ELECTRONIC TECHNOLOGY SERIES

SUPERHETERODYNE CONVERTORS and I-F AMPLIFIERS

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SUPERHETERODYNE CONVERTERS AND I-F AMPLIFIERS

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PREFACE

The utilization of heterodyning action in receiver design via local oscillator, mixer, or converter action marks one of the major steps in the advance of communications. Application of the basic principles of superheterodyne operation solved many of the problems inherent in the earlier tuned radio frequency receivers. Such factors as receiver stability, gain, selectivity, and uniform bandpass over an entire band could be improved by using the superheterodyne receiver. The reasons for the enormous popularity of this design are apparent, as is the need for the technician to understand the theory and operation of superheterodyne converters and i-f amplifiers. This book is organized to provide the student with an understanding of these fundamental principles, with emphasis on the descriptive treatment and analyses. Mathematical formulas or numerical examples are presented where pertinent and necessary to illustrate the discussion more fully.

Specific attention has been given to the essential theory of mixers and converters; basic superheterodyne operation; arithmetic selectivity; image frequency considerations; double conversion; conversion efficiency; oscillator tracking; pulling and squegging; types of converters (both early and modern) ; functions and design factors of i-f amplifiers; choices of i-f frequencies; avc and davc; the Miller effect; and the consideration of alignment procedures. The description of tracking alignment is included, since this book deals primarily with applied theory.

The second edition adds a number of modern topics relating to mixer and converter technology. Doubling the size of the original work, this edition includes general design considerations of mixers, converters, oscillators, and i-f amplifiers which supplements the fundamental ideas presented in the first edition. A detailed discussion of the transistorized counterpart of the vacuum tube versions of superheterodyne converters and i-f amplifier circuity completes the revised work.

Grateful acknowledgment is made to the staff of New York Technical Institute for its assistance in the preparation of the manuscript for this book.

A. S.

New York, N. Y.
July 1963

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Chapter 1

BASIC PRINCIPLES OF SUPERHETERODYNE OPERATION

1. Basic TRF Operation

Prior to the design and development of the superheterodyne (superhet) receiver the tuned radio frequency (trf) set was the most popular type for home entertainment and commercial use. As shown in Fig. 1 the incoming r-f signal is coupled from the antenna to the first of severable tunable r-f stages, in which the received signal is succesively amplified. The amplified r-f signals are then fed to the detector stage, where the r-f carrier is removed and the audio intelligence obtained for application to the audio and loudspeaker sections. Tuning of the various r-f resonant circuits is accomplished by rotation of a multi-section variable tuning capacitor.

In trf receivers, a frequent problem was instability and danger of oscillation, since adjacent stages were all tuned to very nearly the same frequency. This important objection, together with the difficulty involved in obtaining high gain, proper selectivity, and uniform bandpass over the entire band, was overcome with the design of the superhet receiver.

2. Basic Superhet Operation

The superhet receiver (Fig. 2) employs a heterodyne or "signal beating" principle to convert the incoming r-f signals to a carrier of a fixed, lower frequency containing the same audio modulation as the original carrier. This lower frequency signal is fed to an amplifier section, which (because it is fixed-tuned) can be designed for optimum gain. Also, its bandpass characteristic can be tailored to any desired response for purposes of fidelity and selectivity. As shown in Fig. 2, the incoming r-f signal is selected by a tuned r-f

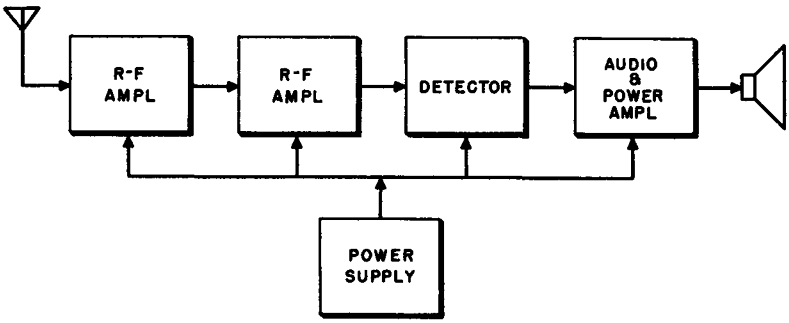

Fig. 1. Block diagram of a TRF receiver.

circuit, amplified in the same manner as in a trf set, and then fed to a mixer (converter) stage. At this point, a locally generated oscillator signal of constant amplitude heterodynes or "beats" with the modulated r-f signal; the difference frequency signal (oscillator frequency minus rf, or vice versa) resulting from the mixing action is selected in the tuned output circuit. The mixer output signal contains the same audio information as the original r-f signal, but employs a lower carrier frequency.

As the receiver is tuned to accept different stations, the local oscillator tuned circuit and the r-f amplifier and mixer tuned circuits are changed so that a constant frequency difference always exists. The action in the mixer stage may be considered a form of modulation or detection since it is here that the r-f carrier is removed and the audio modulation is impressed on a lower frequency (i-f) carrier. For this reason, it is common to refer to the mixer stage as the *first detector,* as distinguished from the second, or audio, detector at which the i-f carrier is detected and the audio information obtained. The i-f signal, selected by the fixed tuned circuit in the mixer output, is amplified by one or more fixed-tuned i-f amplifier stages before application to the audio detector.

3. Mixers and Converters

Any single tube or circuit performs the task of mixing or heterodyning two signals, as well as developing the necessary local oscillator energy, is termed a "converter." A "mixer" tube, on the other hand, combines the two signals, but is supplied (either by inductive, capacitive, or combined coupling) by a separate local oscillator stage. In most modern a-m broadcast sets, a single combination converter tube is used for economy and simplicity in wiring and assembly. However, in multi-band receivers, in which high frequency (short wave) signals are to be received, it is often necessary to utilize separate mixer and oscillator tubes for improved selectivity and stability.

Superhet design forms the basis for high-fidelity f-m tuners and for tv receivers as well as simple five-tube table model a-m radios. The principles of conversion apply in all types and the differences in operation are due mainly to the characteristics of that portion of the r-f spectrum occupied by each.

4. Advantages of the Superhet Over the TRF Receiver

The development of the superhet has resulted in a high gain receiver possessing excellent stability characteristics. Since the

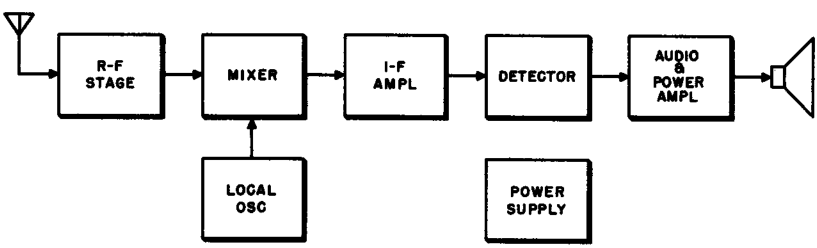

Fig. 2. Block diagram of a typical superheterodyne receiver.

greatest amplification takes place in the fixed-tuned i-f stages, a uniform degree of amplification can be achieved, regardless of the point on the broadcast band to which the set may be tuned. In the trf receiver, the various r-f stages are resonant at the same frequency, thus creating the possibility of instability; the r-f, oscillator, and i-f stages of the superhet are operating at separate frequencies, thereby greatly reducing the tendency toward oscillation due to interstage coupling. A third and decidedly important advantage is termed "arithmetic selectivity."

5. Arithmetic Selectivity

Assume that a short wave station transmitting at 2800 kc is to be received, and that a nearby station is operating at 2810 kc. It would be extremely difficult, if not impossible, for a trf receiver to tune in the 2800-kc signal without interference from the 2810-kc station, because of the small frequency difference of only 10 kc; the station spacing would be 10 parts out of 2800 or approximately 0.36 percent.

Using a superhet receiver with a 455-kc i-f amplifier stage, the mixing action of the local oscillator with the desired 2800-kc signal would develop a 455-kc i-f and a 465-kc difference frequency for the 2810-kc station. The 10-kc station spacing is now 10 parts out of 455 or approximately 2.2 percent, thus effectively making it six times easier to tune in the desired station without interference.

In addition, it is not difficult to design the i-f section so that its response will be sharply reduced at 10 kc from its resonant frequency of 455 kc; this further attenuation of the 465-kc signal decreases the level of the interfering station, resulting in clearer reception and greater ease of tuning, even in crowded portions of the r-f spectrum.

6. Operation of the Mixer Stage

Consider an incoming r-f carrier at 800 kc with a 1-kc audio modulation note entering a superhet receiver with a 455-kc i-f section. With the r-f input tuning capacitor set to resonate at 800 kc, the ganged oscillator capacitor tunes its circuit (*L1*) to 1255 kc. The constant amplitude, locally generated (oscillator) signal is considerably larger than the relatively weak r-f signal, and both are applied to the mixer stage by one of several methods discussed later. The mixer tube is biased by a cathode resistor, so that the two signals swing in the nonlinear region of the tube characteristic curve. The simultaneous application of the r-f and local oscillator signals in the nonlinear detector stage results in a multitude of output signals of different amplitudes. The oscillator signal, and signals of frequencies equal respectively to the difference of the r-f signal and oscillator frequencies predominate. Harmonics of these signals, and their sum and difference components are also present, but are lower in magnitude. The desired i-f or difference frequency signal in the previously-mentioned case is 455 kc, although relatively high-amplitude 800-kc, 1255-kc, and 2055-kc signals are available at the mixer output.

Figure 3 is a circuit diagram of a modern 6BE6 pentagrid converter stage. The plate or output circuit of the mixer consists of a

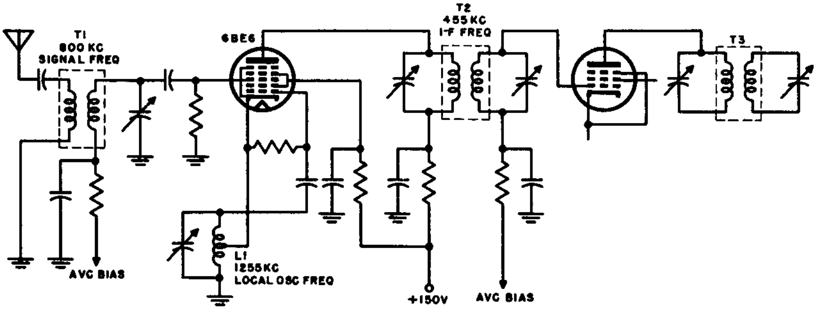

Fig. 3. Schematic diagram of a 6BE6 converter.

parallel resonant circuit fixed-tuned to 455 kc. The 455-kc difference signal, which still carries the original 1-kc audio modulation is offered high impedance at its resonant frequency, and consequently a large 455-kc voltage is developed across the primary winding of i-f transformer *T2* for application to the i-f amplifier section. The 800-kc, 1255-kc and 2055-kc signals are offered low impedance by the 455-kc tuned circuits of *T2*, and are thus shunted or bypassed from the i-f section. The tuned circuits in the mixer output and i-f amplifier stages generally consist of i-f transformers (*T2* and *T3*), with primary and secondary resonant circuits adjusted for desired response, selectivity, and gain.

7. Image Frequency

Perhaps the most serious drawback to superhet operation is the possibility of image interference. This interference is possible because there are two r-f signal frequencies at which incoming signals can mix with the local oscillator signal to produce the desired intermediate frequency. For example, the receiver previously discussed utilizes a 455-kc i.f. and the local oscillator is set at 1255 kc in the reception of an 800-kc station. A strong signal from a 1710-kc transmitter could enter the mixer input (assuming that the input stages had poor selectivity) and beat with the 1255-kc oscillator signal to produce a 455-kc i-f signal with the same audio modulation as the 1710-kc r-f signal. The result would be serious interference and garbled or distorted audio output. The difference between the image frequency and the desired signal is always twice the intermediate frequency. The difference between the 1710-kc image signal and the desired 800-kc station is twice 455 kc. For minimum interaction from image frequencies the input circuit of the mixer should possess fairly high selectivity characteristics, or, preferably, be preceded by a tuned r-f amplifier stage. The ratio of the receiver output voltage produced by the desired signal to that produced by the image signal is termed the "signal-to-image" or "image" ratio. It is important to note at this time that the earlier intermediate frequencies of 175 kc and 262 kc were increased to 455 kc and 465 kc in order to increase the separation between desired and image signals, thus placing the image signal at a lower level of the input circuit resonance curve. The image frequency for the 455-kc intermediate frequency is 1710 kc when the desired station is 800 kc. The output level at 1710 kc on an 800-kc resonant circuit is relatively low. With a 175-kc intermediate frequency, the image signal would be 800 kc plus twice the i.f. (350 kc) or a total of 1150 kc. The output level of the tuned 800-kc input circuit would be substantially higher for the 1150-kc image, and increased interference would result. At higher frequencies or in short wave reception, it becomes more difficult to obtain a satisfactory signal-to-image ratio, because a 455-kc intermediate frequency does not offer appreciable spacing between the incoming rf (in the order of 5000 to 20,000 kc) and the local oscillator. Therefore, in communications receiver design higher intermediate frequencies, such as 1600 kc and 5000 kc, are sometimes used.

8. Additional Sources of Interference

In addition to "image" frequency problems, other undesired signals may enter or be internally developed in the set, producing the specified i-f signal. At the audio detector stage, harmonics are produced during rectification. These harmonics may be coupled back to the mixer stage and heterodyne with the local oscillator. It is also possible that harmonics generated in the local oscillator stage may combine with signals other than the one desired and produce output at the intermediate frequency. To minimize these various interference effects, a receiver must be carefully designed so that internal generation of harmonics is reduced to the lowest possible level. Adequate shielding and proper component layout are of prime importance in realizing maximum circuit isolation with minimum interaction.

9. Double Conversion

The stability, selectivity, and high gain of a low intermediate frequency can be obtained without sacrificing image rejection by employing double conversion. The incoming r-f signal is mixed with a first local oscillator to produce a relatively high intermediate-frequency signal to minimize image interference. Then this first i-f signal is mixed with a second local oscillator signal to produce an i-f signal of lower frequency.

Some of the more elaborate communications receivers employ double conversion, using a *crystal-controlled* first local oscillator, as illustrated in Fig. 4 (A). The r-f and first i-f tuned circuits and the second oscillator frequency are all varied at the same time for tuning. The crystal-controlled first oscillator remains at fixed frequency; its stability is important because tunable oscillators can contribute to receiver frequency drift. The second oscillator (operating at a much lower frequency than the first) is tuned in such a manner that, even though the first i-f signal varies in frequency, the second intermediate frequency is fixed.

A superhet converter connected to a superhet receiver constitutes a double conversion receiver, as illustrated in Fig. 4 (B). The converter includes an r-f amplifier, first oscillator, and first mixer. The i-f signal from the converter is tuned in by the receiver to which the converter is connected. In other words, the r-f section of the receiver becomes the first i-f section of the combination.

Of course, the lead from the converter to the receiver, and the receiver's input circuit must be well shielded to prevent signals of the first i-f from being picked up directly by the receiver, causing interference.

10. Conversion Efficiency

The conversion efficiency of a mixer or converter stage is the ratio of the developed i-f output voltage at the mixer plate load to the r-f signal input voltage applied to the grid. Another measure of mixer performance is termed *conversion transconductance.* It is

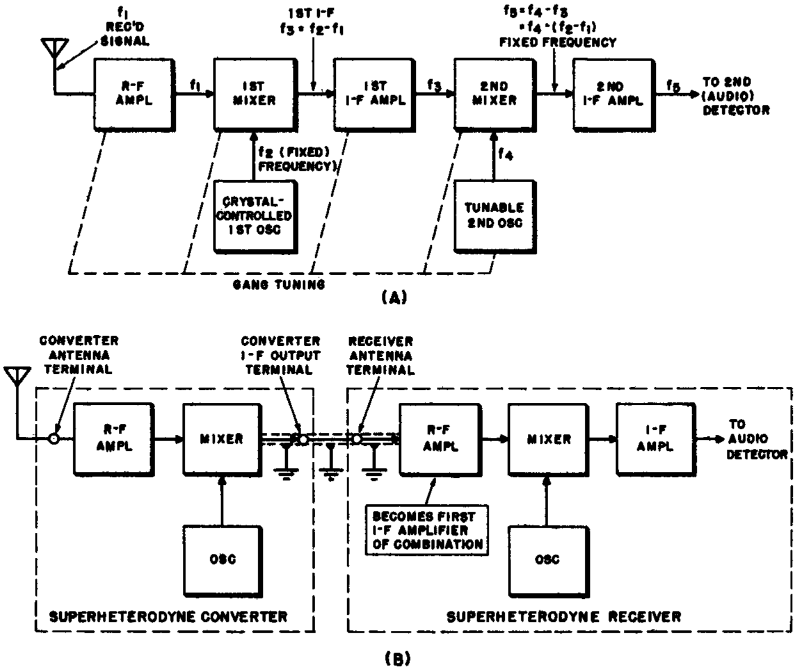

Fig. 4. Double conversion methods. Communications crystal-control type (A) and superheterodyne converter-receiver type (B).

the ratio of i-f current in the mixer output tuned circuit to the incoming r-f voltage required to produce that current. It is desirable to obtain maximum conversion efficiency from a mixer stage in order to achieve best overall receiver sensitivity. To prevent the masking out of a weak signal input, it is important to keep the noise level in the mixer stage as low as possible, especially when preceding r-f stages are not used. In receiver design, optimum conversion efficiency is achieved without depending on an exact oscillator signal amplitude, because it is difficult to maintain a constant oscillator energy output over the entire broadcast band.

11. The Local Oscillator

The local oscillator section of a converter, or a separate local oscillator for use with a mixer stage, must provide a stable, constant voltage output when set to a particular frequency. It must be capable of being tuned, together with the r-f input and mixer tuned circuits, so as to produce a fixed intermediate frequency signal output over the entire broadcast band. In broadcast receiver design, the local oscillator stage is operated at a higher rather than a lower frequency than the incoming r-f signal, in order to reduce the ratio of maximum to minimum oscillator frequency. As an example, consider a 455-kc i-f receiver covering the 550-kc to 1750-kc broadcast band. With the choice of operating the oscillator at a frequency lower than the incoming rf, the necessary oscillator frequency range would be 95 kc to 1295 kc, roughly a 13-to-1 ratio. Choosing to operate the receiver with the local oscillator above the incoming rf, the range becomes 1005 kc to 2205 kc, giving a 2-to-1 ratio. This is obviously much more readily obtainable with a variable tuning capacitor than the 13-to-1 ratio. However, there are situations in which it is desirable to operate the local oscillator at a frequency below that of the incoming rf. For high-frequency reception, the ratio of oscillator tuning coverage is similar in either method and the use of the lower oscillator frequency range permits higher oscillator output and reduces the percentage of the frequency drift that is common at high frequencies. The important factors in a properly-designed local oscillator stage include frequency stability (freedom from drift), absence of harmonics that might produce interference, relatively constant output amplitude over its tuning range, and a minimum of radiation to prevent interaction with nearby sets.

12. Oscillator Tracking

Perhaps the most critical demand of the oscillator stage in a superhet receiver is the necessity for correct *tracking* over the entire band. Since the i-f stages are fixed-tuned at the intermediate frequency, the local oscillator signal at any spot on the dial must beat with the incoming r-f signal to produce a signal of this intermediate frequency.

Since the local oscillator and r-f input circuits resonate at different frequencies, provisions must be incorporated either in the structure of the variable capacitor assembly or in the oscillator circuit to obtain proper tracking. Most modern receivers are built with a "cut plate" oscillator tuning capacitor construction, in which the rotor plates are smaller in size and of different shape than those of the variable capacitor in the r-f tuning section. (See Fig. 5.) Automatic tracking is obtained by shaping the plates carefully and by using properly matched inductance values for the oscillator and r-f coils. The 455-kc spacing between the tuned circuit resonant frequencies is thus kept constant throughout the tuning range of the receiver. The dial must, of course, be adjusted to cover the proper range, but the important thing is that difference between the in-

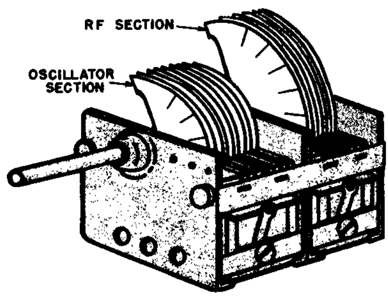

Fig. 5. Tuning capacitor with cut plate oscillator section.

coming radio frequency and local oscillator frequency must equal the intermediate frequency, and that the mixer input and r-f amplifiers (if any) must tune simultaneously to the received r-f signal over the entire tuning range. Only then is the receiver said to be tracking properly.

In some receivers a variable capacitor, called a "series padder" or simply "padder," is inserted in the local oscillator stage to help in achieving proper tracking. As shown in Fig. 6, a typical oscillator tuning circuit consists of the variable tuning capacitor, C_V, shunted by the trimmer capacitor, C_T. The padder capacitor, C_P, is in series with the above combination, and the entire capacitor section is connected across the oscillator coil inductance, L.

It is necessary to recall that wiring capacitors in parallel gives a capacitance equal to the sum of the individual capacitances, while connection in series gives a total capacitance smaller than that of the smallest capacitor. Consequently, the capacitance of the combination of a high capacitance in series with a much lower capacitance is determined almost completely by the value of the lower capacitance.

At the high end of the broadcast band, the variable tuning capacitor (approximately 7 $\mu\mu$f to 115 $\mu\mu$f) is fully open or set at its minimum capacitance. The total capacitance across coil L is low because the relatively higher-capacitance padder C_P (approximately 100 $\mu\mu$f to 360 $\mu\mu$f) has little effect on the series circuit. The high end of the oscillator range may be adjusted by the trimmer capacitor, C_T, which is low in value, but is in parallel with the minimum variable tuning capacitance. When the variable tuning capacitor is fully closed, at maximum capacitance setting, the shunting effect of trimmer C_T is negligible, and the oscillator is operating at its low frequency limit. Padder capacitor C_P is now in series with a larger value of varying tuning capacitance and will have a substantial effect on the resonant frequency of the circuit. The padder capacitor adjustment therefore determines the low frequency oscillator setting and, in a-m broadcast receivers, is usually aligned at the 600-kc point on the receiver dial corresponding to an oscillator frequency of 1055 kc on 455-kc i-f sets. The trimmer capacitor, C_T, is generally adjusted at the highest frequency desired on the band (1650 kc to 1720 kc) depending on the design of the particular receiver. The above tracking methods are necessary because the r-f tuned circuits must operate from 550 kc to approximately 1700 kc, while the local oscillator must cover the range from 1055 kc to 2155 kc (assuming a 455-kc i.f.). As previously indicated, it is of utmost importance

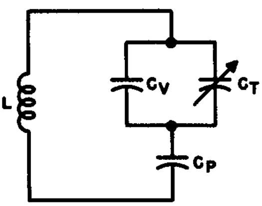

Fig. 6. Simplified schematic diagram of an oscillator-tuned circuit.

to maintain the correct difference frequency at all points of the receiver's tuning range for optimum performance.

