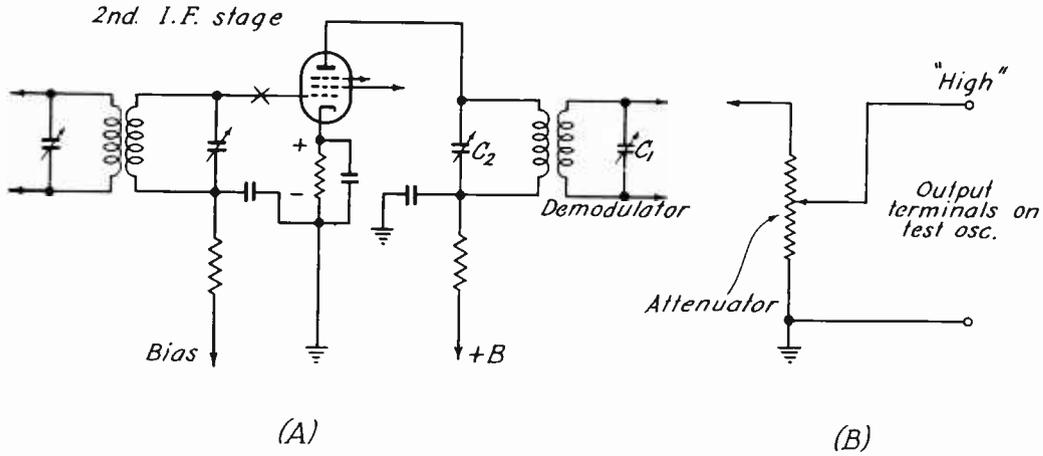


Fig. 36.—I.F. amplifier stages, showing connections for test oscillator and capacitors to be adjusted.

is not a problem since the metal shell is the shield. If the tube is of the "S" type, such as the 6SK7, the grid lead may be unsoldered at the socket. An alternative method is to employ a capacitor-resistor network as shown in Fig. 36(C). The values for the capacitor and resistor are not critical, suggested values are shown.

The aligning may now proceed as follows: The volume control on the receiver is set to maximum and the attenuator on the test oscillator also similarly set, as stated above, and modulated output from the test oscillator fed to the stage. Capacitor  $C_2$  is now varied with the aligning tool (normally an insulated screw driver which may be made by inserting a small metal fin in the the end of a small bakelite rod or purchased for 25 cents or so from a radio supply house) until the output meter reads a maximum. The indication on the input meter is also supplemented in this case by an aural response from the speaker. If the output meter reading is insufficient, change the meter scale to a more sensitive range. The meter reading will be somewhat broad, but the maximum value can be readily identified.

Now adjust  $C_1$  for resonance which is also indicated by maximum response on the output meter. If the demodulator is a diode, the tuning is also somewhat broad but is readily determined. Next, return to  $C_2$  and check it for resonance since the secondary circuit may react on the original tuning of the input circuit.

After one is certain of the correctness of the adjustments of  $C_2$  and  $C_1$ , these tuned circuits are

now considered aligned and the circuit is returned to normal by reconnection of the grid clip or by resoldering any leads which may have been removed.

The lead from the "high" side of the attenuator is removed from the grid of the last i.f. tube and re-applied to the grid of the first i.f. tube (if two i.f. stages are employed), with other circuit changes being exactly as stated for the procedure outlined above. The tuned circuits associated with the first stage output are aligned in exactly the same way. The capacitors now affected are  $C_3$  and  $C_4$  and can be identified in Fig. 36(D). It will be necessary to reduce the output from the signal generator to avoid overload to the last stage. Whether or not the last i.f. stage is overloaded is readily observed by an aural check on the audio output. It may also be necessary to change the scale on the output meter to a less sensitive range. This will normally mean going to a higher voltage scale.

