HOW R. F. AMPLIFIERS WORK IN AM, FM, AND TV RECEIVERS

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For each study step, read the assigned pages first at your usual speed. Reread slowly one or more times. Finish with one more quick reading to fix the important facts firmly in your mind, then answer the Lesson Questions for that step. Study each other step in this same way.

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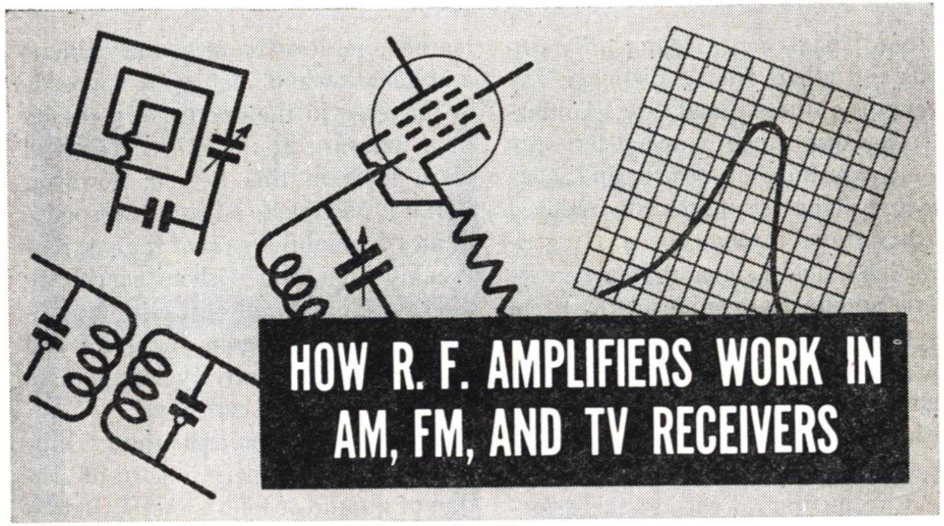


HOW R. F. AMPLIFIERS WORK IN AM, FM, AND TV RECEIVERS

MODERN radio and television receivers all contain r.f. amplifiers, as your earlier Lessons have indicated. These important radio stages have a double job: first, they select the desired signal from among all those picked up by the antenna; and second, they provide most of the amplification the signal gets in the receiver. In this Lesson, you are going to learn just what r.f. amplifiers are and how they work. As a starter, let's see first why they are needed.

Radio signals travel through space in the form of a radio-frequency "carrier" that has been modulated by an intelligence signal. The *radio frequencies,* so called because they can be "radioed" or radiated easily, begin at about 10,000 cycles and extend upward to frequencies close to infra-red and light rays. (As a matter of fact, light and other associated radiations are all part of the same family of electromagnetic waves, differing only in frequency.)

The radio-frequency spectrum is cut up into "bands" that are assigned to different services by the Federal Communications Commission (FCC). The "broadcast" band from 540 kc. to 1600 kc. is one example; other bands are assigned to police, fire, aircraft, trans-atlantic telephone, telegraph, and television services, to name just a few. Within each band, *channels* are assigned to stations; each channel contains at least one modulated carrier. These channels adjoin one another, so we have signals coming to us from adjacent channels. The same channel is frequently assigned to more than one station; in this case, the various stations must share time or be so located as not to interfere excessively with each other within their own service areas.

At the receiving antenna location, signals are present from all kinds of stations, many from bands in which we may have no interest. The average receiving antenna feeds signals from many different stations into the receiving set, all at the same time. Some of these signals come from nearby stations and others from a distance. The strengths of these signals are all different, because some transmitters are more powerful than others, and, also, because signal strengths are reduced as they travel through space. (In general, the induced signal voltages in receiver antennas are quite small.) From all these many signals, we want to pick out *one* at a time, and want to hear or see only the desired program, without interference from other programs.

As we mentioned earlier, this *selection* of the desired signal is the job of the r.f. amplifier section of a radio or television receiver. In addition, the r.f. amplifier has the important task of giving the tiny signals induced in the antenna most of the *amplification* they receive in the set.

Perhaps you wonder why the r.f. amplifier should be called on to furnish most of the amplification when all sets contain low-frequency amplifiers that are capable of building up the signal. There are two chief reasons. The first, as you will learn in a later Lesson, is that it is highly desirable from the standpoint of fidelity to have the signal very strong before the modulation is removed from the carrier by the demodulator or detector stage (the second detector in a superheterodyne). Of course, once the desired signal is selected, we want an acceptable reproduction of the original sound or scene, so we are interested in the fidelity of reproduction.

The second reason for preferring amplification at the radio frequency to low-frequency amplification also has to do with fidelity. As you have learned, in low-frequency amplifiers, amplification and fidelity are opposed: an increase in the one usually results in a decrease in the other. You will learn later in this Lesson, however, that it is possible to adjust the response of an r.f. amplifier so that it gives reasonably high gain without sacrificing fidelity. That is why all modern receivers have high gain in the r.f. end of the set and relatively low gain in the low-frequency end.

You can see, then, that the r.f. amplifier is an extremely important section of a radio or television set. Before we go any further into the subject, let's see what a basic r.f. amplifier consists of.

A BASIC R.F. STAGE

As shown in Fig. 1, an r.f. stage has a signal source, an amplifier tube supplied with operating voltages, and a "load" in the plate circuit, exactly like a basic low-frequency amplifier. As a

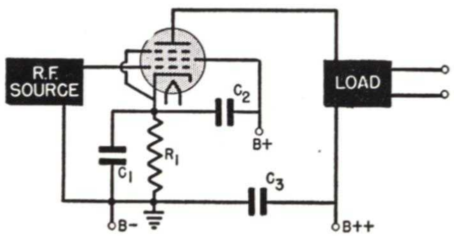


FIG. 1. The basic r.f. stage has the same components as a low-frequency amplifier except that tuned circuits are used as the load on the coupling to the next stage.

matter of fact, the only real differences between this amplifier and the low-frequency amplifier lie in the kinds of loads and couplings used. Transformers with solid or laminated iron cores cannot be used in r.f. stages, and the resistance coupling that we are familiar with in low-frequency amplifiers provides very little amplification at radio frequencies because of shunting tube capacities. Instead, L-C resonant circuits are generally used in r.f. stages, either as the load or as a portion of the coupling network between stages. Let's see why such circuits are desirable for this use.

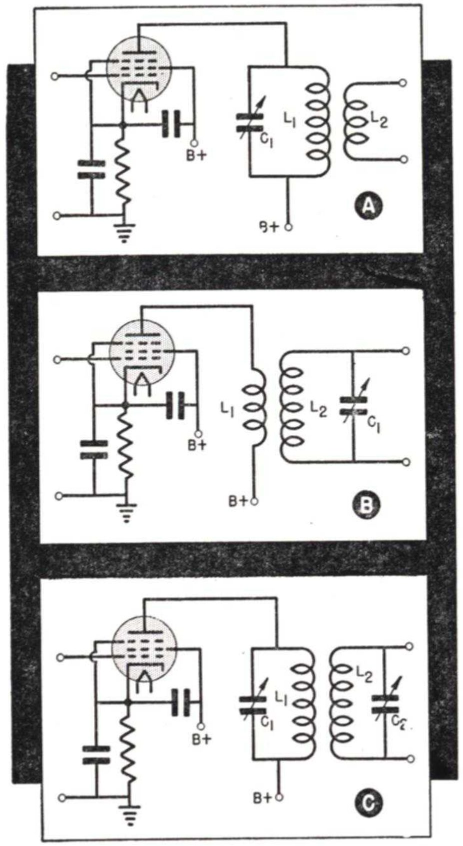


FIG. 2. The three basic tuned-circuit connections.

Fig. 2 shows three basic resonant circuit connections. In Fig. 2A, resonant circuit L_1-C_1 is the "load" for the tube. From your earlier study of parallel-resonant circuits, you know that this circuit will be resonant to some particular frequency that depends on the L-C values.