13. Oscillator Pulling

A change in oscillator frequency caused by tuning the mixer input is termed "pulling." This is a condition in which the incoming r-f signal forces a change in the frequency of the local oscillator output. When "pulling" is present, the oscillator tends to operate at the signal frequency or nearer to it, even though the oscillator is never tuned to the signal frequency. Even if the oscillator frequency moves slightly toward the received signal frequency, there can be no proper i-f output since the signal frequency and the oscillator frequency would no longer be properly spaced. The greater the difference in frequency between the r-f and oscillator signals (the higher the i.f.), the smaller the tendency there is toward "pulling." The spacing between the r-f and oscillator signal frequencies is comparatively small at very high frequencies, even when the i.f. is 455 kc, and pulling effects must be minimized by proper mixer tube selection and circuit isolation to avoid the necessity of using an even higher intermediate frequency.

14. Oscillator Squegging

When the feedback of an oscillator circuit becomes excessive, or develops too great a time constant, it may cause "squegging" (the generation of several frequencies at the same time). These undesired signals may combine with r-f signals and produce interference effects at the receiver output. Insufficient feedback in the oscillator stage, on the other hand, results in low output or, in some cases, absence of output, causing a total loss of i-f signal.

15. Review Questions

(1) Briefly outline the operation of a trf receiver. Include a block diagram.
(2) Briefly outline the operation of a superhet receiver. Show a block diagram.
(3) What is the difference between a mixer and a converter?
(4) Name two advantages of a superhet over a trf receiver.
(5) What is meant by arithmetic selectivity?
(6) What is meant by image frequency interference?
(7) A 455-kc i-f set is tuned to 1160 kc. What is the frequency at which image interference can be caused?
(8) How can image frequency interference be minimized?
(9) What is conversion efficiency? Conversion transconductance?
(10) Describe three important requirements of a well-designed local oscillator stage.
(11) What is meant by oscillator pulling? Oscillator squegging?

Chapter 2

EARLY TYPES OF CONVERTERS AND MIXERS

16. Autodyne

One of the earliest types of converters developed for use with superhet receivers is the autodyne, shown in simplified schematic form in Fig. 7.

Tuned circuit *L1-C1* at the autodyne input is tuned to the required local oscillator frequency required for development of the desired intermediate frequency rather than to the incoming r-f signal. The tuned plate load circuit is adjusted for parallel resonance (high impedance) at the intermediate frequency, and thus offers low impedance to the oscillator and signal frequency currents. The r-f signal is coupled to the autodyne stage by coupling coils *L4* and *L1*. The circuit operates as a local oscillator (tuned grid type) with coil *L3* supplying the necessary feedback energy. The operating bias is developed by grid components *R* and *C*. The heterodyne action between the incoming r-f signal and the generated oscillator signal results in an i-f signal current flowing through the high impedance plate load, *L2-C2*, thereby developing a high i-f voltage for application to the succeeding i-f amplifier stages. Since the autodyne input circuit is tuned to the oscillator frequency, its resonant peak occurs at this frequency, and its response drops off at any other value. Consequently, the incoming r-f signal is applied to a circuit that is tuned to a different frequency and maximum output cannot be attained. The dropping-off in response level is determined by the difference in frequency between the oscillator (to which the circuit is tuned) and the incoming signal frequency (the intermediate frequency). It is therefore apparent that for use with earlier 175-kc i-f systems, the pulling action of the autodyne was fairly serious. On the other hand, the decrease in output at the higher i-f values, 455 kc and 465 kc, resulted in a marked drop in sensitivity. Because of these dis-

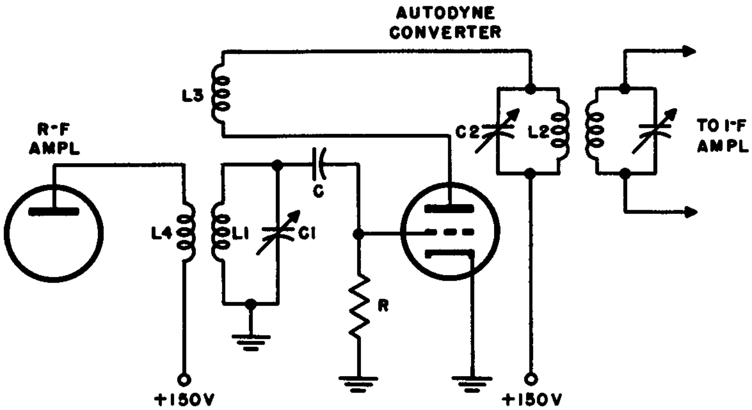

Fig. 7. Simplified autodyne circuit.

advantages, the autodyne was soon replaced by multigrid converter tubes, which combine mixer and oscillator functions, while maintaining a fairly high freedom from interaction.

17. Types of Converters

Originally, the converter in a superhet receiver consisted of one oscillator tube and one mixer tube. Many present-day receivers, especially those used for vhf reception, still employ two-tube converters. However, in most a-m broadcast receivers and in some other types, the entire conversion function is accomplished in one tube, with some of the tube's elements acting as the oscillator, others as the mixer, and some serving both functions. This type of tube is known as a *pentagrid converter*. Examples are the 6A8G, 6SA7, 6BE6, and 1R5. Because the pentagrid is the most common type of converter, its circuits will be discussed first.

18. Pentagrid Converters

As an example of pentagrid converters, we shall consider the 6A8G. The internal construction of the 6A8G, shown in Fig. 8, consists of a cathode, a plate, and five grid elements. The inner grid (No. 1) acts as the control grid of the oscillator section, with grid No. 2 serving as the oscillator plate. (Grid No. 2 is constructed in

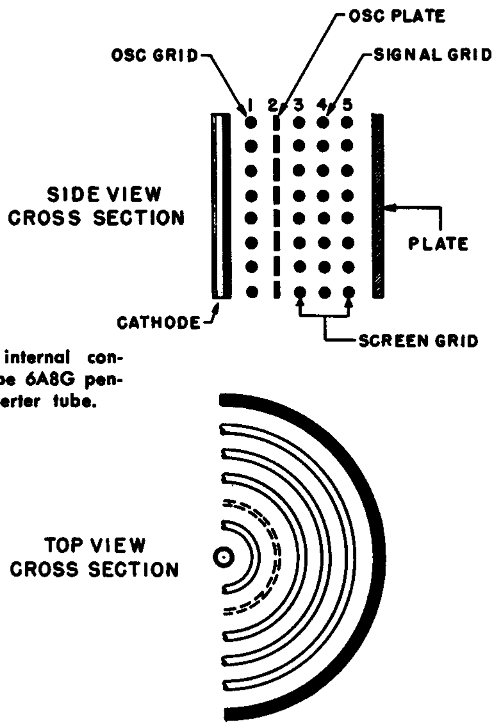

Fig. 8. The internal construction of type 6A8G pentagrid converter tube.

the form of two vertical rods rather than as a complete grid envelope around the cathode.) Grids 3 and 5 act as shields around grid No. 4, the input signal electrode, to isolate the oscillator and r-f voltages, and thus minimize pulling and other detrimental effects. A typical circuit employing a tuned grid oscillator is shown in Fig. 9.

Components *R2* and *C2* form the oscillator grid bias network, while *R1* and *C1* are selected so as to obtain correct mixer bias for operation on the desired non-linear portion of the tube characteristic curve. The oscillator is a tuned-grid type with grid No. 2 acting as the plate; grid No. 2 voltage is obtained through decoupling network *R5C5*. The decoupling network is required to prevent the oscillator current from entering another stage through the common power supply and interacting to produce interference effects. The decoupling action may be illustrated simply by following the flow of oscillator current through grid No. 2 and the feedback coil, to the junction of *R5* and *C5*. At the oscillator frequency, the reactance or opposition to current flow offered by *C5* is considerably lower

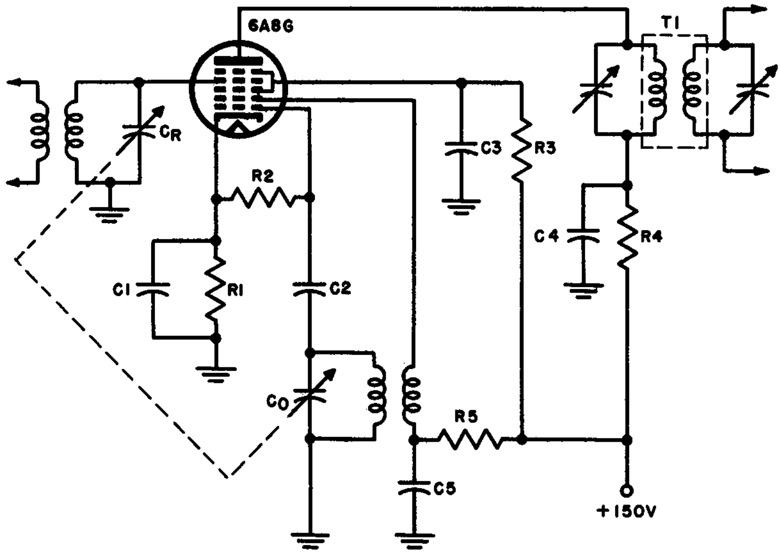

Fig. 9. A converter circuit using a type 6A8G tube.

than the resistance of *R5;* the oscillator current is thus bypassed to ground and is prevented from entering the common power supply. Screen grids Nos. 3 and 5, as well as the plate circuit, contain similar decoupling filters (*C3R3* and *C4R4*) in order to reduce any possible interaction effects. With the oscillator signal at Grid No. 1 and the r-f signal on grid No. 4 modulating the cathode electron stream, the difference frequency is developed and applied across the primary of i-f transformer *T1*. The oscillator variable tuning capacitor, C_O, is mechanically ganged to input variable capacitor C_R.

The 6A8G type of pentagrid converter should not be confused with more modern types, such as the 6SA7 and 6BE6, which are discussed in Chap. 3.

19. Mixer Stage (6L7G) and Separate Oscillator (6C5G)

The internal construction of a 6L7G pentagrid *mixer* is shown in figure 10. The 6L7G contains five grids of standard construction

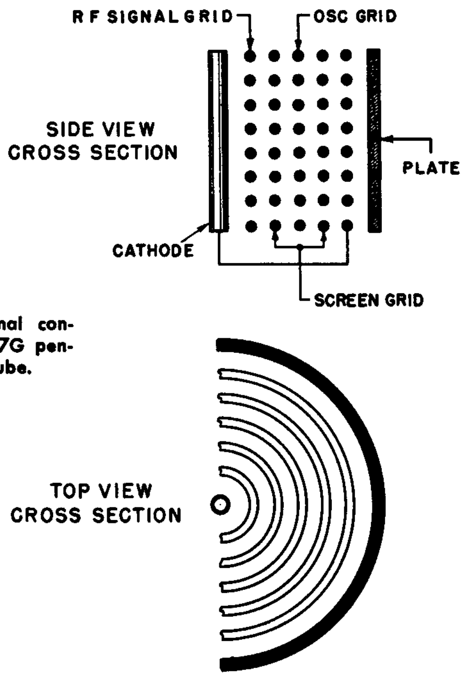

Fig. 10. The internal construction of type 6L7G pentagrid mixer tube.

(compared to the vertical rods of grid No. 2 in the 6A8G tube). The 6L7G, in its use as a mixer with a separate 6C5 G oscillator, is shown in Fig. 11.

This arrangement is an example of inner grid signal injection and outer grid oscillator injection. The 6C5G oscillator is a tuned-grid type with plate supply voltage obtained through decoupling filter *R4C4* and grid leak bias developed across *R6C6*. Oscillator signal is coupled through capacitor *C2* and is applied to grid No. 3, with *R2* acting as a grid return path. Proper mixer operation with maximum conversion efficiency is obtained by properly biasing the mixer tube, 6L7G, by cathode components *R1* and *C1*. The incoming r-f signal is tuned by variable capacitor C_R, ganged to oscillator capacitor C_0 and fed to grid No. 1. The difference frequency current component flows through the primary of *T1* and develops maximum voltage across it for application to the i-f stages. Grids Nos. 2 and 4 provide the shielding between the oscillator and input circuits and

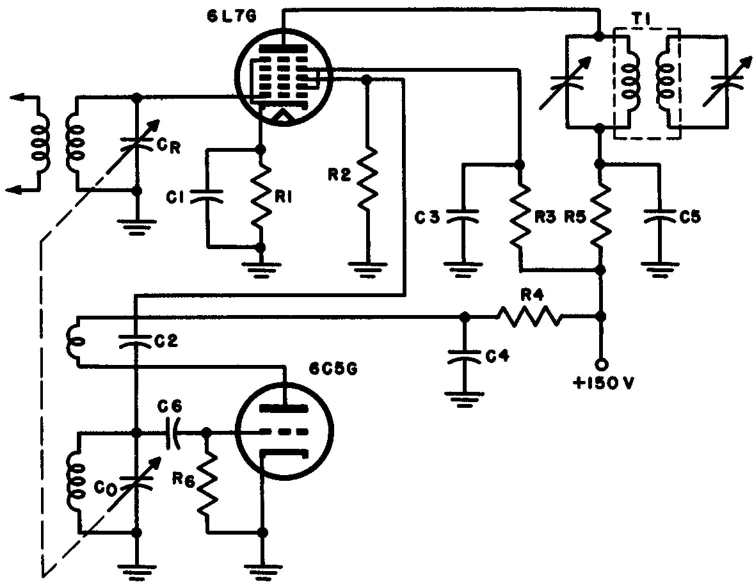

Fig. 11. Circuit using a 6L7G mixer stage and 6C5GT oscillator stage.

are decoupled from the power supply by means of *R3C3*. A suppressor grid, No. 5, is included to increase the plate resistance of the tube and reduce secondary emission effects. Grid No. 1 has a remote cutoff characteristic, and may thus be controlled by avc bias. Grid No. 3 has a sharp cutoff characteristic and produces a large change in plate current with only a small amount of oscillator energy. Grids No. 2 and No. 4 are internally connected and are supplied by a positive voltage through *R3C3* so as to accelerate the electron stream and provide shielding between the signal and oscillator grids.

20. Triode-Heptode Converter (6J8G)

A more complicated tube structure, shown in Fig. 12, is incorporated in the 6J8G vacuum tube. In addition to the five grids, a small triode section is included in the lower portion of the tube envelope

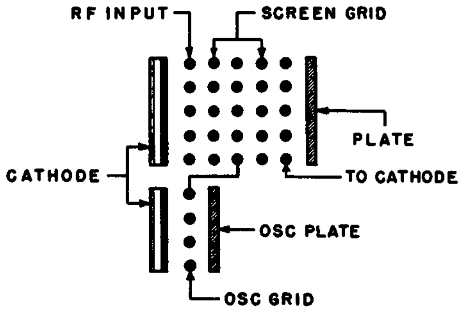

Fig. 12. Internal construction of type 6J8G triode-heptode converter tube.

surrounding the common cathode. The triode grid is used as the oscillator grid and is internally connected to grid No. 3. Figure 13 is a typical 6J8G circuit with oscillator and incoming r-f signal mixing taking place in the common tube envelope.

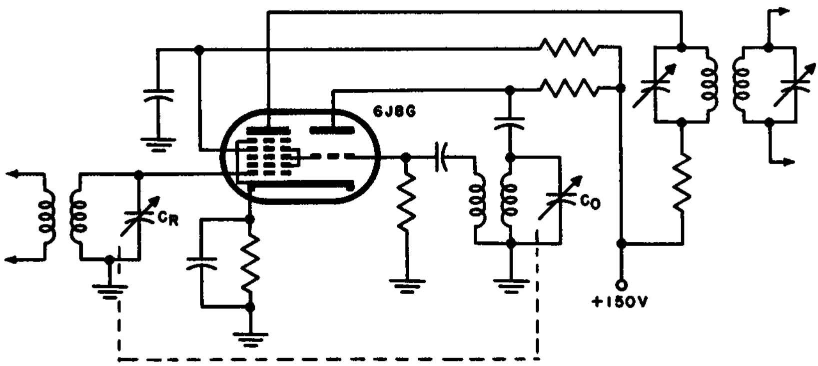

Fig. 13. A converter circuit using a type 6J8G tube.

The circuit description is similar to that of the 6L7G mixer except for the use of a tuned plate oscillator for the triode section to provide increased oscillator output. The bias and decoupling networks are like those in preceding circuit arrangements.

21. Triode Hexode Converter (6K8G)

In the construction of this tube, shown in Fig. 14, the plate electrode of the oscillator section (triode unit) is completely isolated from the mixer section. The first grid (grid No. 1) of the tube

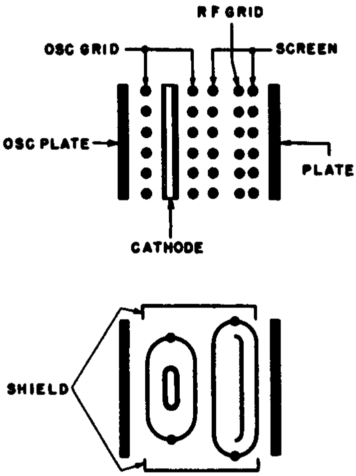

Fig. 14. Internal construction of type 6K8G triode-hexode converter tube.

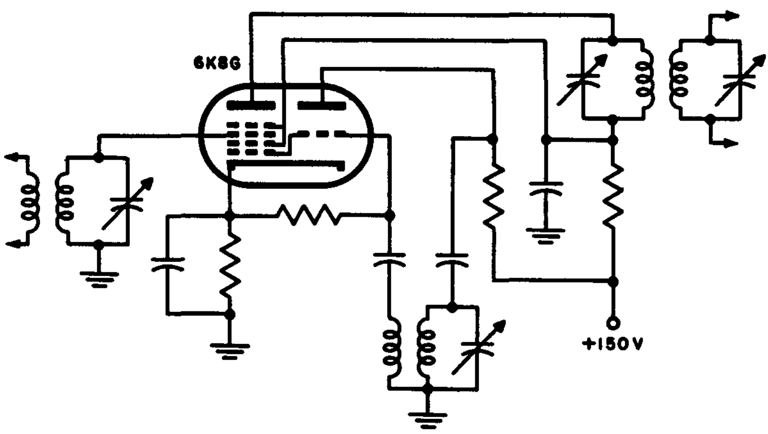

Fig. 15. A converter stage utilizing a type 6K8G tube.

encircles the cathode with the side facing the oscillator plate acting as the control grid of the oscillator while the side facing the mixer varies the electron stream of the mixer section at the oscillator frequency. Internal shields isolate the mixer and oscillator sections and reduce the tendency toward pulling at high frequencies. Thus the oscillator energy is generated by the triode section, transferred to the mixer section by their common grid No. 1, and then mixed with the incoming r-f signal applied to grid No. 3. The plate load of the hexode tube consists of the customary i-f transformer primary. A circuit diagram of a 6K8G triode-hexode stage is shown in Fig. 15.

22. Review Questions

(1) Draw a circuit diagram of an autodyne converter.
(2) What is an important disadvantage of the autodyne?
(3) Draw a circuit diagram of a converter section with separate mixer and oscillator tubes.
(4) What is meant by a pentagrid converter? Describe in detail its functioning.
(5) What is a triode-hexode tube? Name one type.
(6) What is a triode-heptode tube? Name one type.
(7) Sketch the internal construction of a triode-hexode.
(8) Sketch the internal construction of a triode-heptode.

Chapter 3

MODERN CONVERTERS AND MIXERS

23. Pentagrid Converters (6SA7)

The earlier type pentagrid converters operate efficiently at broadcast and slightly higher frequencies (up to about 6 mc) after which they became critical and unstable. The main difficulties encountered at high frequencies are decreased oscillator output and interaction between the r-f input and oscillator signals. To overcome these problems, pentagrid converter tubes, such as the 6SA7 (shown in Fig. 16) were designed so that two electrodes rather than one serve as the oscillator anode. From the circuit schematic shown in Fig. 17, the individual grid functions can be noted.

Grid No. 1 serves as the oscillator grid in a Hartley type circuit with grids Nos. 2 and 4 which are internally connected, acting both as the oscillator plate and as a shield around signal input grid No. 3. Grid No. 5 is internally connected to the cathode and acts as the suppressor grid. Due to the shielding effect about grid No. 3, variations in r-f signal input or applied bias have very little effect on oscillator frequency. Stability and freedom from interaction are also improved by the side rods of grid No. 3; these rods are biased at a negative potential by fixed bias or avc voltage. A miniature version of the 6SA7 is the 6BE6, which is similar in internal construction but has a higher oscillator transconductance factor, due to the special formation of grid No. 1. These tubes are used for high-frequency operation. (The 6BE6 is frequently employed as a converter for f-m receivers operating at 88 to 108 mc.)

24. Pentagrid Converter (1R5)

A popular type converter found in most battery operated receivers is the 1R5. The internal construction of this miniature tube is

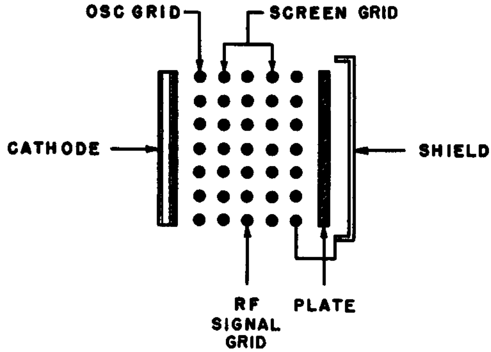

Fig. 16. Cross sectional view of the internal construction of a 6SA7 pentagrid converter tube.

similar to the 6BE6 with the exception of the shields on grid No. 2 and the use of a directly heated filament rather than a heater and cathode arrangement. Circuit operation is similar to that of the 6BE6, except for the use of batteries as plate and filament supplies.

25. Modern Mixer Circuits

Improved pentagrid converters, such as the 6BE6, give satisfactory performance up to about 100 mc. However, even such improved pentagrid converter tubes possess undesirable coupling between the oscillator and r-f input grid sections. This coupling becomes more and more disturbing as frequency is increased. The effective capacitance between the two grids is attributed to their interelectrode capacitance, which is relatively low, and to a *space charge* effect. Assuming grid No. 1 is the oscillator grid, it modulates the cathode current at the oscillator frequency, thereby varying the space charge region around the signal input grid at the same rate. The effect produced is that of a small capacitor connected between the grids; this factor becomes prominent at high frequencies. For this reason, it is necessary to employ separate oscillator and mixer tubes in many communication, f-m and tv receivers, which operate in the vhf (very high frequency) band. The local oscillator may be a Colpitts, Hartley, or tuned plate or grid type; stability and adequate output are its prime requisites. The oscillator output signal is generally connected directly to the mixer grid, while the oscillator energy is inductively or capacitively coupled into the mixer for heterodyning action. Several variations are shown in Fig. 18.