When this stage is properly aligned, the process is now carried to the mixer tube. The process is again identical except that capacitors  $C_5$  and  $C_6$  are the ones to be adjusted. Because of the increased gain, it will again probably be necessary to reduce the output of the test oscillator and possibly to again choose a new scale on the output meter. The main things to remember are to avoid overload of any stage and to get a reasonable indication on the output meter.

This last series of capacitor adjustments constitutes the final alignment at the i.f. frequency. It is desirable to recheck the settings of all the capacitors to be certain

that the best peaking value has been determined. The i.f. alignment is now complete. The test oscillator is disconnected and all circuits are returned to their normal electrical connections.

In the method of alignment, it has been suggested that the input circuits be removed. In certain cases this is not necessary, and the output of the test oscillator can be clipped directly to the grid circuit without the necessity of removing the input circuit. The principal consideration to take into account is whether or not the input circuit will adversely affect the alignment. In most modern receivers the removal of the input circuit is not necessary. Some manufacturers specify certain methods which they recommend for alignment of the i.f., and this information should be carefully followed if a proper aligning job is to be performed.

A second and very important point is that all i.f. transformers are built to provide a certain value of band pass. In communication receivers for telegraphic reception, the band pass is intentionally very narrow with the result that positive peaking is relatively easy. However, receivers intended for the reception of waves carrying relatively wide frequency excursions demand the familiar flat peak characteristic which may vary in width between the peaks from several hundred cycles to several thousand cycles. Television and f.m. receivers demand an even wider band pass.

For most broadcast band receivers, the band pass varies from 8,000 to 10,000 cycles total. This carries the special significance

that the output of the i.f. amplifier must be uniform over a range of 5 kc/s or so on each side of the mid-band i.f. frequency. The alignment of these types of i.f. transformers often demand special attention, and it is always best to carefully consult the manufacturer's aligning instructions. Lacking knowledge of these instructions, the following information will be found useful. If one employs a cathode-ray oscilloscope, the procedure is essentially the same and the certainty of results is improved. Nevertheless, a good aligning job may be done without the cathode-ray oscilloscope.

The transformers are first peaked at the i.f. frequency, and then the primary and secondary are separately tuned, one above and the other below the i.f. frequency by the proper amount to obtain the band pass. For example, if the band pass is 8 kc/s, the i.f. frequency 460 kc/s, the primary might be tuned to 456 kc/s (4 kc/s below) and the secondary to 464 kc/s (4 kc/s above). The variation in output on the meter will indicate the uniformity of pass as the test oscillator tuning is slowly varied from the low to the high frequency of band pass through the mid-band frequency. The adjustments are varied until the desired uniformity is obtained.

It is usually necessary to check these settings carefully and several times before one can be sure of the adjustment. A final check is the application of a modulated signal from a high quality broadcast station when the receiver is finally completely aligned.\*

\*This procedure should not be necessary if the circuit Q's and

**ALIGNMENT OF THE R.F., MIXER AND OSCILLATOR.**—The alignment of the "tuned" end of the receiver consists of obtaining the proper tracking of the various tuned circuits associated with the tuning of the receiver. In general, one always finds at least two tuned circuits and possibly four or five, depending on the cost and desired operating characteristics. For those receivers with a mixer tuned circuit and an oscillator tuned circuit there are but two, and receivers with one stage of pre-selection provide at least three tuned circuits. However, the addition of additional tuned circuits presents no great difficulty.

Earlier in this assignment it was shown that to provide tracking of identical tuned circuits certain basic conditions must be met, and that the alignment was possible through the use of a small capacitor connected across each coil. Also it was shown that the tracking of the oscillator was made possible by the use of a parallel and a series capacitor when identical capacitors are used to tune the oscillator and r.f. circuits. These small capacitors in conjunction with the main tuning capacitors are the ones used in the high-frequency alignment.