If we adjust either the capacity or the inductance of the resonant circuit so that its resonant frequency is the same as the frequency of the signal we want to receive, the parallel-resonant circuit will offer maximum impedance and hence the best load for that signal. Since it will offer far less impedance at other frequencies, the stage will effectively have a variable load—high at the desired frequency and low at other frequencies. The gain of the stage will therefore be greatest at the frequency we want; in other words, the stage *selects* one signal by amplifying it more than any other. We can change this resonant point at will by varying the inductance or capacity with our tuning control.

The resonant circuit in Fig. 2B is coupled through an r.f. transformer to the plate of the tube. It acts much like the circuit in Fig. 2A, because its impedance is reflected into the plate circuit in such a way that the maximum load is offered at the resonant frequency. You will learn more about this action later.

A combination of the two circuits is shown in Fig. 2C. Here the plate load is a parallel-resonant circuit that is coupled to another tuning circuit. This arrangement can be made to give greater selectivity at some sacrifice in gain.

Since the operation of an r.f. amplifier stage is quite similar in most respects to that of the stages you have already studied, we shall devote most of our attention in this Lesson to the actions that are different. Specifically, we shall learn how it is possible to make a resonant circuit accept a band of frequencies and reject others to a sufficient degree to enable us to separate signals that are on adjacent channels, yet at the same time not destroy the fidelity of the desired signal. In addition, we shall see in this Lesson why the triode tube can be successfully used as an r.f. amplifier only in certain special circuits. (As a result of this fact, screen-grid or pentode tubes are ordinarily used in the r.f. section of a receiver.)

Selectivity, Gain, and Fidelity

We mentioned earlier that amplification and fidelity are opposed, since an increase in the one usually means a decrease in the other. The same thing is true of selectivity and fidelity, at least in commercial radios; a highly selective set has rather poor fidelity, and a set that has good fidelity does not have much selectivity. (As we shall point out later, it is possible to achieve good selectivity and high fidelity by using elaborate and expensive circuits, but no ordinary radio made today has these features. Television sets do, however.)

Before you can understand why selectivity and fidelity oppose one another, we must amend one of the basic ideas about the transmission of radio signals that you learned in earlier Lessons. For simplicity, we have spoken of the modulated carrier signal up to now as if it were just one signal that has been modulated in amplitude or frequency. Actually, however, the modulating process produces a whole series of "side-band" frequencies—so called because they extend through frequency bands on either side of the carrier frequency. Each of the frequencies in these side bands must be accepted by the receiver and amplified to the same degree if we are to recapture the original sound or scene completely.

Naturally, then, a highly selective circuit—one that emphasizes one frequency far more than any other—is incapable of truly high-fidelity reproduction, because it is unable to respond equally to all the frequencies in the side bands.

Before we carry this subject any further, let's see just how and why the side bands are produced during modulation.

Amplitude Modulation. Amplitude modulation is used in a.m. sound receivers and for the picture portion of a television (TV) signal. Up to now we have merely considered an amplitude-modulated wave to be a carrier signal that varies in amplitude. Going a little deeper, we find that what actually happens is that the process of amplitude modulation produces two new r.f. frequencies (in addition to the carrier) for *each* frequency in the modulating signal.

For example, let's assume that the carrier frequency is 900 kc. If we modulate this with a 1000-cycle audio tone, the transmitter will not only radiate its 900-kc. carrier, but will in addition radiate two new frequencies; one of these will be equal to the *sum* of the carrier and the modulating frequency, the other will be equal to the *difference* between these two. In other words, when our 1000-cycle signal (equal to a 1-kc. signal) is combined with 900 kc., we will get frequencies of 899 kc. and 901 kc. in addition to the original 900 kc.

Other modulation frequencies combine in the same manner. If the modulating frequency is 500 cycles (.5 kc.) and the carrier is 900 kc., we will have side-band frequencies of 899.5 kc. and 900.5 kc.; if the modulating frequency is 4000 cycles, the side frequencies will be 896 kc. and 904 kc.; and so on. If the modulation is a complex tone having many frequencies, there will be a pair of side frequencies for each component of the modulating voltage; together, these side frequencies will make up two bands of frequencies, one on each side of the carrier.

These new additional frequencies travel along with the carrier, and all must pass through the r.f. stages to the demodulator if we are to obtain the original intelligence without distortion. Therefore, our resonant circuits must pass these frequencies as well as the carrier—in other words, they must accommodate a narrow band of frequencies instead of just a single carrier frequency.

Just how wide a band is necessary for amplitude-modulation systems? Broadcast stations in the 540- to 1500-kc. band are assigned carrier frequencies separated by 10 kc. (Some of those between 1500 kc. and 1600 kc. have channels 20-kc. wide assigned to them.) Each station is therefore considered to have a 10-kc. "channel" that extends 5 kc. on either side of the assigned carrier frequency. In other words, it is possible to modulate the carrier by intelligence frequencies up to 5000 cycles without overlapping a neighboring channel. For example, with a 900-kc. carrier, a 5000-cycle signal will produce side-band frequencies of 895 kc. and 905 kc. A 5000-cycle signal on a 910-kc. channel will have side frequencies of 905 kc. and 915 kc. The 905-kc. side band of this channel just coincides with the 905-kc. side band of the 900-kc. channel; if the two stations were modulated by any frequency higher than 5000 cycles, their side frequencies would overlap and create interference in any receiver that picked up both of them. In general, therefore, a.m. sound stations limit their modulation to require a band of only 5000 cycles on each side, or a total band width of 10 kc. in the broadcast service. (Some few stations in this band are so located that they can and do modulate up to 7500 cycles without undue interference.)

Television channels are assigned so that it is possible to modulate with video signals as high as 4,000,000 cycles. This is made possible by the wider bands assigned, and by the partial suppression of one side band, as we shall learn elsewhere.

Summing up: An amplitude modulation (a.m.) system produces a radiated signal that includes not only the carrier frequency but also, for each modulating frequency, two new frequencies differing from the carrier by the amount of the intelligence frequency. In other words, the transmitted signal is a combination of these side r.f. frequencies and the carrier. The interaction between these new frequencies and the carrier produces the varying-amplitude signal that we know as an amplitude modulation wave. For best fidelity, these new frequencies (known as side-band frequencies because they are on both sides of the carrier) should be handled without attenuation by the radio-frequency amplifier. Fidelity will be sacrificed if the r.f. amplifier reduces or cuts off some of the side-band frequencies.

Frequency Modulation. The system known as frequency modulation is used in the f.m. broadcast band, and is also the system used for sending the sound portion of a television signal. To describe f.m. simply, we can say that an f.m. signal is produced by varying the frequency of the carrier signal in accordance with amplitude changes in the modulating signal. The production of side-band frequencies as a result of this modulation is difficult to describe, so we shall leave the discussion of it for other Lessons. However, you can see that a band of frequencies must be transmitted just from the fact that the f.m. signal varies up and down in frequency from the "resting" frequency (which corresponds to the carrier frequency in a.m.). Therefore, if we are going to get the complete volume range of the original signal, the r.f. amplifier must pass all the frequencies associated with a given carrier within the assigned frequency band. In the f.m. broadcast band (88 to 108 megacycles), signals are allowed to vary 75 kc. on either side of the resting frequency. Hence, at least a 150-kc. band is required under the present standards of f.m. operation, and a band this wide must be passed by the r.f. section of a receiver. Actually, an additional 50 kc. is allowed as a "guard" band to avoid interference from adjacent channels as much as possible; therefore, a total band of 200 kc. is allowed in the f.m. broadcast services. (The television sound channel is allowed to vary only 25 kc. on either side of the resting frequency, making a total band width of 50 kc.)

Basically, therefore, whether we have a.m. or f.m. operation, a *band* of frequencies must be passed without attenuation if we are to have good fidelity. On the other hand, all frequencies outside this band must be cut off as sharply as possible if we are to prevent a station on an adjacent channel from creating interference. Hence, the ideal selectivity curve of the r.f. section of a receiver would have the shape shown in Fig. 3. A circuit having this response would give a maximum and entirely uniform gain to the desired carrier and all its side-band frequencies but would cut off very sharply so as to reject signals from neighboring channels completely. In other words, any frequencies within the shaded portion of Fig. 3 would be passed equally, but any frequency outside this band would be cut off altogether. The width of the "pass" band needed for a particular set would depend on the service—it would be narrowest for a.m. broadcasting and widest for television, but in each case the ideal pass-band shape would be

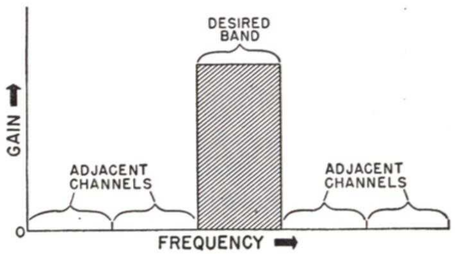


FIG. 3. The ideal selectivity characteristic of an r.f. stage.

approximately the same. Let's turn now to practical amplifiers and see why we cannot get exactly this ideal shape, and what variations we can expect.