The circuit at (A) employs two triodes. The oscillator is of the "grounded-plate" Hartley type. The output of the oscillator is passed from the top of the oscillator coil through a low capacitance

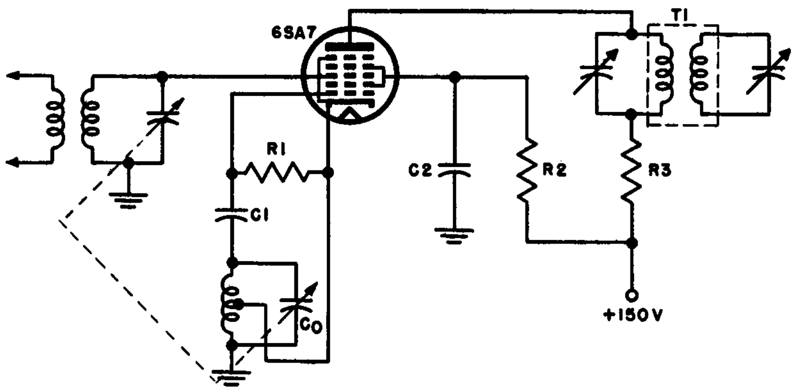

Fig. 17. A typical circuit employing a 6SA7 pentagrid converter tube.

(usually between 0.5 $\mu\mu$f and 10 $\mu\mu$f) to the grid of the mixer which in turn is connected to r-f tuned circuit *L1C1*, from which it obtains the r-f signal. The r-f and oscillator signals mix in the triode mixer. They control the plate current in the non-linear portion of the tube characteristic, due to the cathode bias from *R1C2*. The resulting i-f signal is built up across the primary winding of resonant i-f transformer *T1*.

The circuit at (B) employs a triode oscillator and a pentode mixer. Here the oscillator is of the "tuned-grid" type. The oscillator signal is coupled to the triode mixer by the proximity of tickler coil *L3* to signal coil *L1*. Oscillator voltage is induced in *L1* by mutual inductance *M* between these coils. R-f signal voltage is also induced in *L1* from the antenna or from the preceding r-f amplifier stage. The result is that the oscillator and r-f signals are applied to grid No. 1 of the mixer. *R1* and *C3* keep the mixer properly

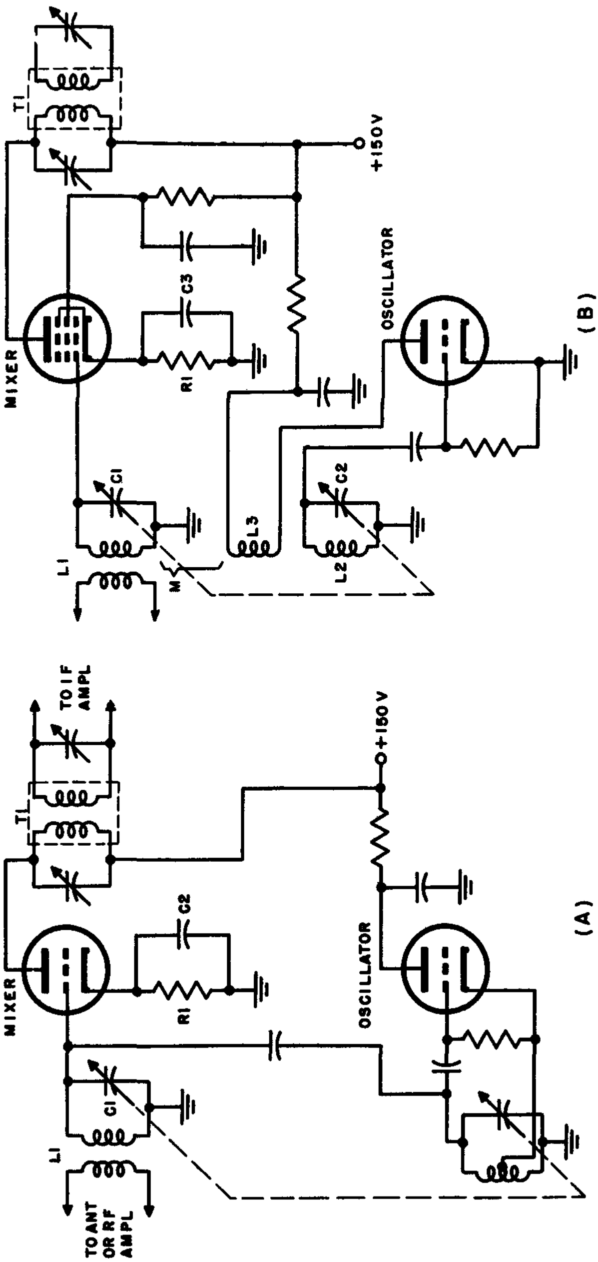

Fig. 18. Separate mixer-oscillator circuit.

biased, so that the plate circuit contains i-f current. Resonant transformer *T1* builds up the i-f voltage for application to the i-f amplifier.

26. Multiband Receiver Mixer Stage

The main difference between multiband and ordinary broadcast receivers exists in the switching arrangement of coils in the tuned r-f and oscillator stages. As shown in the simplified circuit (Fig. 19),

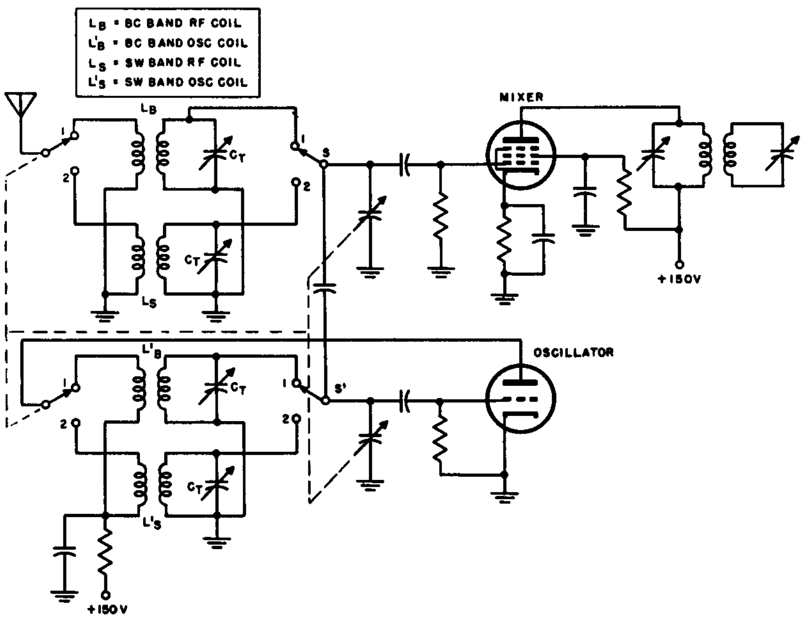

Fig. 19. A multiband receiver converter section.

a two-section (ganged) variable tuning capacitor is connected across the mixer and oscillator tuned circuits. When band switch *S* is in position 1, the broadcast band oscillator coil (L'_B) with its associated trimmer capacitor and the broadcast band antenna coil (L_B) form the remaining tuned circuit components. When switched to position 2, the short wave coils (L_S and L'_S) are inserted and the broadcast coils removed. In addition to the simplified functions shown, the band switch might include additional contacts to (*a*) short out the unused coils to prevent interaction or energy absorption; (*b*) change the input selectivity by resistively loading the input tuned circuit; or (*c*) switch pilot lights on the dial scale for band identification. In spite of its complicated appearance, a multiband receiver is quite conventional in its i-f amplifier, audio detector, and audio output stages.

27. Noise in Converters

One of the most important factors in the choice of a converter circuit or tube is the amount of *random noise* it contributes.

Random noise is generated by thermal molecular action in a resistance, and by random motion of electrons in the electron stream of a tube. At present we are particularly concerned with tubes. The simplest tubes generate the least noise. The more grids a tube has, the more noise it generates. Thus diodes generate the least noise and pentagrid converters are the noisiest tubes of all.

The noise level of the entire receiver is determined in its input circuits. If the first stage in the receiver has a low noise contribution, it can raise the signal so far above noise level that no following stage will have sufficient noise to interfere. It is the job of a good r-f amplifier stage to establish a high signal-to-noise ratio.

Noise increases at high frequencies and with greater receiver bandwidths and is thus an important factor in tv receivers. Under such conditions, a pentagrid converter may contribute enough noise to offset even a rather good signal-to-noise ratio established by a preceding r-f amplifier stage. For this reason triode mixers are popular for tv and other vhf, uhf, and broad band receivers. Although the gain of a triode mixer (conversion transconductance) is ordinarily lower than that of a pentode or a pentagrid converter, the noise added by the latter may more than nullify the gain advantage. A special converter like the 6BE6 is useful for f-m broadcast frequencies only because of the great care taken in design to minimize interelectrode capacitances and to obtain the greatest possible transconductance. Such improvements allow significant increase in conversion transconductance, partially overcoming the noise problem. Even so, virtually all tv receivers use separate tubes for mixer and oscillator (although they are frequently in the same envelope, as in the 12AT7, 6J6, and 6U8).

28. Review Questions

(1) Sketch the internal construction of a 6SA7 pentagrid converter.
(2) Draw a circuit diagram of a 6SA7 pentagrid converter stage.
(3) Name a battery type pentagrid converter tube.
(4) Name two factors limiting the use of a pentagrid converter at very high frequencies.
(5) Name the miniature tube equivalent of the 6SA7.
(6) Why are triodes often used as mixers in high-frequency receivers even though their gain is relatively low?

Chapter 4

I-F AMPLIFIERS

29. Function of the I-F Amplifier

The function of the intermediate frequency amplifier section in the superhet receiver is to select and amplify the difference frequency signal developed at the converter output. The *selectivity* or band-pass characteristics of this section determine, in part, the freedom from interference, as well as the fidelity of the receiver.

30. Selectivity

The selectivity of a receiver is its ability to accept signals within a desired band of frequencies while rejecting others. Tuned circuits and, therefore, tuned stages possess frequency response or selectivity characteristics. A parallel resonant circuit, such as that found in the plate circuit of a converter or mixer, offers maximum impedance at its resonant frequency and decreased impedance as frequency is varied from this value. When fed by a constant current source, such as the pentode tube used in i-f amplifiers, the output voltage developed across the tuned circuit will increase as the impedance of the tuned circuit is increased, by varying signal frequency through resonance. If the tuned circuit has a high Q or a sharp response as shown in Fig. 20 (A), the output voltage will drop sharply as the frequency is varied slightly off resonance. A lower Q or a more broadly tuned circuit (Fig. 20B) produces less output at resonance, but the response decreases more gradually at frequencies off resonance. In radio terminology, the bandwidth of a tuned circuit or stage is often designated as that band of frequencies in which response curve shows an output voltage of 70.7 percent of the peak (resonant) level or higher. Thus the sharply tuned circuit of the Fig. 20 (A) characteristic would represent a bandwidth of 4 kc with rapid attenuation of other frequencies. Figure 20 (B) characterizes a circuit with a 10-kc bandwidth. An overcoupled i-f bandpass

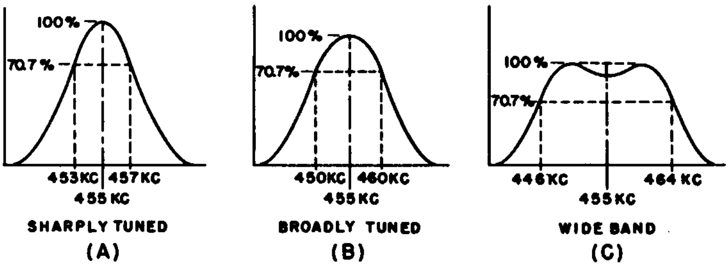

Fig. 20. The response characteristics of tuned circuits with different degrees of Q and coupling.

characteristic, encountered in high-fidelity tuners, is shown in Fig. 20 (C). The bandpass is 18 kc wide for full sideband reception. A decrease in output voltage from a 100 percent level to a 70.7 percent level is equivalent to a 50 percent reduction in power, since power is a function of the square of voltage ($P=E^2/R$). The bandpass characteristics may thus be given in terms of 50 percent power levels as compared to the peak level. A third method of locating bandpass from a response curve is employed by engineers and technicians, and uses "3 db down" points. Reference to db (decibel) tables will disclose that a power reduction of 50 percent is equivalent to a drop of 3 db. Therefore, the 70.7 percent voltage points locating bandpass extremes show the 50 percent, "half-power" or "3 db down" level.

31. Bandpass Requirements in Superhet I-F Stages

In a-m transmission, the audio information is applied to the r-f carrier and is transmitted in the form of sidebands. For example, a 5000-cycle audio signal applied to a 600-kc station carrier results in the transmission of 595-kc, 600-kc, and 605-kc signals; the 595-kc and 605-kc components are termed the *sideband signals* and contain the audio power. During the mixing action in the superhet, 450-kc, 455-kc, and 460-kc signals are produced. If the audio information is to arrive at the audio detector for demodulation, it is necessary that the i-f amplifier allow passage of this band (450 to 460 kc) frequencies. If the plate load (the i-f transformer primary) of the converter stage were a sharply tuned circuit, as shown in Fig. 20 (A), the output at 450 kc and at 460 kc would be quite low and the 5-kc sideband components at those frequencies would be considerably attenuated. The net result, of course, would be a relatively weak 5-kc signal at the receiver loudspeaker. On the other hand, a tuned stage response, as shown in Fig. 20 (B), would produce a relatively high output for the 5-kc sidebands, and would afford satisfactory reception.

Voice frequencies generally do not exceed 8 kc, but musical instruments may produce frequency components above 15 kc. For reception of voice transmission alone, as in short wave ship-to-shore communication, it may be desirable to employ sharply selective (highest *Q*) circuits for maximum gain and minimum noise and interference, since wide band response is not necessary. It has been established by listening tests that the removal of frequency components above 2000 cycles alters the "matureness" of the voice or speech but does not materially affect the intelligibility. For reception of program material in the broadcast band, however, a wider response is required for proper entertainment value. Local a-m broadcast stations in any given area have assigned frequencies spaced at least 40 kc apart. They normally have modulation frequency response up to at least 7500 cps. However, to minimize inter-area interference at night, control noise, and provide low distortion with economical design, the pass band of most a-m broadcast receivers is kept to about 10 kc.

For high-fidelity reception, a-m tuners use several i-f stages, each with low gain and wide bandpass characteristics, to achieve the overall bandpass shown in Fig. 20 (C).

32. Choice of Intermediate Frequency

Many factors were studied over a period of years to establish what is presently accepted as the conventional or standard a-m broadcast intermediate frequency of 455 kc. Early types of a-m receivers utilized 100-kc, 175-kc, and then 262-kc intermediate frequencies. Tubes and i-f transformers were not at the present stage of development in operation, stability, and production. In addition, the choice of a lower intermediate frequency resulted in improved adjacent channel selectivity. Suppose a 1000-kc station and a 1010-kc station are to be separated so that the 1000-kc station is received with minimum interference. Mixing with a local oscillator produces 100-kc and 90-kc i-f signals, assuming that a receiver with a 100-kc intermediate frequency is used. The percentage of spacing is 10 kc out of 100 kc, or 10 percent. For a 455-kc i-f set, the spacing is 10 kc out of 455 kc, or 2.2 percent. Since this percentage of spacing is the criterion of tuner circuit selectivity it is easier to reject the adjacent station (1010 kc in the above example) with a 100-kc i.f. than with a 455-kc system. However, with the improvements in the design of tubes and input tuned circuits, as well as the use of several double-tuned i-f circuits, the adjacent station interference problem has been reduced considerably. By far the more important problem is that of image interference, discussed in Chap. 1. As shown there, the image frequency is spaced at a distance twice the intermediate frequency from the desired station frequency. If the intermediate frequency is raised to 455 kc, the image frequency falls outside of the broadcast band for all received station frequencies above 790 kc. (The image frequency for the 455-kc receiver for a 1000-kc station, 1910 kc, is spaced so far from that of the received signal that the output level from the input tuned circuits would be extremely low.) A 100-kc i-f receiver would have the image interference signal only 200 kc from the desired signal and the output level for the image signal would be quite high. Another disadvantage of a low i-f value is the tendency towards oscillator pulling.

With a high intermediate frequency, gain and stability are reduced, and in some cases local oscillator frequency drift becomes a problem. However, with the present-day pentodes and shielded i-f transformers, the gain and stability problems are overcome; proper component design has also minimized oscillator drift. Extensive field reports of commercial sets, together with laboratory checks, have proven that 455-kc i-f systems are adequate with frequency coverage up to 7 mc, when an r-f stage is not used, and about 20 mc with a tuned r-f amplifier. Multiband radios and most communication receivers use a 455-kc intermediate frequency, while communication receivers covering only frequencies over 2 mc often utilize 1.6-mc i-f stages. In special cases, double superheterodyne receivers (as described in Chap. 1) with 5 mc or 10 mc as the first i.f. and 455 kc as the second i.f. are encountered in wide coverage communication work. Frequency modulated (f-m) receivers use a 10.7-mc i-f system, while tv sets utilize wide bandpass, multi-stage 20-mc and 40-mc i-f amplifiers.

33. I-F Amplifier Circuits

The vast majority of i-f amplifiers found in radios are similar in basic circuitry, even though the specific intermediate frequency, gain, or bandpass characteristics may differ. The ordinary table model a-m radio uses a single-stage 455-kc i-f amplifier with two (one input, *T1*, one output, *T2*) double-tuned i-f transformers, as shown in Fig. 21.

The pentode i-f amplifier tube, *V2*, is a remote cutoff type to permit application of an avc voltage for gain control purposes. In the circuit shown, double-tuned i-f transformers are used for maximum selectivity, although single tuned and even untuned coils may be used for coupling. In general, the greater the number of tuned circuits, the better the selectivity and the lower the interference reception problems.[1] Minimum bias, which adds to avc voltage, is obtained by self bias through cathode components *R1* and *C1*. Class A operation using this bias is required, because any other class operation would result in clipping or distortion, producing undesirable harmonics. Creation of a second or third i-f harmonic (910 kc or 1365 kc) could conceivably result in interference due to heterodyning with some incoming r-f signal. Also, the wave envelope shape is distorted in any single-tube amplifier of class other than A. For these reasons, class A (linear) operation must be maintained in all types of i-f amplifiers. The plate of the mixer stage contains the primary tuned circuit of *T1* (Fig. 21) with mutual coupling transferring the i-f signal to the tuned secondary. Decoupling components *R2* and *C2* in the grid circuit prevent interaction between the various circuits fed by the avc bias line. Due to the relatively low plate and screen voltage in a-c/dc table radios and battery type portables, the tendency towards instability or oscillation is lessened, and the customary plate and screen decoupling filters (shown in preceding diagrams) may be omitted; the signal bypass function is performed by capacitor *C3*. The plate load of the i-f stage is another i-f transformer, double-tuned for improved selectivity,

[1] This is true except in "stagger-tuned" i-f circuits of f-m and tv receivers, in which not all circuits are tuned to the same frequency.

with the secondary feeding the audio detector or demodulator. The i-f transformers are usually metal shield cans containing the coils and associated capacitors. Either air-core or powdered iron-core universal wound coils may be used; iron core types are preferable because higher *Q* and therefore higher gain and selectivity can thus be obtained. Universal winding is more popular than multi-layer winding because distributed capacitance effects between layers of turns are reduced. Some types of i-f transformer use mica-compression or air-dielectric type trimmer capacitors for tuning, while more re-

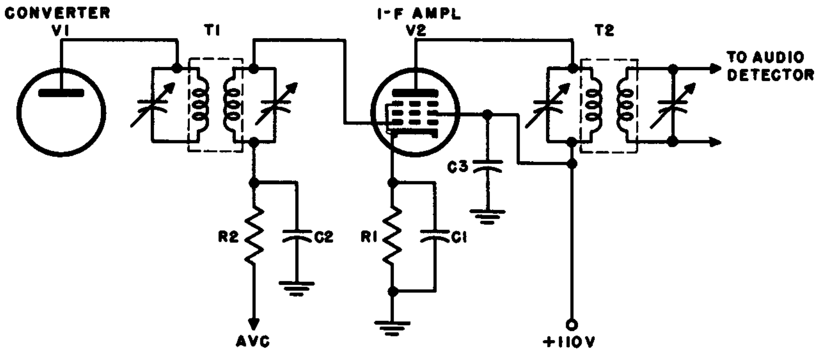

Fig. 21. Circuit diagram of an i-f amplifier.

cent forms employ permeability tuning, in which the inductance of the tuned circuit is varied by changing the position of a powdered-iron slug in the coil.

Common tube types for a-m broadcast receiver i-f stages are the 6SK7 and 6BA6. Table model a-c/d-c sets utilize the above types with 12-volt filaments (12SK7, 12BA6, etc.), while battery portables use similar 1-volt tubes, like the 1T4 and 1U4. As previously mentioned, the i-f tubes used are of the remote cutoff type (sometimes referred to as "variable-mu" or "super-control" types), and are selected for high plate resistance (for least loading effect and therefore maximum selectivity) and high mutual conductance (for maximum gain). For additional gain and selectivity, a receiver may contain two or even more i-f stages, each similar to the amplifier shown in Fig. 21. Additional decoupling and careful component layout are required to prevent oscillation.

34. Automatic Volume Control (AVC)

It is desirable to maintain a fairly constant audio output and prevent overload of the i-f amplifier, regardless of r-f input signal strength or reception conditions. This is done in virtually all modern superheterodynes by automatic volume control (avc). Avc provides bias to the i-f amplifier and converter stages. This bias is a negative d-c voltage, developed at the audio detector stage. Its amplitude is determined by the average signal strength of the incoming signal. The greater the input signal, the more negative is the voltage developed; conversely, a weak input signal produces only a slight negative potential. The avc voltage, in either case, is applied through appropriate decoupling filters to the grids of the remote cutoff i-f, r-f and converter tubes. Thus a receiver tuned to a strong station

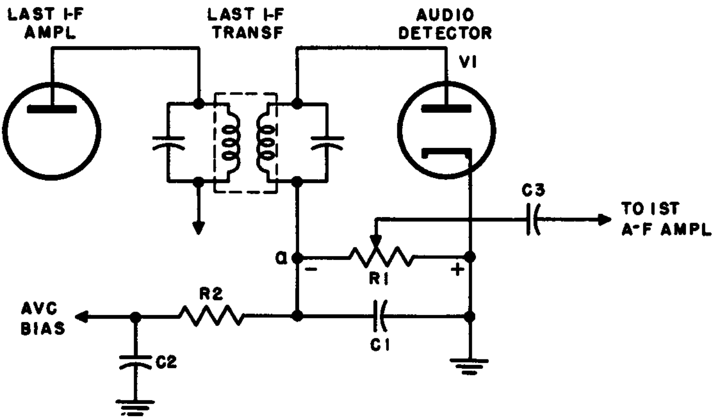

Fig. 22. A simple AVC circuit.

develops a large bias at the control grids of the controlled stages. This large bias reduces the overall gain, so as to produce a certain level of output. As the receiver is tuned to a weaker station, the developed avc bias becomes less negative, permitting higher gain than is allowed with the strong signal. The audio output at the speaker is substantially equal for both signals on a well designed radio.

A simple avc circuit is shown in Fig. 22. Diode section *V1* will conduct on each positive half cycle of the i-f input voltage because its plate will be positive with respect to its cathode. Current through resistor *R1*, the audio detector load, will develop a negative d-c voltage at point *a* with respect to ground; the output across *R1* will be a varying d-c potential whose a-c component corresponds to the audio intelligence sent out from the transmitter. Blocking capacitor *C3* removes the d-c component and feeds the a-c audio signal to the audio amplifier section. Network *R2C2* is a long-time-constant filter (R2=1 megohm and C2=.05 uf, for a .05-second time constant) to smooth out the audio variations and produce an almost pure d-c negative voltage for avc, applied to the remote cutoff i-f and converter stages. Capacitor *C1* acts as an i-f bypass to shunt the i-f carrier component to ground. From the simple circuit, note that a strong input to the diode will result in a large negative avc bias, while a weak signal would produce a correspondingly lower bias potential. The negative avc voltage is applied, through decoupling filters, to the avc-controlled stages.