The test oscillator is connected to the antenna and ground posts with the high side of the attenuator at the antenna position.

coupling  $k$  were correct originally and have not changed subsequently. However, if some changes have occurred, such detuning will compensate for these changes. Thus, it mainly increases the coupling coefficient somewhat over that produced by the fixed spacing of the primary and secondary coils of the i.f. transformer, and thereby increases the band width.

The oscillator section of the receiver is now rendered operative and the aligning frequencies chosen. In the case of the new receiver, these frequencies have been chosen by the designer; in the case of a receiver to be realigned, these frequencies are determined from the manufacturer's data. For the broadcast band, the alignment frequencies are normally 1,500 kc/s, 1,000 kc/s and 600 kc/s.

Suppose the receiver to be aligned at the broadcast band uses 1,500, 1,000 and 600 kc/s. The test oscillator is set at 1,500 kc/s. The modulated output is fed to the receiver with the attenuator setting at a reasonable value. The output meter connection is left unchanged. If the manufacturer recommends a dummy antenna between the test oscillator and the receiver this should be provided.

The schematic for a typical receiver at any band is shown in Fig. 37. This receiver employs an r.f. stage, a mixer tuned circuit and an oscillator circuit in connection with a pentagrid converter.  $C$  is the tuning capacitor and each section is identical.  $C_0$  is the oscillator padding capacitor for alignment at the 600-kc position.  $C_2$  is the oscillator trimmer capacitor for alignment at the 1,500-kc position, while  $C_1$  in both the mixer and r.f. stage is for alignment at the 1,500-kc position.

The procedure is as follows: Tune the test oscillator to the high-frequency aligning position, in this case 1,500 kc/s. Set the receiver dial to 1,500 kc/s. Adjust  $C_2$  for maximum output meter reading. Similarly, adjust  $C_1$  on the r.f. and mixer for maximum output.  $C_2$  will be the most critical ad-

justment of these three capacitors. Use the aligning tool for this purpose. An ordinary screw driver is not recommended. Make sure that all adjustments give maximum output.

Next, set the test oscillator output to 600 kc/s, and similarly adjust the tuning dial on the

circuit capacity. Then the frequency is changed to 600 kc/s and  $C_s$  is checked. This process is continued until the circuits are properly tracked at the 1,500 and 600 kc/s frequencies.

The 1,000-kc/s position is now checked and the receiver should be

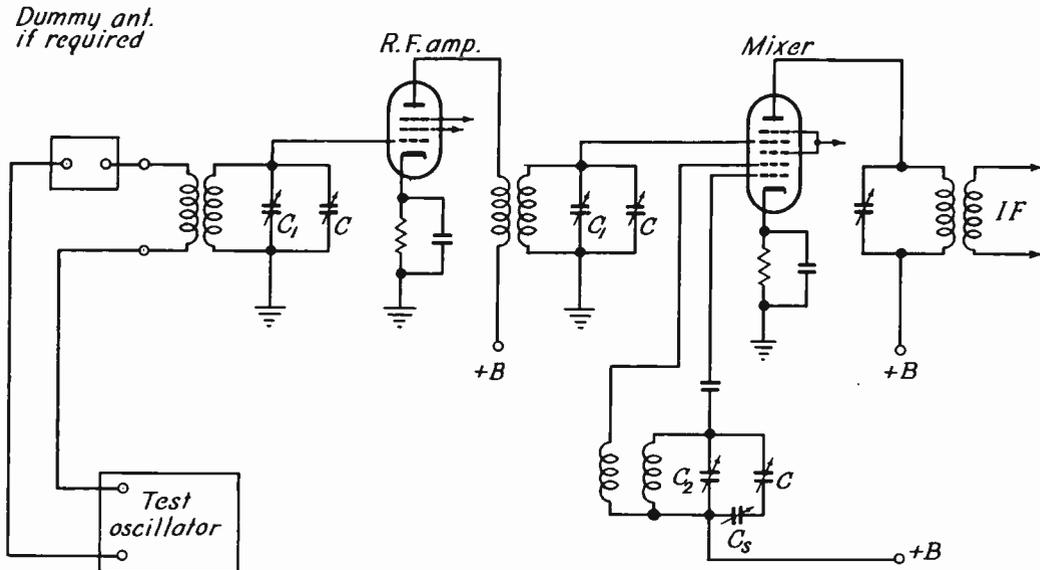


Fig. 37.—Typical test setup for aligning the r.f. end of a receiver.