RESPONSE CURVES

We can find out how a resonant circuit in an r.f. stage will respond to a band of frequencies by determining the output for different frequencies fed into the amplifier from a standard signal generator. A typical laboratory test set-up is shown in Fig. 4. Let's assume the resonant circuit L_1-C_1 is tuned to 1000 kc.

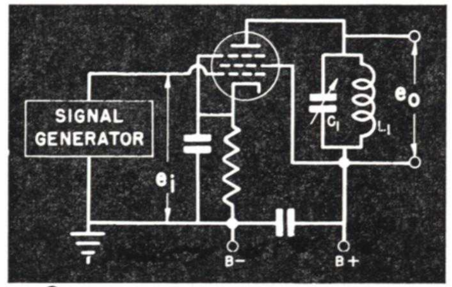


FIG. 4. A test set-up for determining the response of an r.f. stage.

Starting at some frequency lower than the resonant frequency of the tuned circuit—980 kc., for instance—we first adjust the input voltage e_i to some fixed value, say 1 volt. We then read the output voltage e_o for this condition. *Keeping the input voltage* e_i *constant at 1 volt,* we next increase the frequency of the signal generator to some higher frequency such as 985 kc., and read the output voltage e_o again. This process is repeated over and over again, increasing the signal frequency in 5-kc. steps each time, until we have passed the resonant frequency of the tuned circuit (1000 kc.) and gone about 20 kc. beyond (that is, to approximately 1020 kc.).

By plotting the values of output voltage e_o that we obtain for each different input frequency, we may find we have an amplifier response curve like the one in Fig. 5. As we expect, at the resonant frequency of 1000 kc., where the impedance of the tuned circuit is highest, we have considerable gain. In this example, the output voltage e_o at 1000 kc. is 60 volts, or 60 times as large as the input voltage e_i of 1 volt; therefore the over-all gain is actually 60.

For other frequencies that are higher or lower than resonance, the impedance of the tuned circuit drops so that the tube cannot work efficiently, and the gain, consequently, is dropped to smaller values. For example, the gain at 980 kc. and at 1020 kc. is only about 4 and 10, respectively. Hence, this stage favors the frequency to which its L-C circuit is tuned by giving it greater amplification than others. Since the capacity of the L-C circuit is adjustable, it is possible to tune the circuit

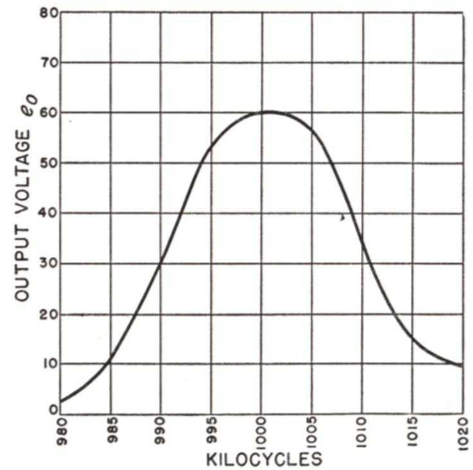


FIG. 5. A typical response curve.

to any desired signal within the range of adjustment of the condenser. (Thus, we could make the circuit have its maximum gain at, say, 1020 kc., in which case the gain at 1000 kc. and at 1040 kc. would be low and that at 980 kc. would be practically zero.)

Now we have a tunable r.f. amplifier that amplifies some frequencies considerably more than others. It can, therefore, act as the input stage of a receiver, because it will select a group of frequencies out of all those induced in an antenna. To know how well it will act as an input stage, we must see how many frequencies it will accept and how evenly it will amplify them.

You can see that the response curve of this circuit does not have the straight sides of the ideal in Fig. 3, nor does it have the flat top. Hence, we haven't ideal selectivity, and the rounded top indicates that all frequencies in the pass band are not amplified equally. Before we can determine how bad the selectivity and fidelity are, we must determine what we can consider the pass band to be on a curve of this shape.

Because of the peculiar hearing characteristics of human ears, the power in a signal may be reduced as much as ½ before the change in level is definitely noticeable. Therefore, all of the frequencies that have at least half the power of the resonant frequency are considered to be in the "pass band" of this curve. Hence, all frequencies between the points at which the output voltage is more than 70% of its maximum are in the pass band, because the power is proportional to the square of the voltage. (.7 × .7 = .49 or about ½ the power.)

This means that all frequencies between the two points giving voltages that are 70% of 60, or 42, are in the "pass band" in Fig. 5. As Fig. 6 shows more clearly, the pass band therefore extends from about 993 kc. to 1009 kc.

As you know, in the broadcast band, the ideal curve should be 10 kc. wide, as shown by the shaded area in Fig. 6. Therefore, our resonance curve shows

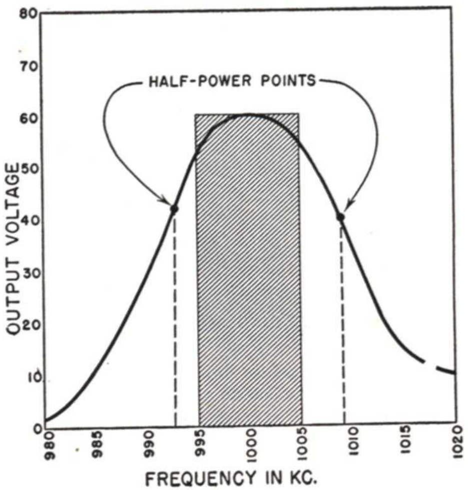


FIG. 6. How our typical curve compares with the ideal response.

that the fidelity is fairly good because practically all the ideal curve is within the pass band—there is but a slight reduction in the "corners" of the ideal. However, the selectivity isn't good, because large parts of the side bands of the adjacent-channel stations lie within our pass band too. As a matter of fact, even the carrier frequency of a 990-kc. station would get a gain of about 30, or half the voltage gain of the desired carrier. As we shall show later, this is far too much—this carrier should be reduced to at least 1/1000 that of the desired carrier if we are to be assured of good separation.

You can see, then, that the selectivity of this stage leaves much to be desired. What can we do to get better selectivity? First, let's see what determines the exact shape of the resonance curve, so we can see how to "sharpen" it.

EFFECT OF Q

As you might suspect, something within the resonant circuit determines where the half-power points will be. It is the Q of the circuit that sets the pass-band width, and, as you learned earlier, the Q is equal to the inductive reactance divided by the a.c. resistance in the resonant circuit, or

$$Q = \frac{\text{coil reactance}}{\text{a.c. resistance}}$$

The a.c. resistance is composed of the d.c. resistance plus certain losses that occur when alternating currents flow through a coil. One of these factors is that a.c. tends to flow on the surface of a wire rather than throughout the wire (a phenomenon known as "skin effect"). This means that the wire has effectively less cross-sectional area, because it acts as if it were hollow. The less the area, the higher the resistance, so this "skin effect" means that the resistance of the wire is higher for a.c. than for d.c. This effect increases as the frequency increases, as do certain other losses that we won't go into here. The result is that the a.c. resistance is appreciable and changes with frequency.

The pass-band width can be found by dividing the resonant frequency by the resonant circuit Q, or

$$\text{Pass Band} = \frac{\text{Resonant Frequency}}{\text{Q of the tuned circuit}}$$

Notice that, if the Q is fixed, the band width increases as the resonant frequency increases. This important fact helps us get better fidelity in f.m. and television (TV), where rather broad pass bands are necessary and where higher resonant frequencies are used than is customary in a.m. broadcasting.