35. Delayed AVC (DAVC)

In some cases, particularly in fringe areas, it is advantageous to incorporate a modified type of avc known as delayed avc in order to operate the receiver at maximum gain for reception of very weak or moderate signals, while retaining full control over the strong or locally-transmitted signals. In the simple avc system, even weak signals develop some negative bias, which is applied together with the minimum bias at the cathode, to reduce the overall receiver gain. By suitably applying a fixed negative voltage to the avc tube (approximately 2 or 3 volts), it is possible to prevent the application of any avc voltage until a predetermined signal input level is reached. As shown in Fig. 23, a negative 3-volt potential (derived from the power supply) is inserted between the cathode and ground return of the avc diode, *V2*. Since the cathode of *V2* is negative by 3 volts with respect to its plate, diode conduction occurs, thereby applying —3 volts to the avc voltage point. (Diode conduction may be considered analogous to that of a switch: when the diode conducts the switch closes.) Weak signals applied to audio detector diode *V1* develop relatively low d-c voltages across *R1*, the receiver volume control. This d-c voltage is applied effectively in parallel with the delay voltage across *V2*. For weak signals, the *R1* d-c voltage is not sufficient to make the plate of *V2* become more negative than its cathode. Thus, for these weak signals, the bias applied to the i-f and converter stages would be —3 volts. This is the minimum bias required for class A minimum-distortion maximum-gain operation, and no cathode biasing is used. A strong signal develops a large enough negative voltage (say —4 volts) at point *a*. This stops conduction of diode *V2* and removes the —3-volt potential from the avc bias point. With the 3-volt bias out of the circuit, operation is similar to that of the simple avc type, and —4 volts would be

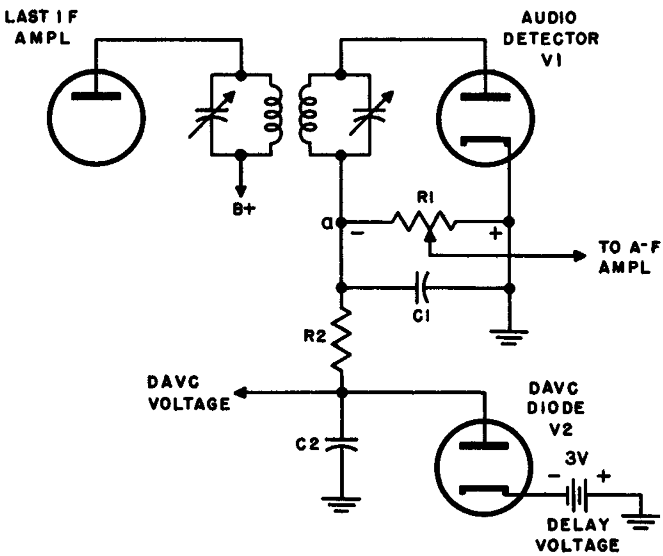

Fig. 23. Circuit of a delayed AVC (DAVC) system.

applied to controlled stages. Thus, for weak signal reception, a fixed bias of —3 volts is applied to the avc line. At a predetermined signal level, the fixed voltage is effectively removed, and is replaced by avc voltage, which is a function of the average signal strength of the incoming signal.

36. Miller Effect

The grid input capacitance of a tube is a function of the grid-to-cathode and grid-to-plate interelectrode capacitances, as well as the gain of the stage. The change of input capacitance with bias is known as the "Miller Effect," which has an important bearing in the design of tuned circuit r-f, i-f, and converter stages, where application of a variable bias, such as avc, is involved. For example, if a given tube has a grid-to-plate capacitance of 1.0 $\mu\mu$f and a grid-to-cathode capacitance of 2.0 $\mu\mu$f, its input capacitance would be determined by the simple relationship:

$$C_{in} = C_{gf} + (M+1)\, C_{gp}$$

where M=stage gain
C_{gf}=grid-filament or grid-cathode capacitance
C_{gp}=grid-plate capacitance.

Assuming that the avc bias is —3 volts on a weak signal, the stage gain might be 19; on strong signals the avc bias might be —7 volts, resulting in an overall gain of only 9. With a weak signal input, the input capacitance shunting the input tuned circuit would be

$$C_{in} = 2.0 + (19+1) \times (1.0) = 2.0 + 20.0 = 22 \ \mu\mu\text{f}.$$

For strong signals, the input capacitance of the stage would be

$$C_{in} = 2.0 + (9+1) \times (1.0) = 2.0 + 10.0 = 12 \ \mu\mu\text{f}.$$

Since the input capacitance is not a constant value, but varies with signal strength, resonant circuit detuning will occur as avc bias is changed. To minimize this effect, tubes with low interelectrode capacitances are selected for r-f and i-f circuitry. In addition, the Miller effect is greatly reduced by the negative current feedback obtained in a stage when the cathode bypass capacitor is omitted; some sacrifice in gain must be suffered but in most cases this is not serious. It might also be noted that i-f transformers may use tuning capacitors in the order of 100 to 200 $\mu\mu$f, so that a change of 5 or 10 $\mu\mu$f due to the Miller effect will not seriously detune their resonant circuits.

Since the detuning effect can be minimized but not completely eliminated, it is important to determine whether alignment should be performed on weak or strong signals. Strong signals can tolerate a partially detuned input circuit because they have a sufficient level to permit some loss; weak signals, unfortunately, may be lost entirely if not received at the peak level of the input tuned circuit response. It is therefore necessary to align receivers with low signal input. Manufacturers' recommendations usually specify application of the minimum signal generator input necessary for lowest range output meter indication. Failure to observe this common precaution may result in inferior receiver performance during reception of weak or fringe area signals.

37. Variable Selectivity

For reception of most broadcast signals, an i-f bandpass of 7.5 kc on either side of the i-f carrier is sufficient, since most stations do not often transmit program material having higher sideband components. Also, sharp tuning is desirable at night, to eliminate interference from adjacent-channel stations in other areas. However, it was previously mentioned that certain stations are licensed to transmit up to 10 kc or more for high-fidelity performance. If receiving conditions are such that full advantage of these sidebands can be realized, it is desirable to incorporate provisions in a receiver to allow variation of the i-f bandpass characteristic. An increase in bandpass, or reduction of selectivity, is accompanied by a loss of

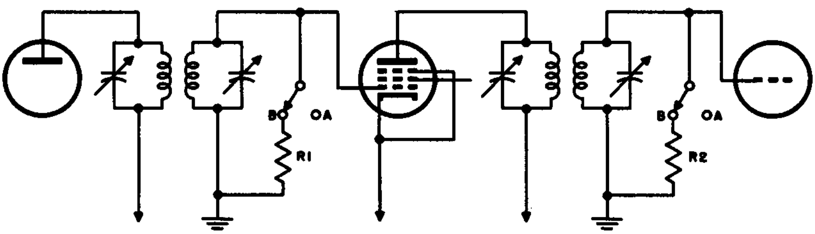

Fig. 24. Variable selectivity by shunt resistor loading.

stage gain; consequently, variable selectivity i-f amplifier stages usually contain two or more i-f stages rather than one. Perhaps the simplest method of increasing bandpass is to shunt one or more of the tuned i-f circuits with resistors; this serves to broaden the resonance curve, decrease the peak value, and allow a wider band of frequencies to pass without appreciable loss. As shown in Fig. 24, when the "variable selectivity" switch is in the A or NARROW or STANDARD position, the tuned circuits are not shunted, and therefore pass sidebands of ± 5 kc. In the B or HIGH-FIDELITY position, shunting resistors *R1* and *R2* are applied across the i-f transformer secondaries, and the circuit passes a sideband of ± 10 kc (20-kc total).

Another common method of obtaining variable selectivity makes use of a tertiary (or third) winding, located between a loosely coupled primary and secondary winding of the i-f transformer. The tertiary winding absorbs energy and thus tends to load the stage, thereby widening the response characteristic. Several types of receivers employ mechanical devices to permit a change in physical spacing between primary and secondary windings, thereby altering coupling and selectivity characteristics. As the coils are brought closer together, up to a certain point (called critical coupling) the gain is increased and selectivity is broadened. At critical coupling, output is maximum. As coupling is increased still further (with coil spacing decreased) an overcoupled or "flat top" response is obtained (see Fig. 20C), with lower gain than that previously obtained, but with a wider pass band.

Receivers utilizing highly selective (narrow band) i-f stages (such as voice communication sets) must contain an accurate, stable local oscillator stage. A change in local oscillator frequency of 3 kc results in a 3-kc change in i-f signal frequency. The result is a marked decrease in output and a substantial increase in distortion.

38. Review Questions

(1) What is meant by selectivity?
(2) Draw the response curve of an overcoupled i-f transformer.
(3) Name two advantages of a high intermediate frequency
(4) Draw a schematic diagram of an i-f amplifier stage.
(5) What class operation is desirable for i-f stages? Why?
(6) Name three tube types for i-f amplifier operation.
(7) Describe the operation of a simple avc circuit. Use a circuit diagram.
(8) Following the procedure of question 7, describe a davc circuit.
(9) What is meant by minimum bias?
(10) Describe the Miller effect. Why is it important?
(11) What is the purpose of variable selectivity?
(12) Describe two methods of obtaining variable selectivity.

Chapter 5

ALIGNMENT

In order to achieve optimum receiver performance, it is necessary to align each of the tuned circuits carefully and properly. The end results of a satisfactory alignment job will be maximum sensitivity and selectivity with minimum interference and distortion. The signal generator should be stable during operation, capable of covering the desired range of frequencies, and accurate in its setting; if possible, it is desirable to spot check the signal generator against a crystal oscillator or a standard frequency transmission such as WWV.

The complete alignment instructions for a five-tube a-c/d-c broadcast band receiver, shown in Fig. 25, are given below. The procedure is explained for the i-f amplifier, the local oscillator, and the mixer sections of the converter, in that order. The operations should be carried out in the same order.

39. I-F Amplifier Alignment

a. Turn the receiver, signal generator, and vtvm on, and allow several minutes for warmup.

b. Connect the a-c leads of the vtvm across the speaker voice coil terminals (A and B). An output meter or multimeter set for a-c voltage may be used in place of the vtvm.

c. Connect the high side of the signal generator cable, through a .01 μf capacitor, to the input, grid No. 3, of the 12BE6 converter tube. Set the signal generator to 455 kc, *modulated.* Turn the volume control on the set to maximum.

d. Ground out the oscillator stage by soldering a short length of wire between the stator terminal and the frame of the oscillator tuning capacitor, *C5*. (See Fig. 26, top view of chassis.) (The stage may also be grounded out by inserting the small blade of a knife or a screwdriver between the plates of *C5* in such a way as to make contact between the rotor and stator.)

e. Adjust the signal generator output level until the modulated note is heard faintly at the speaker and a reading is observed on the output meter (set at its lowest range). It is again important to stress that the minimum signal generator output necessary for meter deflection should be used; the use of an excessively strong signal

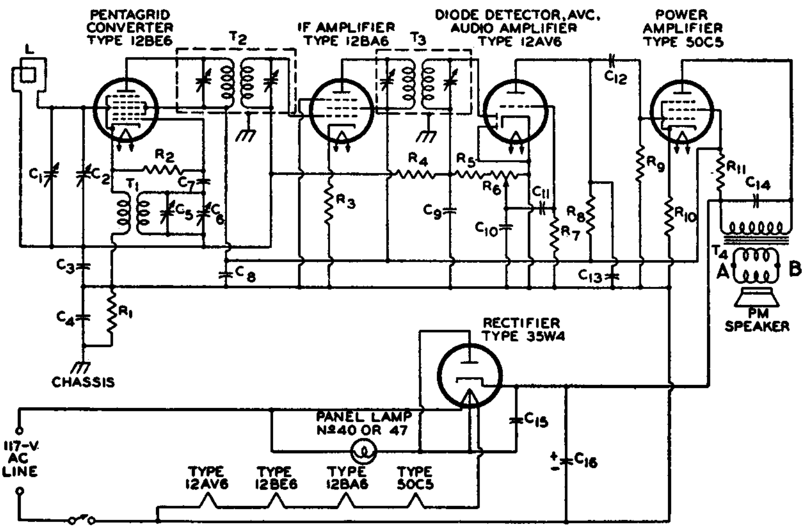

Fig. 25. Schematic diagram of a five tube a.c.-d.c. superheterodyne receiver (Courtesy RCA).

input to the receiver is to be avoided. Adjust the i-f trimmer capacitors of the second i-f transformer, *T3,* for maximum output. As output meter reading increases due to approach to proper alignment, reduce signal generator output level to avoid overloading. Repeat the adjustment of the first i-f transformer, *T2.*

The receiver under discussion is known to utilize a 455-kc intermediate frequency. If in doubt about the exact peaking frequency when attempting alignment, it is best to refer to the manufacturer's notes for specific data. If this is not possible, feed the signal generator into the mixer grid, as in step c. Slowly rotate the signal generator frequency dial until some point between 500 kc and 150 kc is reached, where maximum output is indicated. The most common i-f values are 175 kc, 262 kc, 455 kc, and 465 kc; it is common practice to align

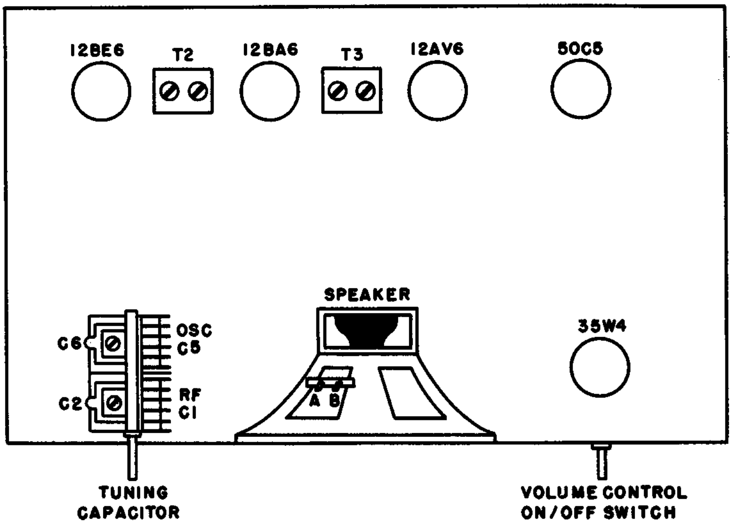

Fig. 26. Top view of chassis of the receiver shown in Fig. 25.

at the intermediate frequency closest to where a peak is observed. For example, if an early type set is out of alignment and the above procedure indicates a broad peak in the region of 270 kc, it is safe to assume that 262 kc is the proper i.f., and this value can be used as the basis for alignment. It is necessary, of course, to have the oscillator section shorted during this test to avoid extraneous beats, which may confuse the technician. It is also important to start tuning the signal generator from 500 kc and work down in frequency to avoid a misleading harmonic output effect. For example, a detuned 455-kc i-f section would indicate some output when fed a 230-kc signal, since the second harmonic would be 460 kc. Should the serviceman begin from the low end of the i-f range, he might be misled by the output indication at the 230-kc setting, and assume the correct i.f. is 262 kc; it would then be difficult to align the receiver. Had he continued up to the 455-kc frequency region, a greater output would have been indicated on the output meter due to use of a signal at the fundamental, or proper, i-f value. On the other hand, if he starts at 500 kc, he encounters fundamental response at 455 kc (or any other intermediate frequency of the receiver) before getting any harmonic response.

A sketch showing the physical layout of the components involved in alignment of the five-tube a-c/d-c set under discussion is shown in Fig. 26. It is strongly advised that the technician determine the location and function of each adjustment slug or trimmer before commencing with the alignment procedure.

40. Converter Alignment: Oscillator Section

a. With the i-f amplifier section properly aligned, the next step involves correct oscillator adjustment. Remove the short across the oscillator tuning capacitor, *C5*.

b. Remove the signal generator from the grid No. 3 point and couple the signal to the receiver loop antenna as follows: Construct a two-turn loop, about 6 inches in diameter (using insulated wire), and connect the signal generator output cable to the ends of the loop. Place the loop about 6 inches from the receiver loop antenna. (With sets not equipped with a loop antenna, the signal generator is fed into the antenna terminals through a 200-$\mu\mu$f blocking capacitor.)

c. Tune the receiver to a point on the high end of the dial that is not in use by any local station, between 1500 kc and 1700 kc. Assuming that 1540 kc is a quiet point, the signal generator is then set to 1540 kc and modulated. Oscillator trimmer *C6* is then carefully adjusted for maximum indication on the output meter.

41. Converter Adjustment: Mixer Section

a. Set the signal generator to 1400 kc, modulated. Slowly rotate the receiver tuning dial until maximum output reading is obtained in the vicinity of 1400 kc.

b. Vary trimmer capacitor *C2* until peak reading is observed on the output meter. This completes the overall receiver alignment.

42. Necessity for Realignment

In many instances, a receiver may exhibit symptoms that are initially interpreted as improper alignment, and yet may not require any retuning at all. Such conditions as poor sensitivity and/or selectivity, oscillation, or distortion may often be traced to open or leaky bypass capacitors, weak tubes, etc. It is wise to check, by simple d-c voltage readings at plate and screen terminals, whether leaky capacitors are at fault, and by tube substitution whether tubes require replacement. A leaky electrolytic filter capacitor or an open screen or avc decoupling capacitor may cause oscillation problems in the form of squeals, as the receiver is tuned through the broadcast range. In general, when the output increases as a trimmer is adjusted away from its correct frequency, realignment is probably necessary; on the other hand, when the trimmer adjustment has little or no effect on output, a defective component is the cause of trouble.

A rather commonplace occurrence in the present do-it-yourself era is the increasing number of misaligned sets brought into repair shops by inexperienced laymen who have attempted to improve the performance of their receivers. In these cases, a complete realignment job is necessary and a strong warning against further tampering should be given. Finally, it is generally approved practice to realign a set completely when a component of a tuned section is replaced. When an oscillator coil or i-f transformer, for example, is newly installed in a receiver that has not been realigned in some time, a full alignment job should be performed to bring the set to its optimum performance capabilities.

43. Review Questions

(1) What equipment is required for alignment of a superheterodyne receiver?
(2) What is the sequence of stages in alignment?
(3) Describe the alignment procedure for an i-f stage.
(4) Describe the alignment procedure for the oscillator stage.
(5) At what frequency is the mixer trimmer adjusted?

Chapter 6

MIXER AND CONVERTER TECHNOLOGY

44. General Design Considerations

In the process of frequency conversion, it must be recognized that a mixer or converter also yields frequencies other than the sum-and-difference frequencies of the fundamental local oscillator and incoming radio signal frequency. In some superheterodynes, particularly those designed for very high frequencies, it is undesirable to have the oscillator working in the same range as the received signal, so that harmonics of the oscillator may be used to mix with the incoming signal to produce the intermediate frequency. Although most of the discussions to follow are based upon the selection of the fundamental oscillator frequency for producing the if, it is important to bear in mind that this is not the only possible heterodyning frequency that may be used.

Modern broadcast radios utilize pentagrid converters of the 6SA7 type almost exclusively. On the other hand, triodes and pentodes may also be used as mixers with separate oscillator injection, especially where a reduction of noise is important. In particular, the factors that must be carefully considered in the design of triode and pentode mixers are (1) conversion transconductance, (2) conversion gain, (3) method of injecting local oscillator signal, (4) input and output impedances, (5) noise figure, and (6) biasing procedures. Each of these factors will be considered in detail in this chapter.

45. Conversion Transconductance

Conversion transconductance is a fundamental property of a mixer tube and is symbolized by g_c. It is related to the ability of the tube to produce signal current at the intermediate frequency in response to input voltage at the radio frequency for which the receiver is designed. Conversion transconductance is defined as follows:

$$g_c = \frac{I_{i\text{-}f}}{E_{r\text{-}f}} \tag{1}$$

where

$I_{i\text{-}f}$ = the peak output current of the mixer at the intermediate frequency, and

$E_{r\text{-}f}$ = the peak signal input voltage at the signal or r-f frequency.

Normally, the peak output voltage of the local oscillator is nearly equal to the cut-off voltage of the oscillator tube, and is quite large compared to the received signal. For this conventional condition, the grid-plate transconductance (g_m) of the mixer tube will be a function only of the local oscillator voltage. A standard example of the dependence of g_m on local oscillator voltage is shown in Fig. 27.

46. Determination of Conversion Transconductance Graphically

Conversion transconductance may be calculated directly from a set of equations representing a Fourier series and a final solution for which the integral calculus is required. It is also possible, however, to obtain g_c graphically from the g_m–vs.–E_c curve by a method described by E. W. Herold.[1] The procedure begins by plotting the g_m–vs.–E_c curve for the particular mixer being considered, with E_c being produced by a local oscillator. The value of g_m is measured for d-c bias voltages equal to the amplitude of the local oscillator signal at 30° steps starting at the negative peak of the local oscillator waveform and ending at the positive peak. Through this range, it is possible to obtain seven different values

[1] E. W. Herold, "Frequency Converters and Mixers for Superheterodyne Receivers," *Proc. I.R.E.*, February 1942.

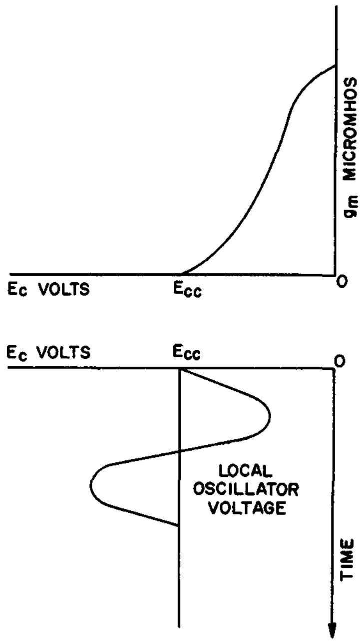

Fig. 27. The variation of mixer transconductance with applied local oscillator signal. The curve shown is generated by the first quarter-cycle of the oscillator voltage; it is called the "g_m vs E_c" curve.

of g_m for the tube. The conversion transconductance is closely approximated by Eq. 2. Note that this equation applies only to the case where the i-f is the difference frequency between the applied signal and the fundamental of the local oscillator.

$$g_c = \frac{1}{12}\left[(g_{m7} - g_{m1}) + (g_{m5} - g_{m3}) + 1.73\,(g_{m6} - g_{m2})\right] \qquad (2)$$

in which the various values of g_m represent the magnitude of this factor at the seven different points read from the curve. This may be clarified by reference to Fig. 28 in which the g_m–vs.–E_c curve for a typical pentode mixer (such as the 6AH6) is shown together with the applied oscillator voltage waveform. Note that Reading 1 for g_{m1} is taken at the negative peak as stated previously, with succeeding values obtained for 30° intervals all the way up to the positive peak. The amplitudes chosen for this figure apply to the solution of the following example.