receiver to 600 kc/s. Now adjust  $C_s$  for maximum output at 600 kc/s. Do not touch the adjustment of  $C_1$  or  $C_2$ . The adjustment of  $C_s$  should be made while the tuning dial is rocked slightly about the 600-kc/s position. The setting of  $C_s$  which produces maximum output on the output meter is the proper one even though this setting of  $C_s$  does not occur with the tuning capacitor exactly at 600 kc/s.

The test oscillator and receiver tuning are now returned to the 1,500-kc/s position, and the adjustment of  $C_2$  and possibly  $C_1$  are checked. Since  $C_2$  and  $C_s$  are in a parallel arrangement a change of one will produce a change in the total

reasonably close to 1,000 kc/s setting. It usually will be. If not, the circuit is defective in the proper choice of oscillator tuning inductance, or the i.f. was aligned at the wrong frequency, or the dial has slipped on the shaft. The first is unlikely normally, the second and third conditions can be readily checked and corrected.

If the receiver has more than one band, the same procedure on the other bands is followed exactly as outlined for the broadcast band. The i.f. need not be changed for the higher bands since it always operates at the same frequency regardless of the input signal frequency. In general, the align-

ment of the higher frequency bands is simpler than the alignment of the broadcast band. This is so because the higher and higher frequency bands operate with a smaller and smaller percentage difference between the oscillator and r.f. signal sections.

### SUMMARY

This concludes the assignment on receivers for reception of amplitude modulation. A general survey of receiver requirements was first presented, and the usual types of amplifier circuits briefly discussed.

Then an analysis of two actual types of receivers was presented: for a table model and for a console type of receiver. The various circuit components, particularly the

mixer and oscillator circuits, were studied in order to see how the requirements are met in actual models.

Following this a section on receiver calculations was presented. Problems as to the calculation of tuning capacitor range, tracking requirements, and r.f. and i.f. design considerations, were furnished to enable the student intelligently to specify or design the various coil and capacitor components in order that the receiver will meet certain design specifications.

Finally, a section on receiver adjustments was given to enable the student to adjust a receiver for optimum performance. Such matters as r.f. and i.f. alignment were discussed in some detail to permit the student to gain an insight into one of the fundamental activities in servicing.

APPENDIX I

DERIVATION OF THE EQUATION  
USED IN THE SUPERHETERODYNE TRACKING  
PROBLEM

The resonant frequency of the r.f. section for any setting of  $C_t$  is

$$F_r = \frac{1}{2\pi \sqrt{L_R(C_t + C_{HR})}}$$

and for proper tracking this frequency added to the i-f frequency must equal the oscillator frequency,

or:

$$\frac{1}{2\pi \sqrt{L_R(C_{HR} + C_t)}} + \text{i. f.}$$

$$= \frac{1}{2\pi \sqrt{L_o \left( C_{HO} + \frac{C_t C_p}{C_t + C_p} \right)}} \quad (1)$$

in which the last term in the right-hand member represents  $C_t$  and  $C_p$  in series.