From the preceding equation, we can find the Q we need for a particular pass band. In our example in Fig. 6,

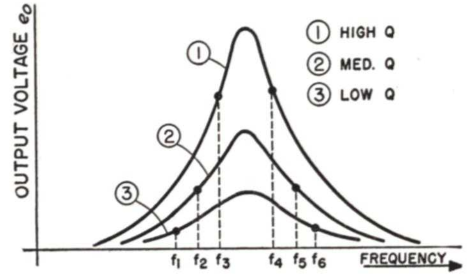


FIG. 7. As the Q is increased, the pass band becomes narrower.

we have a resonant frequency of 1000 kc., and the pass band is 16 kc. (1009 — 993 = 16). By rearranging our formula to be

$$Q = \frac{\text{Resonant Frequency}}{\text{Pass Band}}$$

we find that 100 ÷ 16 = 62.5, which is the Q of the circuit.

If we are to get a pass band of 10 kc., we need a Q of 100 (1000 ÷ 10). Thus, the higher the Q, the narrower the pass band.

Fig. 7 shows how the resonance curve changes with Q. The "high-Q" curve 1 has a pass band between f_3 and f_4; the "medium-Q" curve 2 has a wider pass band between f_2 and f_5; the "low-Q" curve 3 has a pass band between f_1 and f_6.

You can see that increasing the Q constricts the pass band and therefore decreases the amplification offered to frequencies in adjacent channels. If the Q is increased enough to make the amplification of adjacent-channel signals negligible, however, the pass band will be so narrow that the fidelity will be seriously decreased. Furthermore, there is a limit to the Q we can obtain. A Q of 100 to 150 is about the best we can expect, so, for a fixed resonant frequency, there is a limit to how much selectivity we can get in any case.

Gain. Fig. 7 shows that selectivity and fidelity are not the only things affected by the Q. Notice that the voltage output increases (the peak gets higher) as we move from low Q to high Q. This happens because the output voltage in our circuit (Fig. 4) depends on the load impedance, and the impedance of a parallel-resonant circuit equals the coil reactance multiplied by the Q. Hence, the load impedance and output voltage vary directly with the Q.

Conclusions. When a resonant circuit is used as the load in an r.f. stage, the gain and the selectivity both increase if the Q of the resonant circuit is increased. However, if the Q is too high, the fidelity suffers because the pass band becomes too narrow. Conversely, the gain and the selectivity decrease if the Q is decreased, but the fidelity improves. There is a practical limit to the amount of Q we can obtain—values of 100 to 150 can be had, but only by careful design—so there is a limit to the gain and selectivity that a single stage can give.

If more gain and selectivity are needed, we can get them by connecting several stages in cascade—that is, with one feeding the next. Let's see what effect this has.

CASCADE R.F. STAGES

In our study of audio-frequency amplifiers, we found that the over-all gain from several stages in cascade is equal to the product of the gains in each stage. Hence, if we have three audio stages, each with a gain of 10, the over-all gain of the three in cascade will be $10 \times 10 \times 10$ or 1000.

The same sort of calculation can be applied to cascaded r.f. amplifiers. It must be remembered, however, that every r.f. amplifier has its own frequency-response curve. The over-all response curve we obtain from several of them in cascade, therefore, should be approximately equal to what we would get if we "multiplied" the individual response curves together.

As an illustration of what occurs, let us examine Fig. 8. Suppose we have

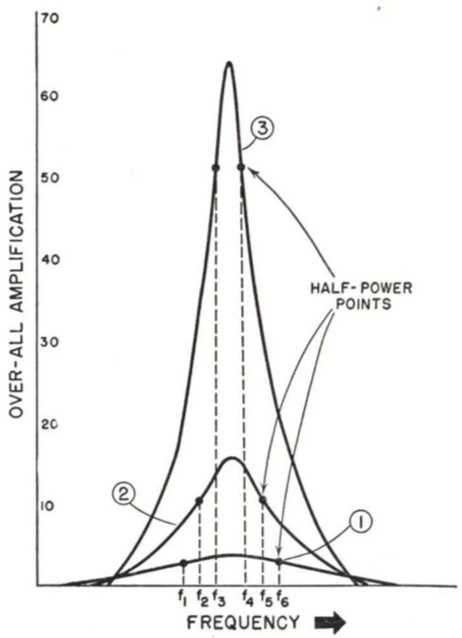


FIG. 8. Representative response curves for a single r.f. amplifier, two amplifiers in cascade, and three identical cascaded stages. Besides the increased gain, note the improvement in selectivity resulting from cascade operation.

three r.f. amplifiers just alike, the response of each one being like curve 1 in the figure. If we now arrange two of these amplifiers in cascade, the over-all response curve will resemble curve 2. Because the two stages are just alike, and we have multiplied their responses together, curve 2 is actually the square of curve 1.

Similarly, if three of the amplifiers are used in cascade, we get the response shown by curve 3. In this particular instance, since the gains of each stage were made the same, curve 3 is the cube of curve 1 (that is, the curve 3 values are the curve 1 values multiplied by themselves three times; if the peak value of curve 1 is 4, then curve 3 will be $4 \times 4 \times 4 = 64$).

Notice the shapes of the curves. By marking the half-power points on each of these response curves, we find that the pass-band of a single amplifier (curve 1) extends between the frequencies f_1 and f_6; for two amplifiers in cascade, this is narrowed to the frequencies f_2 and f_5 on curve 2; and on curve 3, representing three amplifiers in cascade, the pass band is diminished still more and lies between frequencies f_3 and f_4.

This demonstrates the fact that operating r.f. amplifiers in cascade not only increases the gain tremendously, but also improves the selectivity to a great degree. Unfortunately, the use of stages in cascade also makes it possible for the pass band to get too narrow for good fidelity. It's easily possible to get a gain of 1,000,000 out of three r.f. stages having gains of 100—a not unreasonable figure. Such high peaks are always very sharp, so the pass band may become only 5 or 6 kc. wide. Even when stages are connected in cascade, then, it is still necessary to choose either fidelity at the sacrifice of selectivity or selectivity at the sacrifice of fidelity.

Let's go on now to learn a few more facts about practical r.f. stages, and then learn about double-tuned circuits that give *both* selectivity and fidelity at a sacrifice in gain.

Practical R. F. Amplifier Stages

Before we get into a study of typical r.f. stages, let's take a moment to see how these stages are arranged.

T.R.F. Receiver. In the tuned-radio-frequency (t.r.f.) receiver, from two to perhaps six similar stages are arranged in cascade, as shown in Fig. 9A. Each stage has its own resonant circuit, and they are all tuned simultaneously (by varying either the inductance or the capacity) to the modulated carrier frequency desired. When resonant circuits are tuned to various frequencies over a wide band in this manner, the selectivity and fidelity are different at different points on the tuning range. The Q also varies with frequency, another effect that will cause wide variations in the results. These variations have led to the abandonment of the t.r.f. receiver in favor of the superheterodyne, except in a few very inexpensive midget sets.

Superheterodynes. In the superheterodyne, the desired signal plus nearby signals is picked out by the preselector, which may be only a tuned circuit, or may be a complete r.f. amplifier stage or two. Then, as shown in Fig. 9B, the incoming signal is fed into a frequency converter where it is mixed with the signal from a local oscillator (which we shall study in another Lesson). An intermediate-frequency signal on which the original modulation is impressed is produced from the mixed signals in this stage. The local oscillator tuning is adjusted by the tuning controls to produce this same intermediate frequency from any signal in the tuning range of the set; effectively all received signals are converted to one fixed value. This intermediate-frequency (i.f.) signal is then passed through an amplifier consisting of from one to four stages that are all tuned to the one fixed i.f. frequency. Since these i.f. stages need to handle only the fixed i.f. signal plus the modulation side bands, there is no change of Q or of bandwidth with tuning, and since the i.f. amplifier is designed to give most of the selectivity and gain, approximately the same gain, selectivity, and fidelity can be obtained over the complete tuning range. The facts that follow apply to both the variable-tuned r.f. stage as used in t.r.f. sets and superheterodyne preselectors, as well as to the fixed-tuned type used as i.f. amplifiers, except where otherwise indicated.