Example: Using the curve of Fig. 28, determine:
(a) the peak local oscillator voltage
(b) the bias voltage E_{cc}
(c) the grid-plate transconductance at each 30° interval
(d) the conversion transconductance when the fundamental of the local oscillator is used as the heterodyne frequency for producing the if.

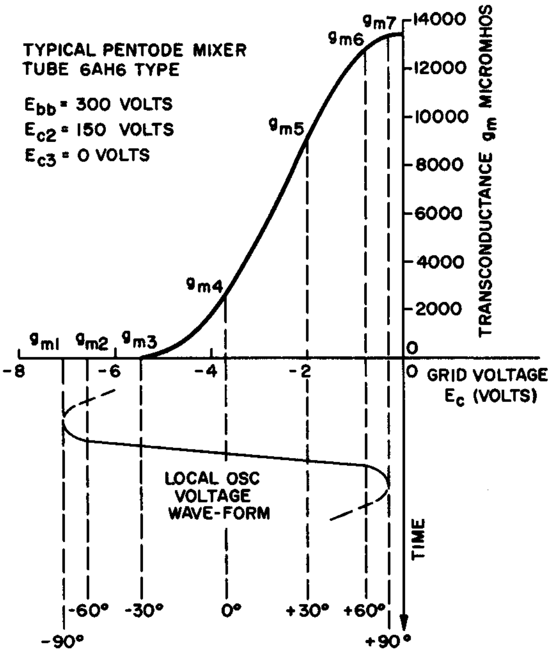

Fig. 28. Determination of g_c by graphical method.

Solution: (a) The peak oscillator voltage is obtained merely by subtracting the appropriate abscissae along the grid volts axis, i.e.,

$$E_{po} = -3.7 - (-0.3)$$
$$= -3.4 \text{ volts}$$

(b) The bias voltage is the voltage measured between the zero abscissa and the 0° point on the oscillator waveform. That is, the bias voltage is –3.7 volts.

(c) As read from the curve, the tube transconductances at the 30° intervals are:

$g_{m1} = 0$ $\quad g_{m2} = 0$ $\quad g_{m3} = 0$

$g_{m4} = 2500$ micromhos $\quad g_{m5} = 9300$ micromhos

$g_{m6} = 12{,}700$ micromhos $\quad g_{m7} = 13{,}200$ micromhos

(d) The conversion transconductance is found from Herold's formula (Eq. 2):

$$g_c = \frac{1}{12}\,(13{,}200 + 9300 + 1.73 \times 12{,}700)$$

$$= \frac{1}{12}\,(44{,}471) = 3706 \text{ micromhos}$$

According to another paper published by E. W. Herold[1], a close approximation to the actual conversion transconductance as obtained above can be elicited very simply by dividing the maximum grid-plate transconductance of the tube by 4. If we do this, we have:

$$g_c = 13{,}400/4 = 3350 \text{ micromhos}$$

This result is approximately 10% lower than the actual value but provides the engineer with an approximate starting figure for conversion transconductance.

47. Injection of Local Oscillator Signal

The received r-f signal is normally applied as a voltage to the control grid of the mixer tube. The output of the local oscillator can be applied to either the control grid or the cathode of a triode. For pentodes, it is possible to inject the local oscillator voltage at either the screen or suppressor grid as well as at the control grid or cathode. In all cases, it must be remembered that intermodulation between two signals can take place only in a nonlinear device if a difference frequency is to be expected. This means that mixers must be biased so that they behave nonlinearly; i.e., they must be biased close to cut-off as for Class B amplification.

Grid injection methods are normally objectionable because there is likelihood of interaction between the local oscillator output and the received signal, making tracking difficult. In addition, grid injection tends to cause radiation of the local oscillator waveform into space — a possible source of interference to other receivers in the vicinity. Despite these objections, grid injection is still found in certain special cases, see Fig. 29. The balanced mixer shown in Fig. 29C reduces signal interaction and interference radiation to a minimum, and is encountered quite often in high-frequency equipment of the radar and navigational variety.

[1] E. W. Herold, "Phase Reversal Modulation," *Proc. I.R.E.*, April 1946.

Injection of the local oscillator signal at the cathode of the mixer minimizes reradiation and interaction effects, but introduces other disadvantages. For example, the cathode input impedance of a mixer is relatively low and tends to load the local oscillator. This

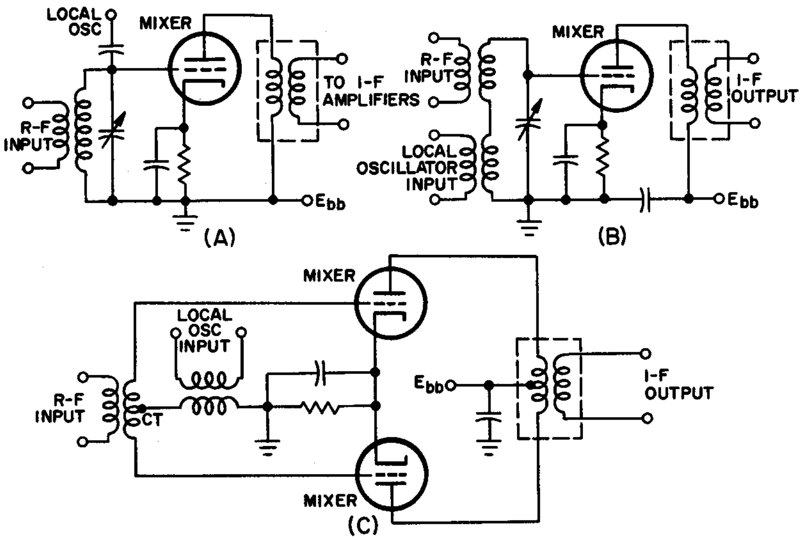

Fig. 29. Grid injection methods: (A) capacitive coupling; (B) inductive coupling, and (C) balanced mixer.

problem must be overcome by extremely careful design, particularly in ultra-high frequency equipment. Two common cathode-injection arrangements are presented in Fig. 30.

A second disadvantage of cathode-injection results from inductive reactance that may either be actually present (as in Fig. 30A) in the cathode lead or may be coupled back into this lead by outside connections. For example, since the normal local oscillator is tuned to a frequency higher than the r-f signal, the local oscillator circuit appears inductive at the signal frequency. This reactance, coupled back into the cathode of the mixer tube, may cause feedback through the grid-to-cathode capacitance with consequent serious loading of the mixer input circuit. This is particularly true at very high frequencies.

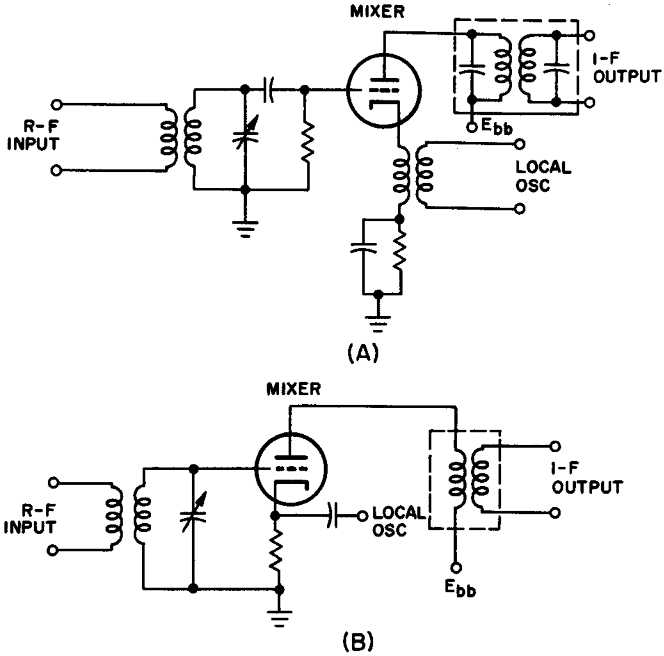

Fig. 30. Cathode-injection of local oscillator signal. (A) Inductive coupling; (B) capacitive coupling.

48. Conversion Gain

Conversion gain of a triode or pentode mixer is generally given in terms of power rather than voltage and is defined as the ratio of available i-f signal power at the mixer output to the r-f signal power at the output of the receiver tuned circuit. Although equations in terms of fundamental quantities are encountered in the literature, conversion gain is generally measured rather than computed.

The *voltage amplification* of a mixer tube is easily determined when the tube is a pentode or a pentagrid converter where the equivalent plate resistance is very much greater than the load impedance into which the tube works. In this case, mixer voltage amplification (often called conversion gain) is given by:

$$\text{Mixer amplification} = g_c Z_L \tag{3}$$

As shown in the example in Section 46, conversion transconductance, g_c, is approximately 1/4 the maximum grid-plate transconductance of the same tube (g_m). Thus, it is found that the amplification of a mixer from Eq. 3 is approximately 1/4 of the voltage amplification that might be expected from the same tube when used as a voltage amplifier. Note that the voltage amplification of a pentode is obtained from an equation similar to Eq. 3 with g_m substituted for g_c.

49. Mixer Noise

In addition to noise coming from outside the receiver (atmospheric, cosmic radiation, machine sparking and radiation, precipitation static, etc.), a significantly large amount of noise is generated within the mixer itself. Such noise derives from phenomena known as shot effect, flicker effect, secondary emission, and positive ion effect. In addition to these vacuum tube noise sources, other places within the receiver itself where noise can be generated are the electrical conductors in which thermal noise is generated, resistors in which contact and breakdown noise occurs, and magnetic components in which the Barkhausen Effect is encountered.

All of the noise sources originating within the mixer tube have been quantitatively analyzed in papers appearing over the years. For the purposes of this book, we are interested in determining the magnitude of the tube noise rather than its sources. This may be done in several ways; the use of an equivalent grid resistance (R_{eq}) approach is perhaps the simplest of all and is treated briefly below.[1]

Suppose we had a perfectly noiseless tube and we placed in its grid circuit a resistance that would generate noise by thermal, contact, and breakdown effects. Suppose that this resistance was then adjusted to generate just enough noise at the signal frequency so that, after conversion to the if, the noise was equal in magnitude to the i-f noise that is actually present. Such a resistor would be designated as R_{eq}. Obviously, if R_{eq} is small, the noise will be small; conversely, if R_{eq} is large, the noise will be more severe.

The formulas provided for triodes, pentodes, and multigrid mixers by Harris are:

$$R_{eq} = \frac{4}{g_c} \text{ (for triode mixers)} \tag{4}$$

[1] W. A. Harris, "Vacuum Tube Amplifiers and Input Systems," *R.C.A. Review*, vol. 5, p. 505, April 1941.

$$R_{eq} = \frac{I_b}{I_b + I_{c2}}\left(\frac{4}{g_c} + \frac{20\ I_{c2}}{g_c^2}\right) \quad \text{(for pentode mixers)} \quad (5)$$

$$R_{eq} = \frac{20\ I_b}{g_c^2\ I_{sp}}\ (I_{sp} - I_b) \quad \text{(for multigrid mixers)} \quad (6)$$

The factor g_c is the conversion transconductance of the tube averaged over the local oscillator input cycle.

In these equations I_{sp} equals total space current, I_b equals plate current, and I_{c2} equals screen current. All are averaged over the oscillator cycle just like the conversion transconductance.

The equations and actual measurements taken on a variety of tubes in operation show that the noise from triode mixers is substantially less than that originating in pentode mixers, and that pentagrid coverters are by far the worst offenders. Three representatives tubes and their values for R_{eq} are given below:

Triode mixer: 6J5	R_{eq} = 6500 ohms
Pentode mixer: 6SG7	R_{eq} = 13,000 ohms
Pentagrid converter: 6SA7	R_{eq} = 240,000 ohms

50. Bias for Mixers

Bias for triodes and pentode mixers may be obtained from an external voltage source, grid-leak bias, or through the use of a cathode resistor. External (or fixed) bias, although sometimes used in special applications, is not as desirable as either of the other two methods. This follows because the conversion transconductance of the mixer is seriously affected by variations in the local oscillator voltage when fixed bias is used. The amplitude of the local oscillator voltage, a function of line voltage and oscillator tuning, is therefore apt to vary over relatively wide limits.

Conversion transconductance is substantially independent of the local oscillator output voltage when either cathode bias or grid-leak bias is employed. As in any circuits in which only grid-leak bias is found, there is danger of excessive plate current if the local oscillator should fail. Hence, it is customary to use an additional small resistor in the cathode circuit of the mixer to serve as a source of protective bias should failure occur.

Another very successful biasing scheme may be used with pentode mixers. Instead of utilizing the safety bias feature provided by an additional cathode resistor, all the bias is obtained by the grid-leak method. Protection for the tube is then added in the form of a series screen resistor selected so that neither plate nor screen current can become excessive if the bias falls to zero. This biasing method reduces the influence of oscillator voltage on the conversion transconductance to a greater degree than any of the other biasing systems described.

51. Multigrid Mixers and Converters

Multigrid mixers discussed qualitatively in Chapters 2 and 3 were originally designed to minimize the interaction between local oscillator and the r-f signal source that tends to occur in triode and pentode mixing systems. To differentiate between a tube such as the 6L7 pentagrid mixer, which requires a separate, external local oscillator, and a tube such as the 6SA7, which contains the elements required for oscillation as well as mixing, the latter type is generally called a converter. The discussion that follows, although specifically applied to the multigrid mixer (6L7), applies equally well to the pentagrid converter (6SA7 type), and the triode-hexode converter (6K8G type).

The physical structure of the 6L7 is shown in Fig. 10. A typical circuit using a separate oscillator is illustrated in Fig. 11. In the latter, it should be observed that the oscillator is coupled to the third grid while the signal input is applied to the first grid. This method of local oscillator injection is known as *outer grid injection.* It is also possible to reverse the two connections to obtain *inner grid injection* in which the local oscillator voltage is applied to the inner grid. The performance of the mixer is quite different for each of these arrangements.

Inner grid injection. When the local oscillator voltage acts upon the inner grid (G1), the space current flowing to all the other tube elements is modulated at the oscillator frequency, see Fig. 31. The outer grid (G3) is the grid to which the r-f signal is applied. The grid-plate transconductance of this grid is a function of the current passing through it; hence, the transconductance varies at the local oscillator frequency. Thus, the mixing action is very similar to the action that occurs in a triode mixer where it has been shown that grid-plate transconductance is a function of oscillator voltage.

The screen grid (G2) isolates the signal and oscillator circuits from each other, but as will be shown, does not completely eliminate the coupling between G1 and G3. The screen grid (G4) is an accelerating electrode and the suppressor grid (G5) serves the same function as in a standard pentode. The mechanical and electrical isolation between the oscillator grid and the signal grid makes it

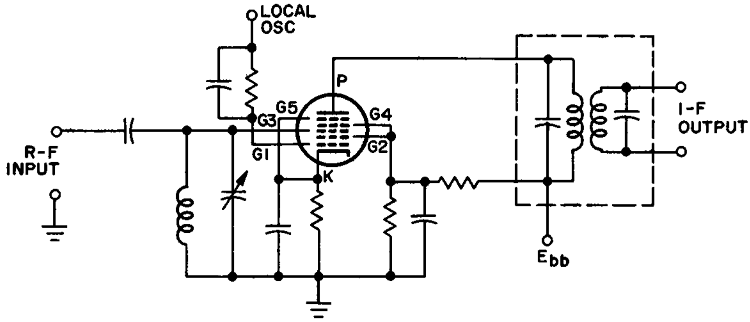

Fig. 31. Pentagrid mixer circuit illustrating inner-grid injection in which the local oscillator voltage is applied to the inner grid and the signal voltage to the outer grid. Refer to Fig. 11 for a circuit using outer grid injection.

possible to drive the oscillator grid into the conducting region without loading the signal source. This is a definite advantage. Grid-leak bias is generally used for the oscillator grid (G1) because this biasing system provides the maximum signal grid-plate transconductance.

Outer grid injection. In outer-grid injection, the voltage applied to the signal grid (G1) modulates the cathode current to all of the other elements. The local oscillator voltage, on the other hand, controls the distribution of space current between the screen grid and the plate. This difference gives rise to transconductance effects that must be considered in selecting the injection system to be used. As the curves in Fig. 32 indicate, the inner grid has a somewhat higher grid-plate transconductance (maximum) than the outer grid. In some applications, this difference may affect the performance of the equipment. An additional divergence in performance is caused by *space-charge coupling* described below.

Space-charge coupling. As previously mentioned, G2, G4, and G5 provide good shielding and reduce capacitive coupling between G1 and G3 to a minimum, as well as coupling from these grids to the plate. Another type of coupling exists, however. Known as space-charge coupling, this effect is explained as follows: particularly with inner grid injection, when the local oscillator voltage on G1 is such as to permit the passage of electrons, a space charge forms between G3 and G1 because G3 is biased negatively. After the local oscillator passes its positive voltage peak and fewer electrons pass through it, the magnitude of the negative space charge decreases so that the cloud moves farther away from the signal grid; on the positive upswing of the local oscillator grid, the space charge increases in intensity, thereby moving closer to the signal grid, G3. This advance and retreat of the space charge induces a voltage between the grid and cathode at the local oscillator frequency, causing a current of the same frequency to flow in the external grid impedance. The current flowing in the grid impedance lags behind the oscillator voltage by slightly more than 90°. This is effective coupling between these two grids and may result in instability.

A current lag of more than 90° may be considered the result of an inductance and a *negative* resistance (imaginary) connected between G1 and G3. Hence, space-charge coupling may be eliminated by resonating a capacitance and a positive resistance with these elements. This is generally done by connecting a small capacitor and resistor in series between G1 and G3 and adjusting both until zero local oscillator voltage appears across the signal circuit.

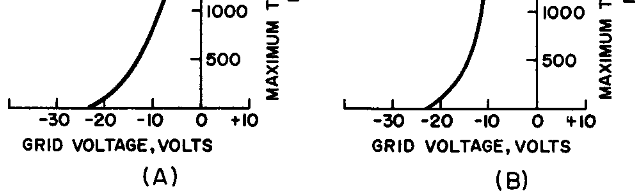

Fig. 32. Maximum transconductances of inner and outer grids considered separately. (A) Outer grid, when $E_{c1} = 0$; (B) inner grid, when $E_{c3} = 0$.

52. Vacuum Diode Mixer

Since a diode is not an amplifier, a diode mixer always introduces a conversion loss rather than a conversion gain. Yet the principal mixer used at very high frequencies (500 mc) is the diode type because of its inherently low noise figure at these frequencies. The diode circuit, as might be anticipated, is quite simple in structure, see Fig. 33.

The conductance of a diode is, generally, a function of the local oscillator voltage since the r-f signal voltage is, generally, so much smaller. If the diode is linear, its conductance remains essentially constant during the conduction period of the tube. On the other hand, *nonlinear behavior* necessary for mixing is obtained through

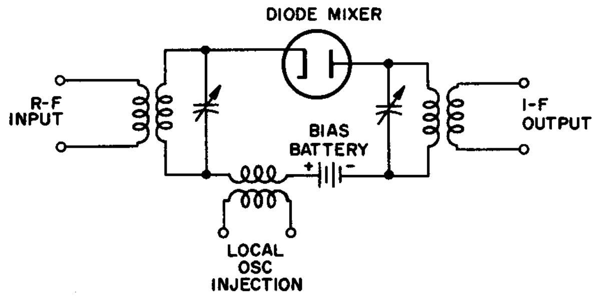

Fig. 33. Basic diode mixer circuit with battery bias. In many high-frequency mixer designs, a high quality bias cell is actually used.

straightforward rectifier action. It can be shown, however, that minimum conversion loss occurs when the local oscillator signal is made very strong, when the bias is made very negative, and when the diode conducts only at the peaks of the local oscillator signal. These characteristics are shown in Fig. 34.

The development of the equations that describe diode action depend upon the application of a Fourier series and subsequent integration over the cycle to determine the coefficients. Although this derivation cannot be undertaken here, the final expressions for i-f and r-f currents are presented merely to show the terms they contain.

$$i_{r\text{-}f} = g_o E_{r\text{-}f} \sin (2\pi f_{r\text{-}f} t) - g_c E_{i\text{-}f} \sin (2\pi f_{r\text{-}f} t) \quad (7)$$

in which $i_{r\text{-}f}$ is the current at the oscillator frequency, g_o is the average diode conductance, $E_{r\text{-}f}$ is the oscillator voltage, f is the frequency, and t is the time. The current at the intermediate frequency is determined by a similar equation:

$$i_{i\text{-}f} = g_c E_{r\text{-}f} \sin (2\pi f_{i\text{-}f} t) - g_o E_{i\text{-}f} \sin (2\pi f_{i\text{-}f} t) \quad (8)$$

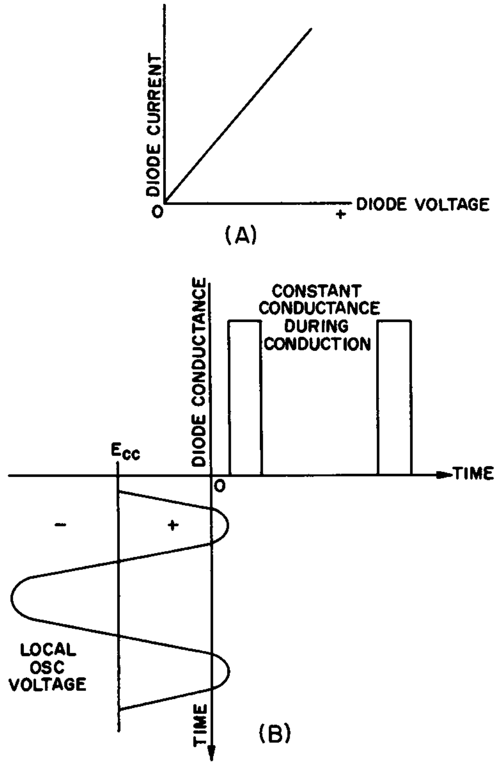

Fig. 34. The behavior of a diode mixer when biased for minimum conversion loss. Note that conditions are selected so that the conduction interval is extremely short. (A) Diode current vs. diode voltage for a linear diode; (B) variation of diode conductance with time for conditions above.

where g_c is the conversion transconductance of the diode.

Thus, the i-f current that flows in the output of the diode mixer is a function of the conversion transconductance and average diode conductance, as well as the applied signal voltage.

53. Implications of Equivalent Circuit

Herold[1] has shown that the equivalent circuit given in Fig. 35 satisfies both Eq. 7 and Eq. 8. The diode mixer, therefore, turns out to be a symmetrical π attenuator[2]. In such an attenuator, maxi-

[1] *Proc. I.R.E.*, October, 1943.
[2] *"Mixers and Attenuators"*, Alexander Schure. John F. Rider Publisher, Inc.

mum power transfer is realized when the input and output (source and load) impedances are matched to the input and output impedances, respectively, of the mixer itself.