If this equation is solved for  $C_t$  in terms of  $C_{HO}$ ,  $L_o$ ,  $C_{HR}$ ,  $L_R$ ,  $C_p$  and i. f. it will be found that there are three possible values of

$C_t$  and only three that will satisfy the equation. Conversely, three arbitrary values of  $C_t$  between the maximum and minimum values that the tuning capacitor can assume, may be chosen and fixed POSITIVE REAL values of  $L_o$ ,  $C_{HO}$ ,  $C_p$ ,  $L_R$  and  $C_{HR}$  found such that tracking occurs at the three chosen values of  $C_t$ . Let the first setting of  $C_t = C_1$ . At this setting the r.f. section will be resonant to a frequency  $F_1$ . Add to  $F_1$  the i. f. frequency and call the result  $F_{10}$ . This is the frequency of the oscillator section if tracking is to obtain, so that

$$F_{10} = \frac{1}{2\pi \sqrt{L_o \left( C_{HO} + \frac{C_1 C_p}{C_1 + C_p} \right)}}$$

or  $(F_{10})^2$

$$= \frac{1}{4\pi^2 L_o \left( C_{HO} + \frac{C_1 C_p}{C_1 + C_p} \right)} \quad (2)$$

Solving this for  $L_o$  by simple transposition yields the third of the three equations given in the text.

$$L_o = \frac{1}{4\pi^2 f_{10}^2 \left( C_{HO} + \frac{C_1 C_p}{C_1 + C_p} \right)} \quad (a)$$

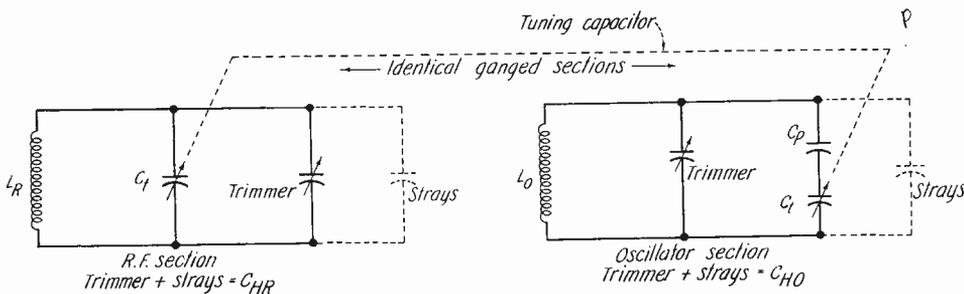


Fig. 1.—The r-f and oscillator tuning section of a superheterodyne receiver.

In a similar manner, corresponding to a setting of  $C_t$  equal to  $C_2$  we may write,

$$(f_{20})^2 = \frac{1}{4\pi^2 L_o \left( C_{H0} + \frac{C_2 C_p}{C_2 + C_p} \right)} \quad (3)$$

Dividing Eq. (2) by Eq. (3) produces,

$$\left( \frac{f_{10}}{f_{20}} \right)^2 = \frac{C_{H0} + \frac{C_2 C_p}{C_2 + C_p}}{C_{H0} + \frac{C_1 C_p}{C_1 + C_p}} \quad (4)$$

This may readily be solved for  $C_{H0}$  and gives the second of the three equations in the text:

$$C_{H0} = \frac{\frac{C_1 C_p}{C_1 + C_p} \left[ \frac{f_{10}}{f_{20}} \right]^2 - \frac{C_2 C_p}{C_2 + C_p}}{1 - \left[ \frac{f_{10}}{f_{20}} \right]^2} \quad (b)$$

By the same process, considering that a setting of  $C_t$  equal to  $C_3$  results in an oscillator frequency  $f_{30}$  it will be found that

$$C_{H0} = \frac{\frac{C_1 C_p}{C_1 + C_p} \left[ \frac{f_{10}}{f_{30}} \right]^2 - \frac{C_3 C_p}{C_3 + C_p}}{1 - \left[ \frac{f_{10}}{f_{30}} \right]^2} \quad (c)$$