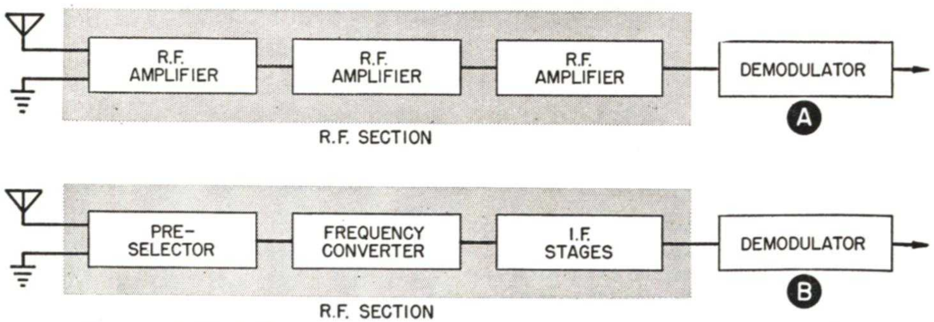


FIG. 9. A block-diagram comparison of the t.r.f. and superheterodyne receivers.

PARALLEL-RESONANT PLATE LOADS

The parallel-resonant circuit used as a plate load, which we have been studying, is no longer used much in sound receivers. However, the variation shown in Fig. 10 is the basis of a type used in several television receivers. Here, R_1 is deliberately chosen to have a low resistance—10,000 ohms or less. It is in parallel with the resonant circuit L_1-C_1, and, since the impedance of parts in parallel is always less than that of the smallest, the plate load for VT_1 cannot be higher than the value of R_1 even at resonance. Also, R_1 acts like a load on the L_1-C_1 circuit, reducing the Q. The effect is shown in Fig. 11. This broadening of the response is necessary in television because we want to pass an extremely wide band of frequencies.

At the same time, such broadening drastically reduces the gain, which probably isn't very high to begin with. Special r.f. pentode tubes are used in television to get around this difficulty.

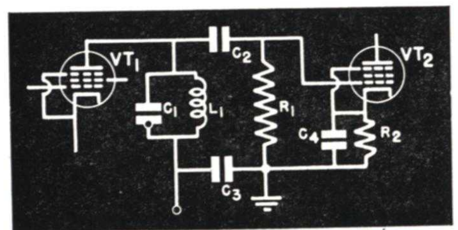


FIG. 10. A basic form of a television i.f. stage.

When the load is far less than the a.c. plate resistance of the tube, as when a low-Q resonant circuit is used as the load for a pentode, the gain is approximately the product of a.c. plate current and the load impedance. In any r.f. pentode, the a.c. plate current is dependent almost exclusively on the grid signal and on the mutual conductance of the tube. The tubes developed for television have very high mutual conductances — much higher than those of the r.f. pentodes commonly used in a.m. and f.m. sound receivers—so they pass high a.c. plate

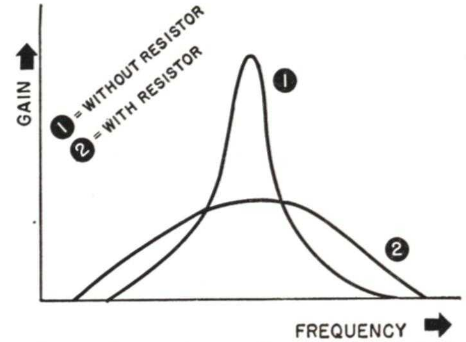


FIG. 11. How resistance loading broadens the response curve.

currents through their loads, thus increasing the stage gain.

TRANSFORMER COUPLING

The load arrangement most commonly found in t.r.f. stages and in superheterodyne preselectors of sound receivers is shown in Fig. 12. Here, the resonant circuit L_4-C_5 is in the grid circuit of another stage, and is coupled to the plate of VT_1 by L_3; here L_3 and L_4 form the primary and secondary of an r.f. transformer. We cannot use the turns ratio of an r.f. transformer to express the effects of the secondary on the primary as we did with iron-core transformers, because we do not have the near-perfect coupling found in iron-core units. The secondary of an r.f. transformer may have many turns, but only a few may be effectively linked with the primary.

At radio frequencies, the *mutual in-*

ductance between the primary and secondary windings is a direct measure of the amount of coupling actually present, and it depends not only on the number of primary and secondary turns but also on the spacing and physical shapes of the two coils. Since the secondary turns are fixed by the requirement that L_4 must be able to be tuned to resonance by C_5, the only way the designer can adjust the mutual inductance is to adjust the size and position of the primary winding.

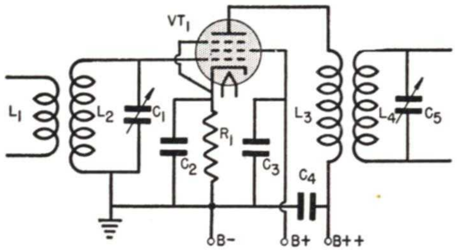


FIG. 12. A typical t.r.f. or preselector stage.

The tuned secondary circuit acts as a load on the primary through the mutual inductance. Whatever the impedance of the secondary circuit may be, it reacts on the primary circuit as a "reflected" impedance that appears to be in series with the primary winding, and this acts as the load for VT_1.* This reflected impedance increases if the mutual inductance is increased and decreases if the impedance of the secondary circuit is increased. At resonance, the impedance of the secondary circuit is effectively just that of the a.c. resistance, because the coil reactance equals the condenser reactance and cancels it. Therefore, the Q of the secondary circuit determines what the load will be. If the Q is good, the secondary a.c. resistance must be low, and therefore the reflected effects on the primary at resonance will be high. If the Q is low, the a.c. resistance will be high and the primary reflected impedance will be low. Therefore, the use of a high-Q secondary means that effectively tube VT_1 has a high load impedance and therefore a high gain.

At frequencies other than the resonant one, there will be an inductive or capacitive reactance in the secondary, so the secondary impedance will rise rapidly. This means that, off resonance, the load impedance reflected into the primary winding will decrease rapidly, so VT_1 will not amplify as much. We are therefore back to practically the same circuit action that we had with our parallel-resonant plate load. Again we have an r.f. stage that has variable gain because its load impedance changes with frequency. In this case, the amount of gain depends on the mutual inductance, and on the Q of the resonant circuit.

Reflected Load. If we were using triode tubes or other tubes with low plate impedance, we would find that the maximum gain would be obtained at one particular value of mutual inductance. In this case, the mutual inductance may be said to be practically matching impedances, as it works out that at the optimum value the reflected load in the primary circuit exactly equals the plate resistance of the tube. As you will recall, this is a condition for maximum power transfer. If the mutual inductance is too low, the gain is decreased because there is not enough transfer of available energy to the secondary; if the mutual inductance is too high, on the other hand, the gain is reduced because the reflected impedance in the primary circuit begins to reduce the plate current, which of course cuts down on the signal transfer.

*The formula for the reflected impedance is $Z_p = \frac{(\omega M)^2}{Z_s}$ that is, the impedance in the plate circuit (Z_p) is equal to the square of the mutual impedance (ωM) divided by the impedance of the tuned secondary (Z_s).

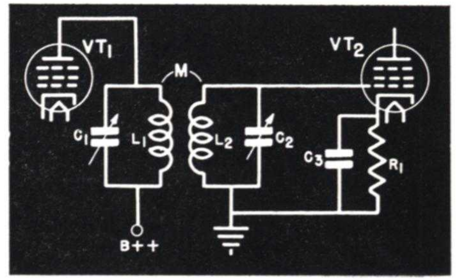


FIG. 13. This dual tuned circuit is typical of the i.f. stages in sound receivers.

Since we are using r.f. pentode tubes with exceedingly high plate resistances, however, it is practically impossible to get enough mutual inductance to cause the reflected impedance to equal the plate resistance value. For this reason, the stage gain will increase with increases in the mutual inductance. There is a practical limit to the amount of mutual inductance we can get, however; if the primary winding is made too large in an attempt to increase the coupling, its inductance will resonate with its distributed capacity and produce an undesirable response at the self-resonant frequency. Also, as the mutual inductance is increased, resistances are reflected from the primary into the secondary in such a way that the Q of the secondary is reduced. In general, therefore, we find that somewhat higher selectivity but less gain and less fidelity are obtained if the transformer is designed for less coupling to the secondary, and that more gain, more fidelity, and less selectivity are obtained by coupling with a higher mutual inductance.