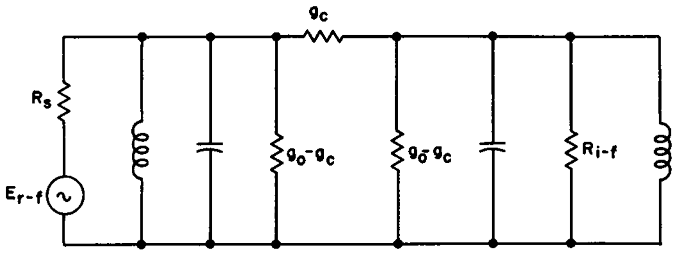

Fig. 35. Equivalent circuit of diode mixer is seen to be a π attenuator. R_s is source impedance (or resistance) and $R_{i\text{-}f}$ is the i-f load resistance.

When the mixer is matched on an image basis, both R_s and $R_{i\text{-}f}$ are equal to the characteristic impedance of the mixer network, and the value of R_s and $R_{i\text{-}f}$, which provide maximum power transfer through the mixer, is given by Eq. 9.

$$R_s = R_{i\text{-}f} = Z_m = \frac{1}{\sqrt{g_o^2 - g_c^2}} \tag{9}$$

For the same condition, i.e., mixer impedances matched at input and output by the source and load resistances, respectively, the conversion power gain of the mixer may be obtained from:

$$G_c = \left[\frac{g_c/g_o}{1 + \sqrt{1 - (g_c/g_o)^2}}\right]^2 \tag{10}$$

It is convenient to plot the conversion power gain for this same condition as a function of the ratio g_c/g_o to observe how G_c varies. This is illustrated in Fig. 36.

Clearly, the conversion power gain is always less than one, approaching unity only as the ratio g_c/g_o approaches unity. The Fourier series previously mentioned shows that this ratio can approach unity only when the diode conducts for an extremely short part of the local oscillator voltage cycle. This explains why diode mixers are biased heavily as in Fig. 34.

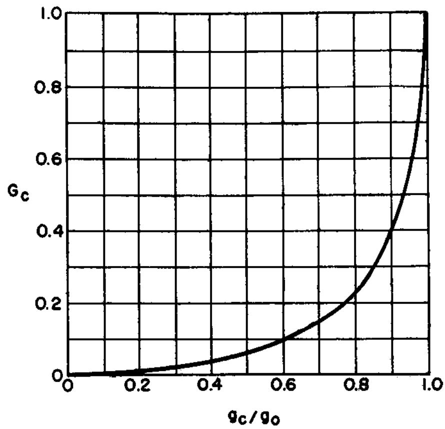

Fig. 36. Conversion gain of a diode mixer for conditions described in the text.

54. Crystal Mixers

Crystal mixers are employed in the microwave region for the same reason that diodes are used: their noise figures are quite low. Although silicon mixers have long been favored as mixers for frequencies in the range of 3000 mc, recent advances in design have indicated that germanium diodes may offer equal advantages over the entire microwave spectrum. In the region around 500 mc, welded contact germanium diodes are superior to both vacuum diodes and silicon mixers with respect to noise.

The performance of a crystal mixer is nearly the same as that of a vacuum tube diode. There are two significant differences, however, that must be taken into account in mixer design: (1) the reverse resistance of a crystal diode is not as great as that of the equivalent vacuum tube, hence some reverse current flows; (2) the best performance with respect to noise is obtained when a crystal is operated with little or no bias.

Figure 37 illustrates a simple form of unbiased crystal mixer circuit. Note its similarity in all respects but bias to the vacuum diode circuit. The volt-ampere characteristic curve of a typical crystal diode is given in Fig. 38.

The noise produced by a crystal diode mixer depends upon temperature and the conversion gain of the stage. Both these fac-

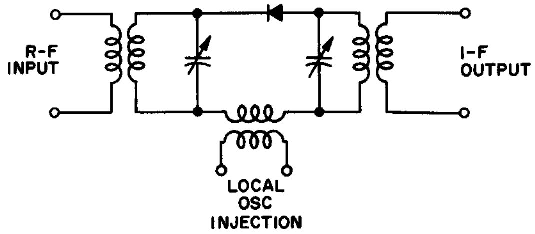

Fig. 37. Typical unbiased crystal mixer circuit showing lumped components. In the microwave region, the usual waveguide "plumbing" replaces the lumped L and C shown.

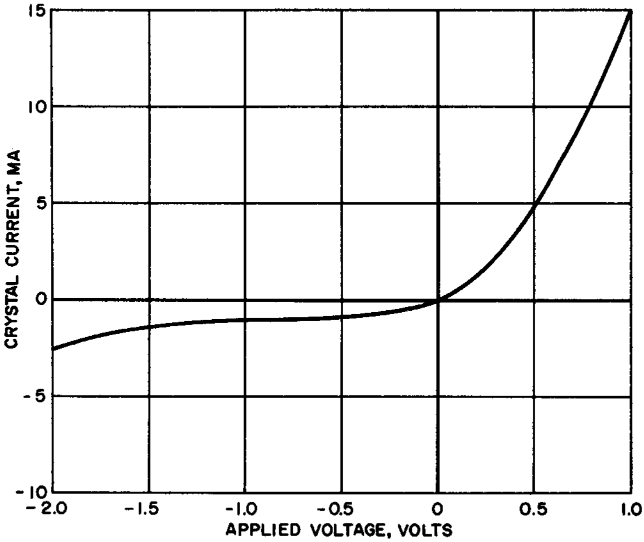

Fig. 38. Volt-ampere characteristics of a typical silicon mixer crystal. Note the reverse current at negative diode voltages.

tors, in turn, are functions of the level of the applied local oscillator signal. Least noise is generally obtained when the crystal current is about 0.5 ma, which corresponds to about 0.5 milliwatts of local oscillator power.

55. Review Questions

(1) Describe the factors which must be taken into consideration in the design of triode and pentode mixers.
(2) Define conversion transconductance.

(3) Explain how mixer transconductance (grid-plate) varies with local oscillator input voltage.
(4) Describe fully the method of obtaining the conversion transconductance of a triode mixer graphically.
(5) How may the conversion transconductance of a triode or pentode mixer be approximately determined very quickly?
(6) Describe the methods whereby a local oscillator signal may be injected in a triode or pentode mixer circuit. State the advantages of each injection method.
(7) Define conversion gain. How is the conversion gain of a mixer generally determined?
(8) State the equation for finding mixer voltage amplification. How does the voltage applification of a mixer pentode compare with its voltage amplification when used as a straight amplifier?
(9) Describe the sources of a mixer noise. What is meant by the "equivalent noise resistance" of a mixer?
(10) Using diagrams, explain the various methods that have been used successfully for mixer biasing. Which method is preferred? Why?
(11) What was the major goal in the original design of a multigrid mixer or converter?
(12) How does a multigrid mixer differ from a multigrid converter?
(13) Describe the circuits used for inner grid and outer grid local oscillator injection. In which arrangement is space-charge coupling least objectionable?
(14) How is space-charge coupling minimized? Why does this system work?
(15) Why are vacuum diode mixers popular for v.h.f.?
(16) What are the principal differences between a crystal diode and a vacuum diode mixer relative to performance and circuit design?

Chapter 7

LOCAL OSCILLATOR TECHNOLOGY

56. Oscillator Output Considerations

As we have shown in the preceding chapter, maximum conversion transconductance in any type of mixer is obtained when the amplitude of the local oscillator signal applied to the mixer is correct. For triodes and pentodes, the peak voltage of the local oscillator signal at the point of injection should be between 6 and 10 volts. Multigrid mixers and converters often perform satisfactorily when the injection voltage is as low as three volts. On the other hand, a vacuum diode requires an injection voltage of the order of 20 volts peak if maximum conversion gain is to be realized. A crystal diode mixer calls for an injection voltage large enough to produce a crystal current of about 0.5 ma.

Since interaction between mixer and local oscillator is to be avoided, the oscillator is generally coupled to the mixer very loosely. This leads to a voltage loss and must be compensated for by raising the oscillator output voltage and power. Normally, the oscillator output voltage is adjusted to be at least 5 to 10 times greater than the voltage required at the injection point to insure adequate mixer drive.

57. Frequency and Amplitude Stability

The importance of good frequency stability of the local oscillator is very evident in ultra-high frequency receivers. The need for retuning FM and television receivers during the first 10 or 15 minutes of operation is generally the result of local oscillator drift. The maximum variation permissible for drift-free receiver operation is approximately 20% of the i-f amplifier bandwidth.

In properly designed receivers operated in locations where there are no severe extremes of temperature, the allowable warm-up time is not usually greater than the time required for all the resonant circuit components to reach thermal equilibrium. For such receivers, no special precautions aside from structural rigidity, care in placement of parts and lead dress, and the use of high quality components are required. If the permissible warm-up time is very short, however, frequency drift may be excessive unless temperature compensated capacitors and resistors are used.

Another important source of oscillator frequency drift is local oscillator voltage variation, and variation in electrode voltages. This effect is minimized in v.h.f. receivers by using voltage-regulated power supplies.

Changes in local oscillator output also seriously affect mixer conversion transconductance and lead to amplitude instability. Most often, the amplitude of the local oscillator varies because the receiver, and hence the oscillator, is tuned over a relatively wide band of frequencies. As tuning proceeds, the Q of the oscillator resonant circuit changes, thereby resulting in output variations. To minimize the effects of local oscillator amplitude changes, grid-leak mixer bias is recommended.

58. Choice of Oscillator Tubes

At relatively low frequencies — such as the range from the broadcast band up to about 30 mc — pentagrid converters and similar multigrid types have proved adequate. Between 30 mc and about 2500 mc, conventional triodes have been used as oscillators with good success. At frequencies much above 2500 mc, the effects of electron transit time become a serious handicap and the local oscillator-mixer circuit becomes almost useless. Special miniature triodes and pentodes provide better service but still leave much to be desired. Thus, in the microwave range above 4000 mc, the reflex klystron has superseded all other types almost completely.

59. The Tuned-Grid Oscillator

The most frequently-used circuits for local oscillators in superheterodyne receivers are the tuned-grid type and the Hartley. Referring to Fig. 39, there is an untuned inductance in the plate circuit which is inductively coupled to the resonant LC grid circuit. The tube causes a 180° phase difference between its grid signal and plate

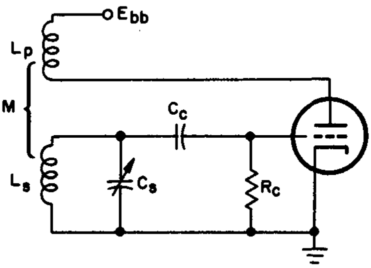

Fig. 39. Circuit of the tuned-grid oscillator showing the relationship between components.

signal; when L_p and L_s are coupled properly, a second phase shift of 180° can be realized. Thus, in-phase feedback from plate to grid can be accomplished and, if the tube's amplification is sufficiently great to make up for the resistance losses in the resonant circuit, sustained oscillation will occur.

In the design of oscillator circuits, two factors must be determined in the design procedure: (1) the minimum amplification factor the tube must have to sustain oscillation and (2) the frequency at which a given circuit will oscillate. Equations 11 and 12 – the design equations – are based upon the assumption that the resistance of R_c, the grid return resistor, is considerably greater than the capacitive reactance of the tuning capacitor, C_s. In these equations, use is made of the following symbols:

μ = tube amplification factor

L_p = inductance of plate coil in henries

Q_p = Q of L_p defined as the ratio of $2\pi fL_p/r_p$

L_s = inductance of grid coil in henries

Q_s = Q of L_s defined as $2\pi\ fL_s/R$

C_s = capacitance of tuning capacitor in farads

R = resistance in series or part of L_s in ohms

r_p = resistance, plate, in ohms

M = mutual inductance of primary to secondary in henries

f = frequency of oscillation in cps

Equation 11 provides the information required to determine the minimum amplification factor in terms of the other parameters.

$$\mu \geqq \frac{L_p}{M}\left(\frac{1}{1+\frac{Q_s}{Q_p}}\right) + \frac{M}{L_s}\left(\frac{1}{1+\frac{Q_p}{Q_s}}\right) + \frac{C_sRr_p}{M} \qquad (11)$$

Equation 12 provides the necessary information to determine the frequency at which the oscillator will operate.

$$f = \frac{1}{2\pi\sqrt{L_sC_s}\left(1+\frac{RL_p}{r_pL_s}\right)} \qquad (12)$$

A qualitative analysis of Eq. 11 is rendered substantially easier by assuming that the ratio of Q's is unity, a rather normal situation. In this case, Eq. 11 simplifies to:

$$\mu \geqq \frac{L_p}{2M} + \frac{M}{2L_s} + \frac{C_s\ Rr_p}{M} \qquad (13)$$

or further to:

$$\mu \geqq \frac{L_p + \frac{M^2}{L_s} + 2C_s\ Rr_p}{2M} \qquad (14)$$

The effect of each of the variables on the right side of the equation on μ now becomes clear. Thus:

(1) As L_s is made larger and C_s made smaller to maintain resonance at a given frequency, the magnitude of μ that is required for sustained oscillation decreases. Thus, a tube with lower amplification factor will oscillate more easily in a circuit in which the resonant L/C ratio is high.

(2) If L_p (the feedback winding) has more turns or has its inductance increased in any other manner, the necessary amplification factor increases.

(3) From Eq. 13 it can be seen that as M increases, the amplification required increases. Hence, to make a low amplification factor tube oscillate more readily, a smaller mutual inductance (or looser coupling) may be used.

(4) Since both R and r_p appear only in the numerator, these should be made as small as possible to reduce the required μ. However, r_p is determined chiefly by the tube design, hence only R is under control by the designer.

Equation 12 may be similarly analyzed; in this case, it is comparatively easy to show that the term RL_p/r_pL_s approaches zero in many oscillators. When this occurs, the frequency of oscillation becomes equal to the frequency of resonance of the grid circuit. Thus:

(1) R is made as small as possible and, with r_p generally quite high, the ratio R/r_p becomes quite small.

(2) Since L_s is a coil having many times the inductance of L_p, the ratio L_p/L_s may be very small.

(3) These two small ratios in product form become even smaller so that RL_p/r_pL_s may approach zero.

At best, this ratio may be considered as a rather small correction on the equation for simple resonance so that an oscillator's frequency is quite close to the resonant frequency in most practical circuits.

60. The Hartley Oscillator

Virtually all pentagrid converters of the 6SA7 type utilize Hartley oscillators for local heterodyning. This circuit requires only one coil with a tap, rather than two individual coupled coils. The basic Hartley is shown in Fig. 40.

The grid is connected to the top of the tuning coil through C_c, placing it at the same instantaneous potential as the top of the coil. Similarly, the plate is joined to the bottom of the coil through C_{BP} so that the instantaneous voltage phase difference between the grid and plate is 180°. Since the tube produces a second phase inversion of 180°, any feedback or coupling between plate and grid is regenerative and may lead to oscillation. Feedback from one circuit to the other occurs through the mutual inductance, M, the extent of the feedback being a function of the number of turns included between the tap and common ground. It is also feasible to operate the grid or the cathode at ground potential if desired.

The design equations for the Hartley oscillator are substantially simpler than those given for the tuned-grid type. The frequency of oscillation is given by Eq. 15:

$$f = \frac{1}{2\pi\sqrt{LC}} \tag{15}$$

where L is the total inductance included between grid and plate and is equal to $L_1 + L_2 + 2M$.

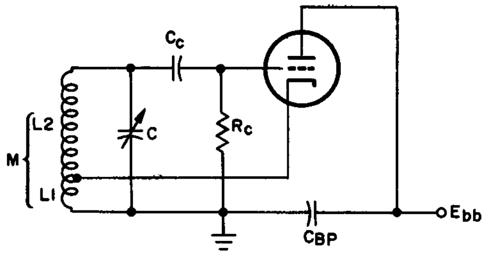

Fig. 40. The Hartley oscillator circuit. Note the mutual inductance (M) between the two sections of the tuning inductance.

For sustained oscillation, the amplification factor of the tube must be equal to or must exceed the value of the right-hand member of Eq. 16:

$$\mu \geqq \frac{L + L_1 - L_2}{L - L_1 + L_2} \tag{16}$$

substituting $L_1 + L_2 + 2M$ for the total inductance L, we have

$$\mu \geqq \frac{L_1 + M}{L_2 + M} \tag{17}$$

and since the inductance sections are in the same ratio as the square of the number of turns in the respective sections, we may write:

$$\mu \geqq \frac{N_1^2}{N_2^2} \tag{18}$$

All of the foregoing equations are based upon the assumption that the Q's of the coils are infinite. The corrections required for practical circuits having coils with finite Q are relatively small and inconsequential.

61. Oscillator Tracking

The standard broadcast superheterodyne ordinarily is adjusted so that its local oscillator tunes above the r-f circuits in frequency by an amount equal to the intermediate frequency of the receiver. Since the r-f and oscillator tuning units (generally the capacitors) are ganged so that rotation for a given adjustment is equal for each, a tracking problem arises: the local oscillator must tune over a smaller percentage of its center frequency than the r-f resonant circuit. Consider the broadcast band which has a tuning range from 550 kc to 1600 kc (a band of 1050 kc). The center frequency of this band is 1075 kc, so that the band-to-center-frequency ratio is 0.975. The oscillator working 455 kc higher at all times than the rf, therefore ranges from 1005 kc to 2055 kc, also a band of 1050 kc. The center-frequency of this band, however, is 1530 kc so that the band-to-center-frequency ratio is 0.685. This means that simple resonant circuits having similar characteristics cannot be used without the two systems going "out of track" over most of the tuning range.

There are two ways to solve this problem. The first one — a very popular solution for inexpensive broadcast radios — is based upon the use of two identical tuning inductances (for rf and oscillator). A specially-shaped, oscillator tuning capacitor is then used to maintain a constant difference between the two resonant circuits over the entire band. The construction of the rotor and stator plates must be planned so that the capacitance of the local oscillator tuning capacitor is always related to the capacitance of the r-f tuning capacitor according to the equation:

$$C_o = \frac{C_{rf}}{(1 + f_{i\text{-}f}/f_{r\text{-}f})^2} \qquad (19)$$

where C_o is the capacitance of the local oscillator tuning capacitor for any frequency $f_{r\text{-}f}$, and $C_{r\text{-}f}$ is the capacitance of the r-f tuning capacitor for the same frequency. Since small tracking errors due to imperfections of the special tuning capacitor have negligible effect on the broadcast band where the signals are usually very strong, this solution is quite acceptable.

Receivers used for short-wave reception, however, handle the prolem in a different manner. Here the ganged capacitor sections are generally identical and auxiliary capacitors are added to the oscillator circuit to maintain good tracking. A series capacitor of large value, called a *padder,* is used for tracking near the low-frequency end of the desired band, and a small parallel capacitor called a *trimmer,* (see Fig. 41) aligns the high frequency end of the band. If the trimmer, padder, and tuning capacitors, and inductances are properly computed, the two stages will also track perfectly at one other frequency very close to the center of the band.

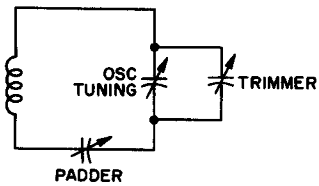

Fig. 41. Position of trimmer and padder capacitors in the local oscillator circuit.

Throughout the remainder of the band, the tracking error may be maintained reasonably small with correct design. The curve of Fig. 42 shows the type of error and normal magnitude of "out-of-track" condition for a well-designed broadcast superheterodyne.

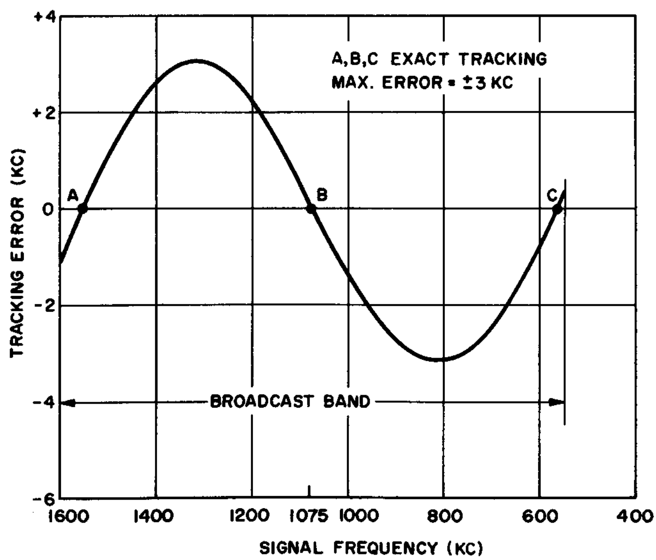

Fig. 42. Trimmer-padder tracking system provides tracking at three frequencies and a maximum tracking error of only ±3 kc over the rest of the band.

Equations for calculating the required components are lengthy and complex; they will not be considered here. For those interested in this type of design, a complete description of the calculation procedure may be found in several handbooks on this subject.[1]

62. Automatic Frequency Control of Local Oscillators

As the frequency to which a receiver is tuned increases, it becomes more difficult to maintain the local oscillator frequency constant in the face of temperature and other ambient changes. Variation of local oscillator frequency for a given setting of the tuning system causes proportional variations in the if produced and results in detuning of the signal. This effect, especially noticeable in FM receivers and in television, is referred to as *oscillator drift.*

Automatic frequency control, or AFC, is an electronic system in which the local oscillator frequency is partially controlled by a feedback network which maintains a constant difference between the r-f signal and the oscillator frequency. In superheterodyne receivers of the entertainment class, the arrangement used is called "difference frequency AFC". In this system, the r-f signal to which the mixer is tuned is called the *reference frequency*. In operation, the local oscillator signal is heterodyned with the reference signal, and the difference signal is fed to a discriminator. The output voltage of the discriminator is dc, and is proportional to the difference between the *desired* if and the *actual* if. This d-c voltage is used to control a reactance tube connected across the tuned circuit of the oscillator. The block diagram in Fig. 43 illustrates the interconnections used in this AFC system.

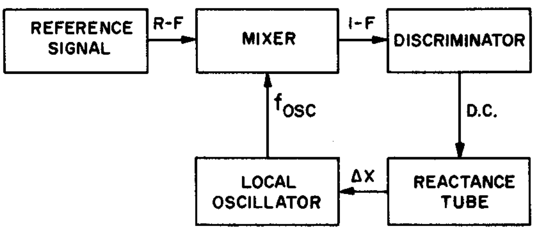

Fig. 43. Block diagram of "difference-frequency AFC". The reference signal is the incoming rf to which the receiver is tuned.

[1] Most of these discussions are based on original material appearing in the *Radiotron Designer's Handbook,* an RCA publication.

Consider a simple example: a superheterodyne receiver is used to receive a transmitter at 710 kc; if the intermediate frequency of this receiver is 455 kc, then the oscillator output should be 1165 kc. Assume that the oscillator frequency drifts to 1180 kc, thus detuning the receiver by +15 kc from its normal if. This new difference frequency (470 kc) is fed to the discriminator which then produces a correction voltage such that the reactance tube undergoes an increase in reactance, thus lowering the oscillator frequency. When properly adjusted, the control voltage produced by the discriminator will restore the oscillator to the exact frequency required by the i-f system of the set.