To derive the equation for the value of  $C_p$ , first for ease in handling let

$$\left[ \frac{f_{10}}{f_{20}} \right]^2 = A$$

and

$$\left[ \frac{f_{10}}{f_{30}} \right]^2 = B.$$

Then divide Eq. (b) by (c) to obtain

$$\frac{\frac{AC_1 C_p}{C_1 + C_p} - \frac{C_2 C_p}{C_2 + C_p}}{\frac{BC_1 C_p}{C_1 + C_p} - \frac{C_3 C_p}{C_3 + C_p}} \times \frac{[1 - B]}{[1 - A]} \quad (5)$$

Make use of common denominators and cancel as shown

$$\frac{AC_1 C_p (C_2 + C_p) - C_2 C_p (C_1 + C_p)}{(C_1 + C_p) (C_2 + C_p)} \times \frac{[1 - B]}{[1 - A]} = 1$$

$$\frac{BC_1 C_p (C_3 + C_p) - C_3 C_p (C_1 + C_p)}{(C_1 + C_p) (C_3 + C_p)} \quad (6)$$

$$\times \frac{[1 - B]}{[1 - A]} = 1$$

Invert, multiply and rearrange so as to have

$$[1 - B] [AC_1 C_p (C_2 + C_p) (C_3 + C_p) - C_2 C_p (C_1 + C_p) (C_3 + C_p)] - [1 - A] [BC_1 C_p (C_2 + C_p) (C_3 + C_p) - C_3 C_p (C_1 + C_p) (C_2 + C_p)] = 0 \quad (7)$$

Multiply out, cancel where possible and factor out  $C_p$  to obtain:

$$C_p [AC_1 C_2 + AC_1 C_p - C_1 C_2 - C_2 C_p + BC_2 C_3 + BC_2 C_p - BC_1 C_3 - BC_1 C_p + C_1 C_3 + C_3 C_p - AC_2 C_3 - AC_3 C_p] = 0$$

One solution of this would be  $C_p = 0$  but a glance at the circuit diagram shows that it is an inadmissible root since a zero capacity is an open circuit and tuning would be

impossible. Therefore the quantity within the brackets must be equal to zero. This is easily solved for  $C_p$  and after factoring results in

$$C_p = \frac{C_1 C_2 (1 - A) + C_2 C_3 (A - B) - C_1 C_3 (1 - B)}{C_3 (1 - A) + C_1 (A - B) - C_2 (1 - B)}$$

This is a complete solution but may be transformed, if desired, into the form used in the text thus: Write the quantity  $(A - B)$  as  $(A - 1 + 1 - B) = [ - (1 - A) + (1 - B) ]$  and substitute to obtain

$$C_p = \frac{C_1 C_2 (1 - A) - C_2 C_3 (1 - A) + C_2 C_3 (1 - B) - C_1 C_3 (1 - B)}{C_3 (1 - A) - C_1 (1 - A) + C_1 (1 - B) - C_2 (1 - B)}$$

Factoring this gives

$$C_p = \frac{-(1 - A)(C_3 - C_1)C_2 + (1 - B)(C_2 - C_1)C_3}{(1 - A)(C_3 - C_1) - (1 - B)(C_2 - C_1)}$$

Multiply numerator and denominator of the right member by  $-1$  and then divide by  $(1 - B)(C_2 - C_1)$  to get

$$C_p = \frac{\left[ \frac{1 - A}{1 - B} \right] \left[ \frac{C_3 - C_1}{C_2 - C_1} \right] (C_2) - C_3}{1 - \left[ \frac{1 - A}{1 - B} \right] \left[ \frac{C_3 - C_1}{C_2 - C_1} \right]} \quad (d)$$

which is the formula as it appears in the text since

$$A = \left( \frac{f_{10}}{f_{20}} \right)^2$$

and

$$B = \left( \frac{f_{10}}{f_{30}} \right)^2$$

From Eq. (d), Eqs. (a) and (b) can be solved for  $L_o$  and  $C_{H0}$ , respectively. Then, knowing the magnitude of the stray capacity, it can

be subtracted from  $C_{H0}$  to yield the value of the trimmer capacitor in the oscillator circuit.