DUAL TUNED CIRCUITS

As long as selectivity and gain are most important, single tuned circuits arranged in either of the two manners we have just described may be used. However, it is possible to get much more fidelity (particularly if a wide pass-band is necessary) and still get very good selectivity and reasonable gain by using dual tuned circuits arranged as shown in Fig. 13. Here we have an arrangement of two tuned circuits, both alike and tuned to the same frequency. This is the arrangement used in almost all i.f. amplifiers of superheterodyne a.m. or f.m. sound receivers.

If we run response curves for this arrangement, we will find that the kind of curves we get will depend on the mutual inductance.

If we run a response curve on this arrangement when there is very little coupling, we will find there will be little energy transferred from the primary to the secondary winding, and the primary resonant circuit as a resistance equal to the resistance already in the primary circuit. Effectively, therefore, by the time this coupling has been reached, the Q of the primary circuit has been reduced to one-half the value it would have if the secondary circuit were not coupled to it. Therefore, the gain of the VT_1 stage is half what it would be if the stage had just the primary winding as a

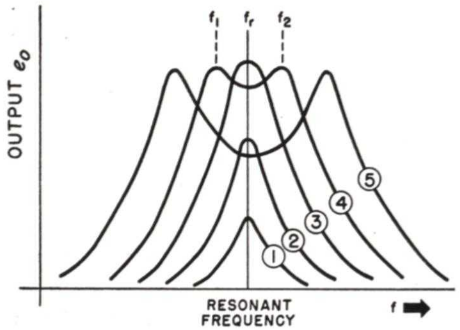


FIG. 14. How the response of a dual tuned circuit varies with the mutual inductance.

consequently the output voltage will be rather low. The response curve will be something like curve 1 in Fig. 14.

If there is a larger mutual inductance (that is, if the coils are closer together), we will get something like curve 2. As you might expect, the closer coupling will permit more energy to be transferred, and the output voltage will therefore become greater. If the coupling is made still closer, the output voltage will continue to increase until a maximum value is reached (curve 3). At this point we will have what is called critical coupling—no greater output voltage can be obtained either by increasing or decreasing the coupling.

Critical coupling is reached when the secondary impedance reflects into parallel-resonant load. However, the selectivity is considerably greater than that of the resonant load by itself. This increase in selectivity occurs because the primary and secondary resonant circuits are effectively connected in cascade, and their output is the product of their response curves. In other words, the voltage induced in the secondary depends on the current in the primary resonant circuit. This current reaches a peak value at resonance; therefore a peak voltage is induced in the secondary. Effectively, then, the primary circuit feeds the secondary, producing a cascade connection. As you learned earlier in this Lesson, cascading produces a resonance curve with steeper sides.

As we go beyond critical coupling,

a very interesting thing occurs. As shown by curve 4, the response curve now has two peaks, one on either side of the resonant frequency, at frequencies f_1 and f_2. (The response is lower at the resonant frequency than at f_1 and f_2 and is also lower than it was in curve 3.) Beyond f_1 and f_2 the response falls off sharply. If the coupling is increased even more, these peaks separate farther and farther as shown in curve 5.

This double-humped response is the result of the impedance that is coupled from the secondary into the primary. The closer coupling means that increased resistance is reflected into the primary at resonance; therefore, the primary Q falls at the resonant frequency. This decreases the primary current at resonance, producing the "dip" in the center. At frequencies on either side of resonance, however, the impedance reflected from the secondary into the primary has a reactive component in addition to resistance. This reflected reactance is the opposite of what the actual secondary reactance is—if the secondary is capacitive, the primary "sees" an inductance; if the secondary is inductive, the primary "sees" a capacity. At frequencies above or below the resonant one, therefore, the secondary reflects into the primary a reactance that cancels some of the reactance of the primary. In consequence, the primary current becomes higher.

In other words, when the coupling exceeds the critical amount, the primary circuits acts as if it were no longer tuned to resonance with the secondary — instead it apparently automatically tunes itself to two frequencies, one on either side of the actual resonant one. Therefore, the primary current has peaks on either side of resonance, which of course means that the voltage induced in the secondary will have similar peaks.

The two humps in curve 4 of Fig. 14 are not very far apart, and the hollow between them is quite small. Therefore, this curve more nearly approaches the ideal response curve than does any other we have previously shown, because it has practically a flat top. In other words, all the frequencies in the band between f_1 and f_2 are amplified by approximately the same amount. Because of this band-amplifying response, this is sometimes known as a band-pass curve.

Obviously an r.f. amplifier having this double-peaked response is superior in fidelity to one having a single-peaked response, because a wider band of frequencies is amplified equally; yet no selectivity is sacrificed, because the sides of the response curve are as nearly vertical as the circuit Q's will permit. Gain is a little less for the double-peaked response, however.

As soon as we get away from a single-peaked response, the half-power points can no longer be used to determine the pass-band width. Engineers measure the selectivity of a band-pass circuit in terms of how much better the amplification is at the desired carrier than at the nearest undesired carrier. Good selectivity requires that the desired carrier be amplified at least 1000 times more than the nearest adjacent carrier.

The fidelity of such a band-pass response is measured by the variation in gain over the band of frequencies to be passed. The response is considered to be of high fidelity if the ratio between the highest gain and the gain at the resonant frequency is not more than 1.25 to 1. This represents a very flat output characteristic, particularly in comparison with the part of a single-peak curve that is included between the half-power points.

CIRCUIT VARIATIONS

The three basic r.f. stages that we have studied up to now include by far the majority of the r.f. amplifier circuits found in a.m. and f.m. sound receivers. However, there are a few noteworthy exceptions.

The first basic exception, which is used in a few small table models as an i.f. stage, is shown in Fig. 15. This is essentially the parallel-resonant plate-load circuit, coupled to another coil that is untuned. The gain of the VT_1 stage depends on the Q of the C_1-L_1 resonant circuit, and since the secondary is untuned and is loosely coupled, there are no reflections into the primary to reduce this Q apprecia-

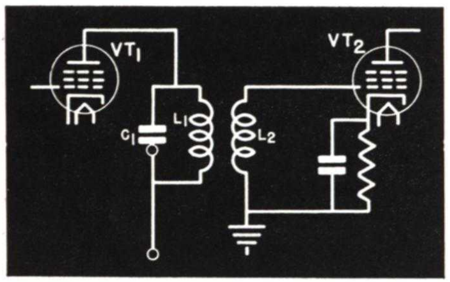


FIG. 15. This is basically a parallel-resonant plate-load circuit.

bly. About twice as much gain can be obtained with this arrangement as with the double-tuned i.f. transformer. However, this gain is obtained at a sacrifice of the fidelity and selectivity, so the circuit is not used very much.

Untuned R.F. Stages. In the early days of radio, certain patent restrictions forced the use of untuned r.f. stages. In such cases, the tuned circuits were arranged in a group ahead of the first tube. Each tuned circuit fed into the next through an inductive or capacitive coupling. (Such band-

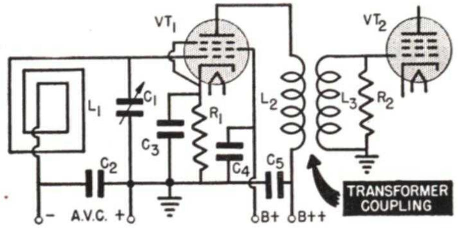


FIG. 16. A broadly tuned transformer coupling.

pass couplings will be covered elsewhere in your Course.) Then, the r.f. amplifier tubes were coupled to each other by what appeared to be non-resonant r.f. transformers. Such stages had low gain, of course.

Today there are a few examples of this basic type remaining, usually only in table-model receivers. An example is in Fig. 16. Here, the tuned circuit is L_1-C_1, with L_1 being the loop antenna of the set. R.F. amplifier VT_1 then is coupled to the mixer VT_2 by an r.f. transformer L_2-L_3. This transformer appears to be untuned; actually, however, its inductance is such that it resonates with the distributed circuit capacities over the entire broadcast band. This broad peak is achieved by loading the secondary with resistor R_2. Even though the peak is very broad, VT_1 contributes a little gain, which is helpful in overcoming some of the noise that superheterodynes introduce in the frequency conversion process.