63. Review Questions

(1) What are the important qualifications of a good superheterodyne local oscillator?
(2) Explain why both amplitude and frequency stability in a local oscillator are important to the good performance of a superheterodyne receiver.
(3) Draw the circuit diagram of a tuned grid oscillator and explain its operation.
(4) Describe the factors that determine the minimum amplification factor permissible for a triode tuned-grid oscillator.
(5) Describe the effect of raising the plate load resistance of a tuned-grid oscillator upon the frequency of oscillation.
(6) Prove that the minimum amplification factor of the tube used in a Hartley oscillator is a function of the ratio of the square of the numbers of turns in the two sections of the tuning inductance.
(7) Explain why an oscillator and mixer having identical tuning elements will not track properly even over a limited frequency range.
(8) Describe two ways that are commonly used to accomplish oscillator-mixer tracking.
(9) Why is a maximum tracking error of 3 kc not considered serious in a broadcast superheterodyne?
(10) What is meant by AFC? What is the usual method used to accomplish it in high-quality FM receivers?

Chapter 8

INTERMEDIATE-FREQUENCY AMPLIFIER TECHNOLOGY

64. General Requirements for I-F Amplifiers

As mentioned briefly in Chapter 4, the function of the i-f amplifier in a superheterodyne is to amplify the difference frequency developed by the local-oscillator-mixer action. The characteristics of the i-f amplifier are determined by the application of the receiver and by the behavior of other sections. Before an i-f amplifier can be designed, it is desirable to specify the requirements as discussed below.

Gain. The minimum satisfactory voltage gain that the i-f amplifier produces is determined by the amplitude of the smallest mixer signal to be amplified, considered in conjunction with the signal level needed at the input to the second detector to insure linear detection. Once this minimum gain is determined, the design procedure generally includes steps that will provide for a substantial surplus of gain so that trouble will not be experienced as tubes age, or as the set strays from perfect alignment.

Bandwidth. The bandwidth and selectivity characteristics of a receiver are usually established by the design of the i-f amplifier. The bandwidth ultimately selected depends upon the application of the receiver, i.e. whether it is to be used in applications where selectivity is the prime consideration and reproduction quality the secondary consideration, or vice versa. In addition, the bandwidth of the i-f amplifier contributes materially to the determination of the signal-to-noise ratio, hence it must be chosen so as to make this ratio as large as possible within the limits set by selectivity considerations.

Noise Figure. Although descriptions of the performance of receivers have been given in terms of signal-to-noise ratio for years, a more recent measure of equipment noise quality that has been given widespread acceptance is the *noise figure* of the equipment. Noise figure represents a direct comparison between the actual noise coming from a circuit and the noise that would be present if the circuit were free of *internal* noise sources. That is:

$$F = \frac{S_i/N_i}{S_o/N_o} \tag{20}$$

in which S_i = available signal input power

S_o = available signal output power

N_i = noise input power (ideal amplifier with zero internal noise)

N_o = actual noise output power

In a superheterodyne receiver, the mixer and succeeding i-f stages are in cascade. It may be shown that the over-all noise figure of two stages in cascade can be expressed by:

$$F_{ab} = F_a + \frac{F_b - 1}{G_a} \tag{21}$$

where F_{ab} = the noise figure for the cascaded pair

F_a = the noise figure of the first stage

F_b = the noise figure of the second stage

G_a = the power gain of the first stage.

The interesting feature of this equation is this: if the gain of the mixer stage (together with any r-f amplifiers that may precede it) is sufficiently great, G_a may grow to proportions that make the second term of Eq. 21 negligible. For this situation, the noise figure of the i-f amplifier, F_b, becomes inconsequential in determining the overall noise figure of the system. Thus, in designing a high quality receiver, considerably more attention should be given to reducing the noise figure of r-f amplifiers and mixer than to the noise figure of the i-f amplifier.

Stability. Except in special cases (e.g., very selective and carefully controlled amateur band receivers), the intermediate-frequency amplifier should not be regenerative. Even when not in an oscillatory state, regeneration can cause the amplifier to distort over large portions of its passband. Care must be exercised in applying automatic volume (or gain) control to i-f amplifiers. Due to the Miller effect, the change in gain caused by avc voltages may result in input capacitance variations that are severe enough to detune the amplifier.

Signal Handling Ability. An i-f amplifier must be capable of handling a wide range of signal amplitudes. It must handle large signals without overloading or producing limiting effects, while at the same time being capable of amplifying weak signals with little noise so that they yield useable output at the second detector. Normally, such a wider range of signal handling ability is achieved by using some form of well designed avc (agc) system.

Intermediate Frequency. Some of the factors that contribute to the final selection of an intermediate frequency were mentioned in Chapter 4. These factors, plus others not heretofore discussed, are summarized below.

Factors that speak for the use of a *low* intermediate frequency are:

(1) Noise figure: the grid induced, shot-effect noise power rises as the square of the frequency. Thus, a low intermediate frequency may be expected to produce less noise from this cause than a high if.[1]

(2) Limit on power gain: due to the effects of cathode lead inductance and electron transit time, the input impedance varies inversely as the square of the frequency. This means that at high frequencies the input impedance may decrease significantly, hence limiting the power gain of the tube.

(3) Bandwidth: a low i-f makes it possible to achieve smaller bandwidths with coils having a given Q. The reason for this is that the bandwidth is equal to the ratio of the resonant frequency to the Q of the circuit." Thus, bandwidth is directly proportional to frequency of resonance.

Factors that favor the use of a high i-f are:

(1) Image rejection: discussed in Chapter 4

(2) Detector design: a high if makes detector design considerably

[1] See *R. F. Amplifiers,* edited by Alexander Schure, pp. 22-29, John F. Rider Publisher, Inc.

Discussed in Section *69.*

simpler. When the if is low, it is possible that some components of the detected modulated signal may begin to approach the if. If this happens, a very sharp-cutoff low pass filter is needed in the detector output to pass the desired modulation components but block the if.

65. Choice of I-F Amplifier Tube

Gain and Bandwidth. In broadcast and communications superheterodynes, where narrow-band operation is generally employed, the transconductance of the tube used is most important in determining the merit of the tube as an amplifier. The higher the g_m is, the higher the gain will be. Thus, for high-gain service, tubes such as the 6AC7 (g_m = 9000 micromhos) and 6CB6 (g_m = 8000 micromhos) are very frequently encountered.

In other types of receivers, where bandwidth as well as gain is a criterion for tube selection, the *gain-bandwidth* product establishes a figure of merit for the particular tube selected for amplifier service. This product is a function of tube constants and circuit wiring and is given by:

$$\text{gain-bandwidth product} = \frac{g_m}{2\pi C_t} \qquad (22)$$

where C_t is the total shunt capacitance in the interstage circuit including tube, socket, and wiring capacitances. Evidently, a high gain-bandwidth product is obtained by selecting a tube with a high transconductance and as small input and output capacitances as possible. In this category are tubes such as the 6AK5, the 6CB6, and the 5840. For example, a typical value for the gain-bandwidth product for a tube such as the 6AC7 is only 89, whereas, under similar conditions, a 6CB6 may have a product of 120[1].

Input Conductance. If the input conductance of an i-f amplifier tube is high, the maximum realizable gain from the preceding stage may be severely limited. Stated otherwise, it is important that the input impedance of an amplifier be as large as possible to permit realization of desired gain in the preceding stage. This means that the permissible grid return resistance must be high, as well as the input capacitive reactance to the tube.

[1] If g_m is in micromhos and C_t is in micromicrofarads, then the g-b product is in megacycles.

Noise. Although the noise figure of the i-f amplifier stage does not contribute significantly to over-all receiver noise (see Section 64) in normal applications, there may be some receiver types in which attention must still be given to this factor. In such cases, the tube must be chosen so that its noise figure is as low as possible. Triodes have the smallest noise-equivalent resistance but may have insufficient gain; in most cases a reasonable compromise is made between gain, gain-bandwidth product, and noise by choosing tube types that offer the best of each. For example, a 6AC7 has a relatively low R_{eq}, about 200 ohms, while still being quite satisfactory with regard to g_m and bandwidth. Tube manufacturers are constantly improving the essential characteristics of tubes for special service, including the successful reduction of noise.

Signal Handling. An i-f amplifier is called upon to handle a relatively wide range of signal amplitudes. To do this without overloading, remote cut-off pentodes and avc circuits are employed. Simple automatic-volume-control circuits are discussed in Sections 34 and 35. The action of a remote cut-off pentode is given in the illustration of Fig. 44.

In a simple avc system, the operating bias applied to the i-f amplifier becomes larger as the signal amplitude increases. When the tube has a sharp cut-off characteristic, the permissible avc voltage must be very low to prevent the type of distortion shown in Fig. 44 (A). The use of remote cut-off types, however, extends the permissible bias considerably without introducing cut-off distortion as before, thus increasing the signal-handling ability of the i-f amplifier stage.

66. I-F Amplifier Circuits

Once the i-f amplifier tube or tubes have been selected, the circuits to be used must be determined from the performance specifications required. In most i-f amplifiers, gain and bandwidth are the principal factors to be considered. Whether single-tuned circuits, double-tuned circuits, or stagger-tuned circuits are employed depends upon the specific requirements. For a complete discussion of various i-f circuits and their characteristics, the reader is referred to two other books of this series.[1] An extensive treatment of coupling in double

[1] *R-F AMPLIFIERS,* and *TRANSFORMERS,* edited by Alexander Schure, John F. Rider Publisher, Inc.

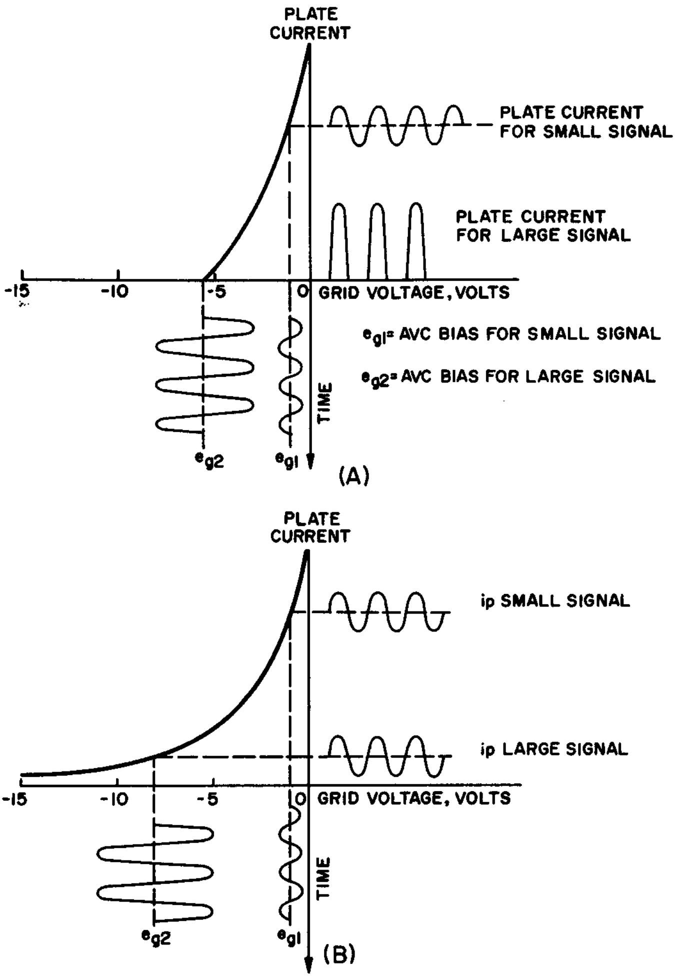

Fig. 44. How distortion due to large signal grid swing is eliminated by using a remote cut-off pentode, in conjunction with AVC. (A) Distortion due to plate current cut-off with sharp cut-off tube; (B) elimination of distortion with remote cut-off tube.

tuned circuits (loose coupling, critical coupling, transitional coupling, and overcoupling) may also be found in the latter of the two books mentioned.

67. I-F Amplification at Ultra-High Frequencies

Ordinary broadcast and communications superheterodynes normally make use of relatively low intermediate frequencies. In these applications ordinary tubes and circuits are capable of providing acceptable performance and trouble-free operation. In uhf television and radar equipment, however, substantially higher intermediate frequencies come into use, bringing with them effects which are either not present, or if present, are not detrimental at the lower frequencies. These phenomena, unique to the ultra-high frequencies, are treated in the following sections.

68. Increased Input Conductance of Vacuum Tubes at UHF

At low frequencies, the grid input circuit of a vacuum tube presents a very high impedance (low conductance) to the i-f signal, provided that the tube is correctly biased and is not overloaded. As the frequency rises, two important phenomena begin to appear and assume significant porportions in the ultra-high-frequency range. Both of these phenomena have the effect of increasing the input conductance of the tube; unless steps are taken to nullify them, the input impedance may drop so low as to load the signal source down to serious proportions.

Electron Transit Time. A cold vacuum tube has a very definite, fixed value of input capacitance. When operated even at medium frequencies as an amplifier, this input capacitance is increased by effects due to electron transit time as follows: the time required for electrons to move from cathode, through the grid, and to the plate is an appreciable fraction of one cycle of the applied grid signal. Electrons in motion represent moving charges; as they approach and retreat from the grid, they induce charges on the grid structure that cause a small current to flow in the external grid impedance. When the input frequency is low, the induced grid current leads the signal voltage by about 90° so that the effect is that of a small added capacitance across the grid input terminals. This, of course, increases the input conductance.

Increase of Transconductance. When the gain of an i-f amplifier is changed either manually or by avc action, the transconductance of the tube is varied. As the g_m rises, a second increase of input capacitance occurs as a result of the movement of the space charge inside the tube toward or away from the grid. The effect is similar to the induction of charges due to electron transit time. Both these phenomena in combination are capable of increasing the input capacitance of a tube by 40% or more. For example, the increment of input capacitance for several common i-f tubes are given below:

	Cold Input Capacitance $\mu\mu f$	Hot Input Capacitance $\mu\mu f$	Increment (c)
6AH6	6.8	10.0	3.2
6CB6	4.5	6.3	1.8
6AU6	2.9	6.5	2.6

Detuning Effect of Added Input Capacitance. The influence of the Miller effect upon i-f tuning was discussed quantitatively in Chapter 4. It was shown that the input capacitance to an i-f amplifier is a function of interelectrode capacitance modified by the Miller effect according to the equation:

$$C_{in} = C_{gf} + (M + 1)\ C_{gp} \tag{23}$$

where C_{in} = the total input capacitance

C_{gf} = grid-filament or grid-cathode capacitance

M = stage gain

C_{gp} = grid-plate capacitance

This equation demonstrates that the input capacitance varies in the same sense as the stage gain; that is, as the gain increases, the input capacitance increases and vice versa. Since stage gain is a direct function of tube transconductance, it is clear that the input capacitance increment due to the Miller effect may be reduced by reducing the transconductance of the tube. The usefulness of this phenomenon is explained in the next paragraph.

As shown previously, the input capacitance is a function of transconductance in a second way — as a result of space charge movement. Thus, any resonant circuit between the grid and cathode will be detuned if the transconductance of the tube is varied. With avc applied to such a stage, the difference in tuning between strong and large signals may be substantial because avc action causes transconductance to vary. The change in input capacitance due to space charge may be minimized, however, by artificially *reducing* the g_m of the tube so that the Miller effect is reduced at the same time. The diminution in "Miller effect capacitance" can be made to compensate for the increase in input capacitance due to space charge movement. This is accomplished by having a part of the cathode bias resistor unbypassed as shown in Fig. 45. The degeneration produced by this measure reduces the g_m sufficiently to counteract

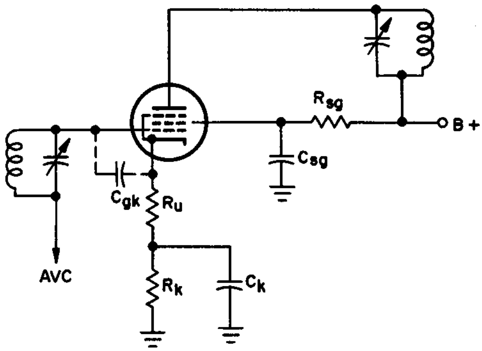

Fig. 45. Making use of reduction of Miller effect to compensate for input capacitance increment due to movement of space charge.

the variations of input capacitance due to space charge movement, provided that the unbypassed portion of the cathode resistor has such a value as to meet the requirements of Eq. 24.

$$R_u = \frac{cI_b}{g_m C_{gk} I_k} \tag{24}$$

in which c = capacitance increment due to space charge, in $\mu\mu$f

I_b = quiescent plate current at operating point, in ma

C_{gk} = grid-cathode capacitance of cold tube, in $\mu\mu$f

I_k = cathode current, quiescent at operating point, in ma

R_u = unbypassed portion of cathode resistance, in ohms

g_m = transconductance, in mhos

Example: Determine the unbypassed value of cathode resistance needed to compensate for the space charge capacitance increment for a 6AU6 i-f amplifier operating under the following conditions:
Quiescent plate current at operating point = 10.8 ma
Quiescent screen grid current at operating point = 4.3 ma
Transconductance = 5200 μmhos
Grid-cathode capacitance = 3.2 $\mu\mu$f

Solution: From the factors given, the known quantities needed for Eq. 24 are:

I_b = 10.8 ma
g_m = 5200 μmhos
c_{gk} = 3.2 $\mu\mu$f
I_k = 15.1 ma
c = 2.6 $\mu\mu$f (from table on page 00)

Substituting in Eq. 24, we have:

$$R_a = \frac{2.6 \times 10.8}{5200 \times 3.2 \times 15.1 \times 10^{-6}}$$
$$= 111 \text{ ohms}$$

The foregoing effects have been described adequately in connection with relatively high radio frequencies. As the frequency at which an i-f amplifier must work is raised even further, the transit time for electrons becomes even more significant in its effect. When the transit time becomes a relatively large fraction of one cycle, the induced grid current exhibits a component that is *in phase* with the applied signal. This in-phase current represents a further increase in the input conductance of the tube; for this condition, input loading may become extremely severe, causing the stage gain to decrease to the point of usefulness. Several investigators[1] have shown that the grid input conductance increases as the *square* of the frequency in the microwave region, necessitating, therefore, special tube designs and structures for operation in these ranges.

69. Effect of Cathode Lead Inductance

At very high frequencies, the straight leads between the tube socket terminals and the leads that connect to the tube elements begin to exhibit inductive reactance. In particular, the inductance of the cathode lead assumes major importance because of the likelihood of feedback from this source. Figure 46 is an equivalent circuit which shows the cathode lead inductance and its physical relationship to the grid-plate and grid-cathode capacitances.

As seen in the diagram, the voltage developed across the inductive reactance of the cathode lead inductance is in series with the applied signal voltage since the cathode current flows up through L_k. Since L_k is a reactive rather than a resistive component, this feedback voltage produces a current component through C_{gk} which is *in phase* with the input voltage. In other words, where a pure resistance

[1] Especially F. B. Llewellyn (1941), Cambridge University Press, London

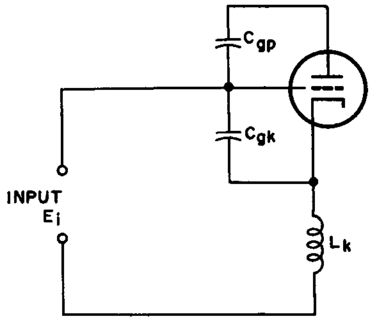

Fig. 46. Equivalent circuit showing cathode lead inductance in relationship to grid-plate and grid-cathode capacitances.

in the cathode circuit produces degeneration, an inductive reactance in combination with the capacitive reactance of C_{gk} produces a regenerative effect which often leads to instability and oscillation, and certainly lowers the input impedance of the tube. Thus, cathode lead inductance reduces the available power gain of the stage.

With reference to maximum power gain of a stage, it is evident that an amplifier is useful only as long as its gain exceeds unity. The input conductance increases as the square of the frequency as a result of electron transit time (see Section 67); it can also be shown that input conductance increases as the square of the frequency as a result of cathode lead inductance as well.[1] This may be interpreted to mean that one of the most important factors which limits the usefulness of a tube as an amplifier at very high frequencies is grid conductance. This is shown in Eq. 25.

$$\text{Gain}_{max} = \frac{g_m^2 \ R_i \ r_p}{4} \tag{25}$$

where g_m = transconductance in mhos

R_i = reciprocal of input conductance in ohms

r_p = output resistance of the stage in ohms.

70. Other UHF Effects

Two additional phenomena limit the frequency for which conventional vacuum tubes are useful. These are (1) grid-induced, shot-effect noise and (2) tube resonance.

[1] F. B. Llewellyn shows that input conductance due to cathode lead inductance $= g_m \times (2\pi f)^2 L_k C_{gk}$

Grid-induced, Shot-effect Noise. Fundamentally, all shot-effect noise originates from the randomness of emission of charged particles from a thermionic cathode. Such emission results in a cathode current having minute but significant variations.

When the electrons that constitute the space current in the tube approach the grid, a tiny grid current consisting of electrons moving out of the grid is induced by the changing electrostatic field in the vicinity of the grid; as the electrons recede from the grid plane (toward the plate), the direction of this induced current changes. Since the space current itself fluctuates due to shot effect, the current induced in the grid circuit varies in proportion. Thus, there is the equivalent of a shot-effect noise input voltage present in the grid circuit. Due to the finite transit time of electrons, the magnitude of the induced noise voltage is a function of frequency and increases linearly with it. Carrying this one step farther, the grid-induced shot effect noise *power* must increase as the *square* of the frequency. The result is that all of these limiting factors considered together (increasing input conductance due to electron transit time, and cathode lead inductance, in addition to shot-effect noise) limit the usefulness of conventional vacuum tubes to frequencies below 500 mc.

Tube Resonance Effects. One additional frequency limitation is imposed by tube resonance effects. As Fig. 46 shows, the grid-cathode capacitance (C_{gk}) is effectively in series with the cathode lead inductance. As the frequency rises, a point is reached where L_k and C_{gk} form a series resonant circuit from the grid to the cathode of the tube. At this point, the input impedance drops virtually to zero, short-circuiting the signal source. Large amplifiers such as the 6AC7 reach the series-resonant frequency much more quickly than smaller types such as the 6AK5. Thus, the resonant frequency for a 6AC7 is about 200 mc while the resonant frequency for a 6AK5 is nearer 500 mc.

71. Review Questions

(1) What is meant by *gain-bandwidth* product? Explain why this is an important factor in appraising the performance of an i-f amplifier.

(2) Define *noise figure*. Explain why the noise figure of an i-f amplifier may not be important in determining total receiver noise if the mixer gain is very high.

(3) Summarize all the factors which make the choice of a low intermediate frequency desirable. Repeat with a summary of factors that encourage the choice of a high if.

(4) Explain why remote cut-off i-f amplifiers are capable of handling wider ranges of signal amplitude than sharp cut-off tubes.
(5) What is meant by electron transit time?
(6) Explain why finite electron transit time increases the input conductance of an i-f amplifier.
(7) Explain how to compensate for the de-tuning of an i-f stage caused by transconductance changes arising from space-charge movement.
(8) What is the effect upon input conductance produced by cathode lead inductance? Explain.
(9) Why does reduced input resistance to a vacuum tube (increased conductance) lower the available power gain?
(10) Explain how grid-induced, shot-effect noise comes into existence. What is the effect of this noise on the upper limit of usefulness of the tube?