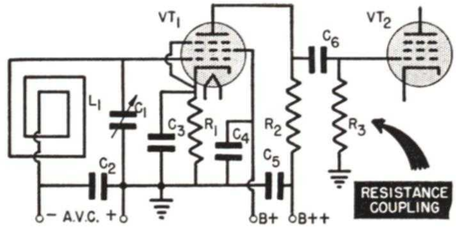


FIG. 17. An example of r.f. resistance coupling.

A few sets have used resistance coupling: an example is shown in Fig. 17. Such coupling gives very little gain at radio frequencies, however.

VIDEO I.F. STAGES

We have already said that a loaded parallel-resonant circuit is used in television as a broad-band amplifier. Because of the extremely broad band that has to be passed, we would have to load the video i.f. stages so much (if we tuned them all to the same frequency) that we would have practically no gain. Hence, it is more practical to use stagger tuning with less loading in each stage. In this arrangement, one stage is tuned to one resonant frequency, the next stage is tuned to another resonant frequency slightly removed, and so on. The result is that the over-all pass band is the product of all the response curves.

An example of a typical response curve for two stagger-tuned stages is shown in Fig. 18. One circuit is tuned

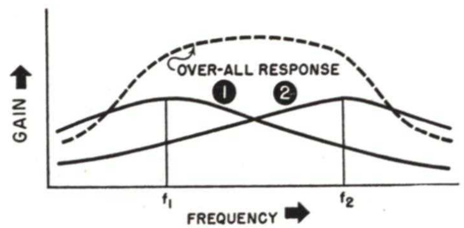


FIG. 18. Stagger tuning broadens the response and improves the selectivity.

to f_1 and the other to f_2. The over-all response, which is the product of the two response curves, is considerably broader and has steeper sides.

Even so, this kind of response does not always offer enough adjacent-channel selectivity, so special trap circuits are used in television receivers. We shall leave the detailed discussion of these circuits for later Lessons. In general, however, they operate by being tuned to the *undesired* signals. These resonant circuits are then used

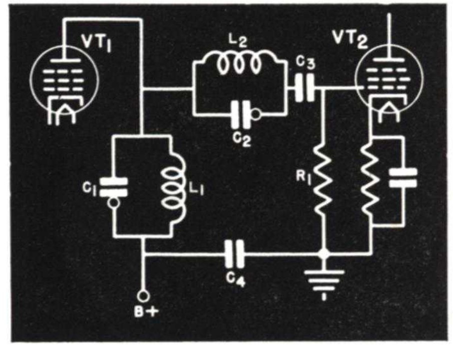


FIG. 19. An example of a television trap tuning circuit.

to reduce the response to the undesired frequencies.

A basic circuit of this kind is shown in Fig. 19. Here the resonant circuit L_1-C_1 is tuned to the *desired* signal and acts as a plate load as before. Then, the trap circuit L_2-C_2 is tuned to an undesired signal. This is a parallel-resonant circuit, so it offers maximum impedance at resonance. It acts therefore as a voltage divider with R_1; at its resonant frequency, most of the undesired signal will be dropped across it, leaving little for R_1. Since it is tuned to an undesired frequency, this is just what we want. At the desired signal frequency—the one to which

L_1 and C_1 are tuned—it no longer offers such a high impedance, so most of this signal is transferred from L_1-C_1 to R_1 and on to the next stage.

Now that you have a general idea of the types of circuits used in r.f. stages, let's learn something about the tubes that are used. You are generally familiar with the basic types, but there are certain important modifications in those intended to be used in the r.f. stages of radio receivers.

More Facts About Triode Tubes

We mentioned earlier that screen-grid or pentode tubes are used in r.f. and i.f. amplifier stages. Let's learn a little more about triode tubes to see why they aren't good r.f. amplifiers in the same circuits used for pentode tubes.

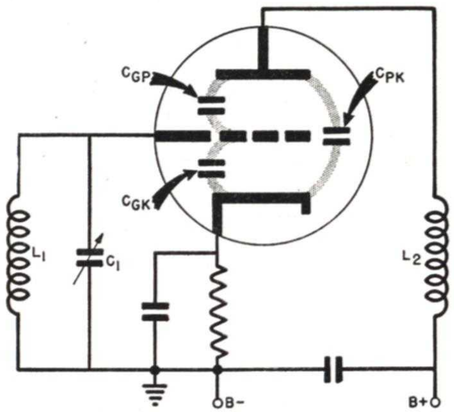


FIG. 20. The internal tube capacities in a triode tube.

One of the most important drawbacks of the triode tube is the appreciable inter-electrode capacity that exists between the grid and plate within the tube. You know that there is always a capacity between any two conductors separated by a dielectric; therefore, all tubes have such capacities between all their elements. Fig. 20 shows all those in a triode tube. Let's see what two important effects these capacities have.

INPUT CAPACITY

Fig. 21 shows the equivalents of these tube capacities and how they appear across the input circuit. As you see, the grid-cathode capacity C_{GK} is directly across the input, and the other capacities are in series across the input circuit. These capacities are therefore in parallel with the tuning condenser C_1 in Fig. 20, so they serve to detune the resonant circuit L_1-C_1. Since these capacities are small, we could compensate for their effect by adjusting the tuning slightly. However, the grid-plate capacity C_{GP} provides a feedback from the plate circuit that apparently increases the input capacity to a far higher value. Let's see how this happens.

If we lump the effects of the actual interelectrode capacities and add a load R_L to the circuit, we get the equivalent circuit shown in Fig. 22, in which C_{IN} represents the lumped capacities. This figure shows that the grid-plate ca-

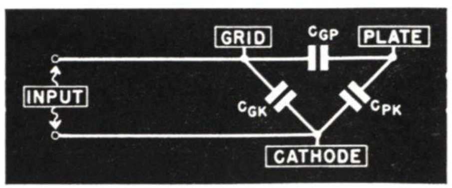


FIG. 21. How the triode tube capacities shunt the input circuit.

pacity C_{GP} effectively connects the load R_L across the input terminals of the tube. This means that the signal voltage produced across the load will also be impressed back across the input terminals through this capacity. Since the voltage across R_L is large compared to the input signal voltage, an appreciable current passes through C_{GP}. Depending on the phase relationship of the input and output voltages, and on the stage gain, it is possible for this

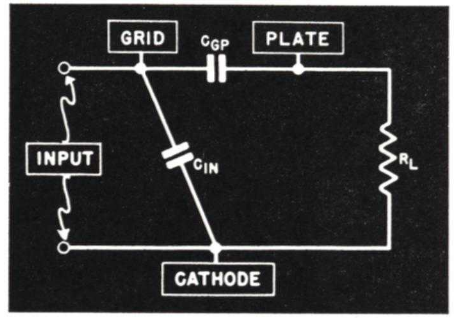


FIG. 22. The grid-plate capacity provides a feed-back from the load to the input.

feedback to have the same effect on the grid circuit as if the input capacity had been increased greatly.

If the load is resistive—that is, if it is a resistor or a parallel-resonant circuit tuned to resonance—the feedback voltage is of such phase that the effective capacity across the input is a maximum. If the load becomes reactive, on the other hand, the phase relationship changes so that the capacity drops. Even so, it will generally be much higher than just the interelectrode capacity.

Incidentally, their high input capacity limits the range of frequencies that can be handled by ordinary tubes. At the ultra-high frequencies, the input capacity of an ordinary tube becomes as large as or larger than the capacity needed for tuning; once the inductance has been reduced as much as possible, therefore, the circuit cannot go to higher frequencies. This is one of the reasons why ultra-high-frequency tubes are especially designed to have low capacities.

REGENERATION

There is another and even more important effect of the feedback path furnished by the grid-plate capacity. Besides changing the apparent input capacity, the feedback voltage may aid or oppose the incoming signal in such a way that the stage gain and the selectivity are seriously affected.

It is possible to show the effects of this feedback to the grid circuit by assuming that it adds a resistance as well as a capacity to the grid circuit. As shown in Fig. 23, these additional parts are R_A and C_A, (C_A combines with C_{IN} and has the effect we have described of increasing the total input capacity.)