Chapter 9

TRANSISTORIZED CONVERTERS AND I-F AMPLIFIERS

72. General Information About Converters

No book dealing with superheterodyne converters and i-f amplifiers would be complete without discussing the transistorized counterparts of the vacuum tube versions of these stages. Although most of what has already been said about converters, mixers, and i-f amplifiers of the tube variety applies to transistorized stages as well, certain important differences must be emphasized. These differences arise from several sources; among these are input and output impedances, the relative sensitivity of transistors to temperature variations, and the fact that transistors are basically current amplifiers rather than voltage amplifiers.

As would be expected, these differences in behavior give rise to many circuit modifications; in addition the design philosophy used for transistorized converters, mixers, and i-f amplifiers is completely at variance with the corresponding procedures used for vacuum tube stages having similar functions.

The traditional difference between a "mixer" and a "converter", first established for vacuum tubes, is carried into transistor circuits without change. That is, if the local oscillator is part of the mixer, the device is called a converter. The use of a single transistor as both mixer and oscillator is economical, and is therefore popular in broadcast receivers. Although the performance is adequate at these frequencies, there are definite disadvantages to combining both functions in a single semiconductor. Among these are the limitation in the maximum signal voltage that can be applied to the mixer because full agc canot be used. If an attempt is made to utilize the entire available agc voltage for a strong signal, the oscillator would tend to discontinue operation somewhere in the agc range. In addition to this effect, at the higher frequencies difficulties are encountered with the variation of local oscillator frequency with agc voltage, particularly in narrow-band systems.

73. Basic Design of a Transistor Converter

The design of a transistor converter is based upon the same factors that enter into the design of an r-f amplifier at the same frequency.[1] The input impedance of the mixer will be essentially the same as that of the transistor itself at the quiescent bias point. An important point of departure, however, is that the effects of normal feedback parameters which might cause oscillation may be ignored in the design of a mixer because the collector (output circuit generally) is tuned to a different frequency from the input circuit, making oscillation impossible, or at least highly unlikely.

Other design considerations are presented below. The approximate effects and values, although initially calculated by transistor engineers, have been thoroughly verified by experiment.

Conversion Gain. The conversion gain of a transistorized converter is roughly 6 to 8 db less than the voltage gain of the same transistor in an idealized, straightforward amplifier circuit.

Conversion Conductance. If the emitter current is 100% modulated by the local oscillator signal, the conversion conductance is very closely ten times the static emitter current. That is:

$$\text{Conversion Conductance} = 10\ I_E$$

where the conductance is measured in mhos and the emitter current is in amperes.

Output Impedance. The i-f output impedance of a transistorized converter closely approximates the output impedance of the same transistor when used as an amplifier with shorted input connections.

Noise Figure. The noise figure of a converter is from 6 to 10 db higher than it is when the same transistor is used as an amplifier. This is particularly true at the lower intermediate frequencies (455

[1] *R. F. Amplifiers,* edited by Alexander Schure, John F. Rider, Publisher, Inc.

kc). At the higher frequencies, the performance of the transistor as an amplifier deteriorates to the point where noise figures are very much the same for either the amplifier or mixer application.

Local Oscillator Voltage. Transistor converters display a rather fortunate characteristic: there is a wide range of local oscillator voltages over which the rate of change of gain is relatively small. Thus, the injection voltage is not highly critical. The usual value ranges from 35 to 100 millivolts. Excessive injection voltages tend to raise the noise figure without appreciably improving the conversion gain.

Injection Points. The local oscillator voltage is usually applied between emitter and ground while the input signal is applied between base and ground.

74. Autodyne Converter Circuit

The autodyne converter is designed for efficient and economical broadcast-band operation using only one transistor. The incoming r-f signal is applied between the base and ground; the i-f output is taken between the collector and ground. The complete circuit is illustrated in Fig. 47.

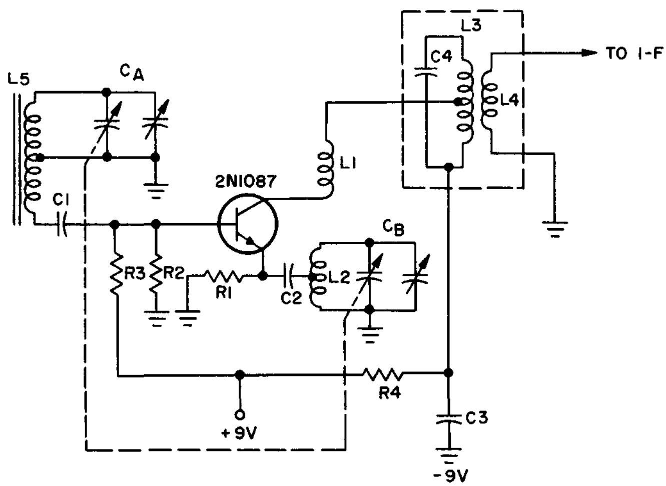

Fig. 47. An autodyne converter which combines in a single transistor stage the functions of oscillator, mixer, and i-f amplifier.

In order to simplify the explanation of the oscillator circuit action, those components which play a part in the sustaining of local oscillations have been redrawn in their proper relationships in Fig. 48.

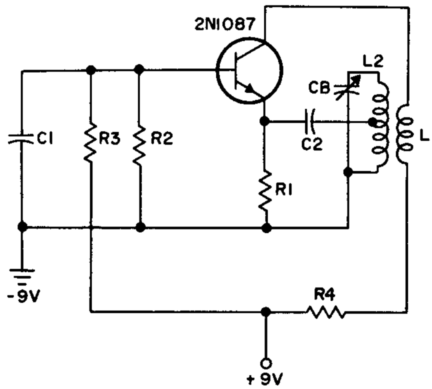

Fig. 48. The oscillator section of the autodyne converter pictured in Fig. 47. Only those components essential to sustained oscillation have been included in the drawing.

Like most oscillators, the process of building up oscillations is begun by a random variation in the base current. This variation is amplified in the collector circuit, the amplified variation causing a current flow in coil L1 which serves as the primary of a transformer. The secondary winding, L2, is tuned to resonance at the desired oscillator frequency by C_B so that a voltage drop at this frequency appears across the terminals of L2 by induction from L1. A small part of this signal is fed back via the tap in L2, through C2 to the emitter of the transistor. With proper phasing of L2 and L1, the feedback will be regenerative and sustained oscillation will result. The tap in L2 serves another very important purpose: it matches the relatively high impedance of the tuned coil (L2) to the low impedance of the emitter circuit. R2 and R3 form a voltage divider that provides the proper d-c bias for the transistor, while R1 is an emitter stabilizing resistor. Capacitor C1 effectively bypasses the biasing voltage divider resistors to ground; hence, the base is maintained essentially at ground potential with respect to the signal voltages in the circuit. From this it is clear that the oscillator operates in the grounded-base configuration; the circuit actually is a typical, two-winding feedback (tickler) oscillator comparable to the familiar Armstrong circuit used with vacuum tubes.

The mixer section of the autodyne converter has been redrawn in Fig. 49. Those components common to both oscillator and mixer functions have been included in this schematic.

L5 is a high-Q ferrite-rod antenna which is tuned to the desired incoming radio frequency by C_A. The signal voltage is applied between base and ground via the tap on L5; here again the coil is tapped to establish an impedance match between the high impedance resonant circuit, and the relatively low base input impedance. Resistors R3 and R2 are so chosen as to bias the transistor in a

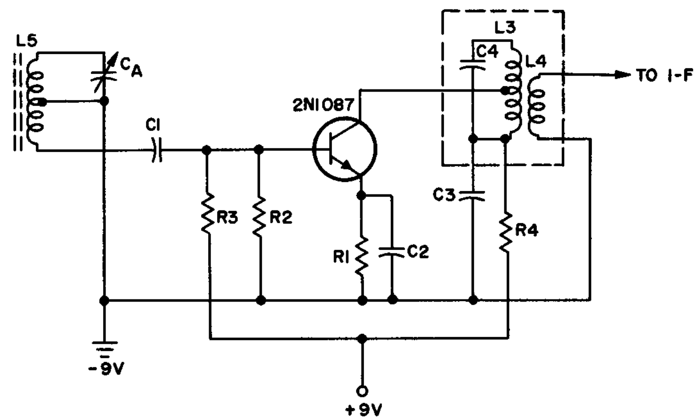

Fig. 49. The mixer section of the autodyne converter redrawn to show the components that play a part in mixer function. Note that some components are common to both oscillator and mixer actions.

relatively low current region. Under these conditions, the transistor exhibits nonlinear properties required for successful mixing. Since the base circuit already carries the local oscillator current, mixing between this signal and the incoming signal occurs in the nonlinear transistor to produce the usual sum and difference heterodyne frequencies. As in a vacuum tube superheterodyne, L3 serves as an output load impedance tuned to the difference frequency (if). R4 and C3 make up a decoupling filter which prevents undesirable feedback from collector to base circuit; such feedback, if positive, would lead to instability and, if negative, would lead to loss of gain. Since the emitter is grounded through C2 and the incoming signal is injected into the base circuit, the mixer portion of the autodyne converter may be classified as a grounded emitter type.

75. VHF Mixer and Oscillator Using Two Transistors

At the very high frequencies, as in television transmission and reception, an autodyne converter is generally unsatisfactory. High frequency tetrode transistors having relatively low noise figures, high gain at these frequencies, and good stability can be used very effectively. Figure 50 shows a portion of a television tuner using such transistors; normally, the mixer stage in a typical tuner would be preceded by an r-f amplifier (not shown in Fig. 50) to obtain

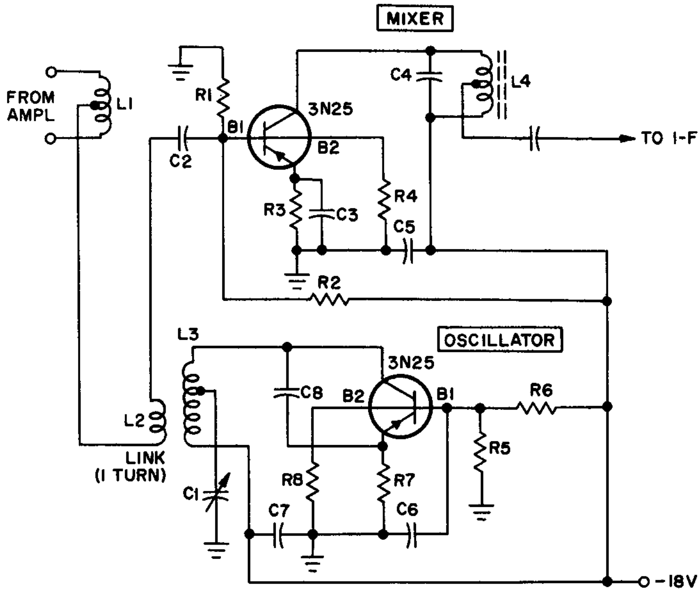

Fig. 50. VHF Mixer using a separate oscillator transistor. The r-f amplifier that would normally precede the mixer in a television tuner has not been shown in this diagram. (Courtesy Texas Instruments, Inc.)

increased selectivity and decreased oscillator radiation. It is interesting to note that the r-f amplifier degrades rather than improves the over-all noise figure of the system, yet is included for the reasons given above. The additional gain is incidental because it could be obtained more readily and economically at the i-f level.

Analysis of this circuit might begin with the d-c biasing method used for both stages. Base 1 (B1) in each transistor is biased by a voltage divided network of its own. R1 and R2 develop fixed bias for the mixer; R5 and R6 perform this function for the oscillator.

A unique coupling method is used to transfer the r-f energy and oscillator signal to B1 of the mixer: the incoming r-f feeds through L2 and C2, to B1 of the mixer tetrode while the oscillator energy is coupled inductively via the same single turn link (L2). The link coupling method was used because it is easy to apply and also because injection of the local oscillator signal at the base requires slightly less power than coupling at the emitter, the more usual method. The oscillator is a simple tuned-collector type. The feedback required for sustained oscillation occurs through the transistor internal impedance and a very small external capacitance (C8 = 1.0 $\mu\mu$f). Base 2 (B2) bias is established for the oscillator by R8, and for the mixer by R4.

Conversion gain ranges from 4 to 10 db with various types of 3N25 transistors, while the average noise figure is about 11 db over the entire television band. The conversion gain remains flat over the oscillator voltage range from 0.07 to 0.1 volt. Since the input impedance of the mixer is about 125 ohms, the local oscillator is called upon to deliver power in the order of 40 microwatts. By actual measurement, the oscillator can furnish 700 microwatts of power at 257 mc with a 15-volt battery supply, hence the conversion gain is maintained over the entire range of tuning.

76. Transistor I-F Amplifiers

All three transistor configurations (i.e., common or grounded base, common emitter, common collector) have been used in i-f amplifier circuits. Regardless of configuration, however, it has been shown that transistor i-f amplifiers tend to be unstable somewhere in their frequency range. In earlier i-f designs, a certain degree of stability was achieved by intentionally inserting losses. This procedure, however, does not eliminate feedback. With feedback present, the input impedance of the transistor depends upon source impedance to a significant extent; similarly, the output impedance is a function of the load impedance. Both of these conditions are undesirable because they may lead to difficulty in aligning tuned circuits, or because they may distort bandpass characteristics.

The process of making feedback parameters equal zero in high frequency transistorized amplifiers is generally known as *unilater-*

alization. Neutralization, on the other hand, is a similar process carried only far enough to remove the instability responsible for oscillation of an amplifier. While unilateralization is an idealized approach, neutralization is a practical one and is generally used. Many circuit arrangements have been proposed for neutralizing i-f amplifiers[1]; these will not be discussed in detail here since they are available to the interested designer in the original literature.

77. One Stage I-F Amplifier with Neutralization

The single stage i-f amplifier shown in Fig. 51 embodies all of the principles of good design in the present state of the art. Utilizing one transistor between a single-tuned output transformer and a double-tuned input transformer, this circuit provides an approximate usable gain of 37 db.

The input transformer is tapped down to provide the proper load impedance for the mixer output; the secondary winding of the

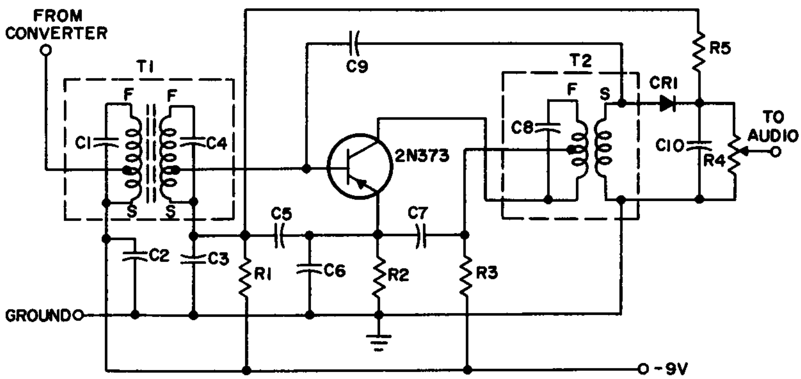

Fig. 51. Modern single stage i-f amplifier intended for use on 455 kc for broadcast radios. This circuit also includes a simple AGC system described in text. Crystal rectifier (CR1) is the detector, shown here to illustrate the AGC take-off. (Courtesy of RCA)

[1] A. P. Stern, "Considerations on the Stability of Active Elements and Applications to Transistors," 1956 IRE Convention Record, Part 2, *Circuit Theory,* pp. 46-52.

G. Y. Chu, "Unilateralization of Junction Transistor Amplifiers at High Frequencies", *Proc. I.R.E.,* vol. 43, August 1955.

A. J. Cote, Jr., "Transistor Neutralizer Networks", *I.R.E. Trans. on Circuit Theory,* vol. CT-5, June 1958.

input transformer (T1) is also tapped, but the tap is positioned to provide an intentional *mismatch* to the input of the i-f transistor. As mentioned in the previous section, mismatches of this kind are introduced for the purpose of producing stability and, although this i-f amplifier is neutralized, a mismatch is used as well to prevent oscillation.

The design T2 is such that the tapped primary winding provides the proper load impedance for the output of the transistor collector circuit, and in turn provides additional mismatch considered necessary for stability. The total mismatch turns out to be about 16.5 db resulting in a usable gain of about 37 db. A 180°-phase shift is produced by T2 by reversing the connections to the secondary winding (note: S and F symbolize start and finish of windings, respectively). This phase shift is required to produce neutralization through capacitor C9. This capacitor couples a voltage equal and opposite in phase to the signal that is inherently coupled from output to input through the internal feedback capacitance of the transistor, thus achieving sufficient neutralization to prevent oscillation.

The agc system is simple and straightforward. When no signal is applied to the amplifier, the base of the transistor, and the junction of R4 and R5 are negative. At this time, the emitter current has a quiescent value of about 1 ma. When the signal appears at the diode, the junction of R4 and R5 goes positive. This causes the base to become more positive, increasing the bias. The gain of the stage thus decreases. Within relatively wide limits of signal voltage, the stage gain varies linearly with the agc voltage developed at the output of the crystal detector.

78. Emitter Tuned I-F Amplifier

The availability of complementary transistor types (n-p-n and p-n-p) permits the design of unusual transistor circuits. One very useful deviation from the normal tuned i-f system is illustrated in Fig. 52. Its selectivity is realized by using series resonant networks placed in the emitter circuit of each of the stages. Note the direct coupling permitted by the complementary nature of the transistors and the absence of coupling transformers.

Although the overall gain is less than that of conventional i-f strips using the same number of transistors, this circuit arrangement

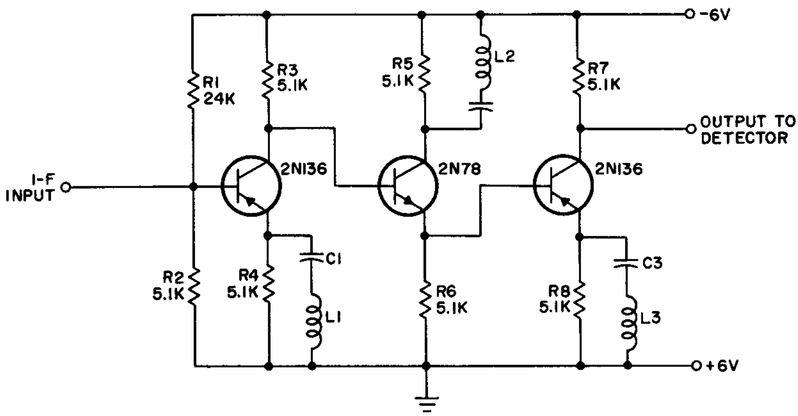

Fig. 52. Emitter tuned i-f amplifier utilizing complementary transistors and direct coupling. Although the total gain is somewhat less than a conventional i-f system, the circuit permits ready transistor interchange without serious effect. (Courtesy of General Electric Company)

has several desirable features. One of the most important is the relative independence of tuning and alignment with respect to transistor replacements; alignment seldom needs changing even when replacement transistors differ widely from the originals. The alignment procedure is extremely simple, particularly when narrow band operation is desired. Finally, the wiring of the circuit is far simpler than the conventional i-f amplifier containing the same number of transistors.

Neutralization is completely unnecessary at normal intermediate frequencies because there is sufficient mismatch between the output of one transistor and the input of the next to provide good stability.

One attractive feature of the circuit is the reduction in the number or different component values — a good recommendation for commercial assembly. Note that all the resistors (except R1) are 5.1K ohms. Furthermore, depending upon the if used, all the coils and capacitors have the same values, respectively, thus reducing the chance for assembly line errors significantly.

79. 30 MC I-F Amplifier Strip

The i-f amplifier shown in Fig. 53 is an excellent example of the simplicity that can be obtained, even in a relatively high frequency amplifier, with modern transistors having high cut-off frequencies.

The circuit philosophy is novel in several ways: (1) synchronous, single-tuned transformers are used throughout; (2) the base of each transistor is at d-c ground potential and separate batteries are used for collector and emitter, allowing excellent bias stability with simple circuitry; (3) the required bandwidth is obtained by resistive loading of the tank circuits which, at the same time, reduces the

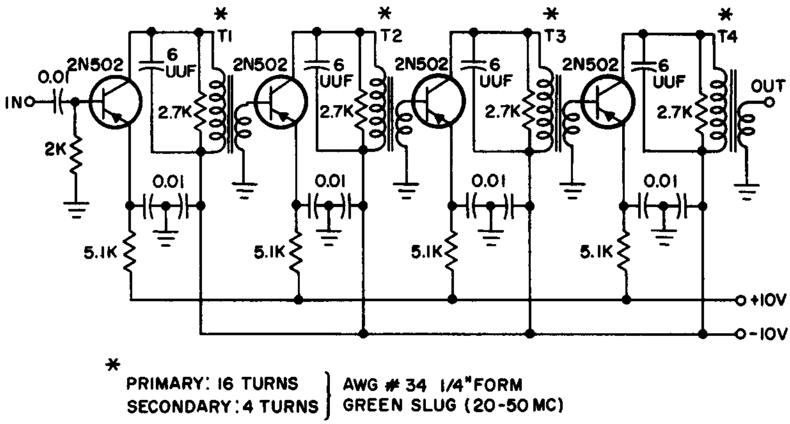

Fig. 53. A 30-mc, 4-stage, synchronously tuned i-f amplifier having relatively wide bandpass. Note the absence of neutralizing networks; these are made unnecessary by resistive loading of the tuning coils. (Courtesy of General Electric Company)

individual stage gains so that instability is overcome, making neutralizing unnecessary. Bandwidths in transistor amplifiers where the response has been broadened by overcoupling or stagger-tuning tends to vary as a function of temperature and transistor operating points. Resistive loading reduces this type of variation, although gain is reduced at the same time. In many applications, the advantages of resistive loading more than outweigh the disadvantage of the need for one or more additional amplifier stages to compensate for the loss of gain.

Another rather unique feature of this circuit is the use of fixed capacitor loading across all the tank coils. Most amplifiers of this type utilize the transfer and stray capacitances to establish resonance at these high frequencies. The use of capacitor loading reduces the sensitivity of the amplifier to bias and temperature drift, generally stabilizing the center frequency to within $\pm 2\%$.

A gain of 84 db is easily obtained with this amplifier over the temperature range of $-40°$C. to $+65°$C.

80. Review Questions

(1) Why are single-transistor converters used even though they are not especially efficient?
(2) What specific advantages do separate mixer-oscillator combinations have over single-transistor converters?
(3) Analyze the operation of the autodyne converter shown in Fig. 47, using "break-down" diagrams to help you.
(4) Explain why the mixer section of a converter does not need to be neutralized to prevent oscillation.
(5) Why is the autodyne transistor biased in a relatively low current region? Explain this point carefully.
(6) What characteristics are common to modern tetrode transistors that make them suitable for use as high-frequency amplifiers?
(7) In the VHF mixer of Figure 50, explain why an i-f amplifier is included even though it degrades the noise figure of the system.
(8) Describe the phase relationships present in the i-f amplifier of Figure 51 which allow capacitive neutralization.
(9) Analyze the operation of the circuit in Fig. 52.
(10) State the advantages of resistive loading to achieve bandpass in a circuit such as that given in Fig. 53.

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