If the load in the plate circuit is resistive, the feedback voltage appears across the input circuit in such a manner that the resistance R_A in Fig. 23

is very high—so high, in fact, that it is virtually an open circuit and can be neglected.

In practical circuits, however, the plate load is usually somewhat reactive. For example, when a parallel resonant circuit is used as the plate load, it is resistive only to the resonant frequency. To frequencies slightly above resonance, it is a capacitive load; to frequencies below resonance, it is inductive. Even when we use resistance coupling, the capacities in the the feedback voltage is in phase with the input voltage and adds to it—producing the same effect as if the Q of resonant circuit L_2-C_1 were made larger. What actually happens is that the feedback of energy from the plate circuit makes up for some of the losses in the tuned circuit. When this occurs, the Q of the tuned circuit is apparently higher, because there is a larger circulating current in the tuned circuit for the same amount of applied signal voltage. Engineers say that R_A now

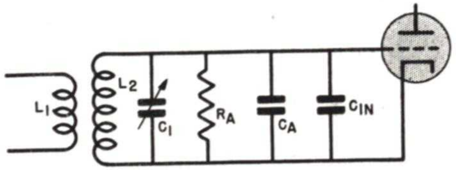


FIG. 23. The input effects caused by tube capacities.

circuit are likely to make the load have a capacitive effect in addition to being resistive. Under this condition of a capacitive load, the feedback voltage will be out of phase with the input and will tend to reduce the amount of input voltage fed to the tube. In other words, the circuit acts as if R_A were a low resistance that loads the tuned circuit and reduces its Q.

In radio-frequency stages it is far more common for the load to be a transformer with a tuned secondary. In this case, the primary inductance provides an inductive load in the plate circuit. This inductance may be partially but never completely cancelled by the reflected effects from the tuned secondary, or it may even be increased by such reflections when the secondary is not tuned exactly to resonance. In any case, when the load is inductive, has a *negative* value, since it cancels some of the resistance in the tuned circuit.

When the feedback voltage opposes the input and effectively reduces it, we call the feedback a *degeneration* effect. When, on the other hand, it aids the input signal, we call it *regeneration.*

A certain amount of regeneration is not undesirable, for it reduces the losses in the tuned circuit and effectively makes the amplifier have more gain. However, with the average triode, the feedback voltage will frequently become large enough for the circuit to become an oscillator—that is, the circuit will produce its own signal. This will destroy the effectiveness of the circuit as an *amplifier* because the feedback voltage will "take over" and prevent true reproduction of the input

signal in the plate circuit. All you can get out of an oscillating r.f. amplifier is a squeal produced by the new r.f. signal beating with the incoming signal.

GETTING RID OF FEEDBACK

Suppressor Methods. One method of controlling regeneration is to introduce losses in the circuit. This may be accomplished in two ways. In one, resistance is added in series with the resonant circuit. This greatly lowers the Q of the resonant circuit, so the feedback merely raises the Q of the circuit back toward normal. This method is not too desirable, since the set will have subnormal amplification if the feedback becomes less for any reason.

Another losser method is to use a grid suppressor as shown in Fig. 24.

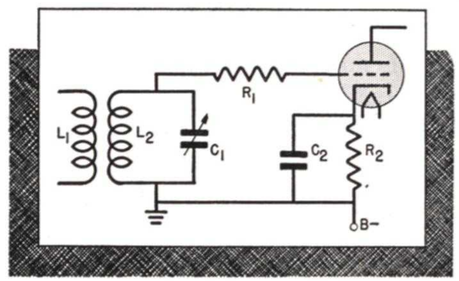


FIG. 24. A grid suppressor "losser" method of controlling regeneration.

Here, resistor R_1 is not in the tuned circuit, so it has little effect on its Q. Furthermore, it has but little effect on the input signal, because practically no signal current normally flows to the grid from the resonant circuit. (The grid is considered to be a voltage-operated device whose effect depends only on the voltage between it and the cathode.) However, the feedback current from the plate circuit must flow through R_1 and the resonant circuit to complete its path back to the load; in flowing through R_1, much of its energy is dissipated in the form of heat. (Resistor R_1 is not by-passed, because it is desired to lose this energy.)

The value of R_1 is usually between 100 ohms and about 5000 ohms. However, its value has to be found by trial in each stage. This is impractical for mass production of radio receivers, so this idea is no longer used in radio receivers intended for home use.

Neutrodyne Circuits. The development of the neutrodyne principle made triode r.f. stages much more practical. The neutrodyne receiver used an ingenious method of removing the effects of regeneration while retaining all the amplification of the stage. In this system, an a.c. voltage from the plate circuit of the amplifier is fed back to the grid circuit in such a way that it is out of phase with the voltage fed back through grid-to-plate capacity. Then, the two voltages are made to balance out or neutralize each other. In other words, no attempt is made to reduce the grid-plate feedback; instead, another voltage is deliberately introduced to neutralize the feedback voltage.

Fig. 25 is an example of a neutrodyne circuit. Here, the voltage across coil L_1 (the plate load) feeds back through the grid-plate capacity C_{GP} (within the tube) to the grid circuit. To offset this, coil L_2 is coupled with coil L_1 in such a way that the voltage across L_2 is out of phase with the L_1 voltage. (When the plate end of L_1 is positive, the lower end of L_2 is negative with respect to their junction.) Hence, the voltage from L_2 that is fed through the condenser C_N to the grid is out of phase with the undesired feedback voltage. If condenser C_N is adjusted so that the two voltages are equal, they will cancel each other exactly, and the regeneration effects will be eliminated.

Another typical circuit is shown in Fig. 26. As you can see, this is merely the reverse of the one shown in Fig. 25. The neutralizing feedback energy goes through coil L_2, which induces a voltage in grid coil L_1 that is out of phase with the feedback through C_{GP}. The operation is otherwise exactly like that of Fig. 25.

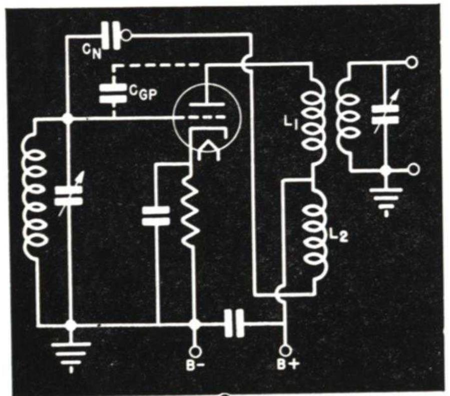


FIG. 25. A basic neutrodyne circuit.

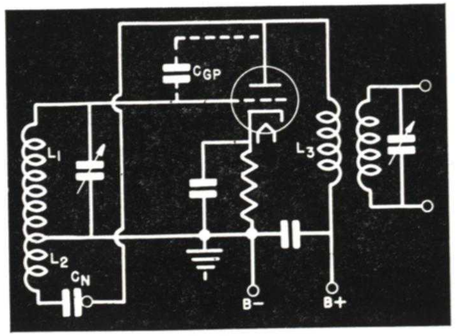


FIG. 26. Another neutrodyne circuit.

Many other variations of this same principle may be used, but essentially all of them include a neutralizing condenser connected so that out-of-phase energy can be fed from the plate circuit of the tube to the grid circuit.

Every time the tube is changed, it is necessary to readjust the neutralizing condenser to compensate for the new grid-plate capacity, because no two tubes are exactly alike.

These methods of overcoming the effects of grid-to-plate capacity make it possible for triode tubes to be used as r.f. amplifiers. However, the modern practice in sound receivers is to use tubes in which the grid-to-plate capacity is small, thus eliminating the problem. We shall study these tubes next.

Screen-Grid and Pentode Tubes

The screen-grid or tetrode tube was the first to have a low grid-plate capacity. A typical circuit using this tube is shown in Fig. 27. As you can see, there is another grid between the control grid and the plate. Let's learn what effect this has.

REDUCING GRID-PLATE CAPACITY WITH THE SCREEN GRID

In Fig. 27, notice the screen-grid by-pass condenser C_2 and the cathode by-pass condenser C_1. Both these condensers are large enough to have very little opposition (negligible impedance) to radio frequencies. Hence, for r.f. voltages, the screen grid is effectively connected directly to the cathode. This gives us the equivalent circuit shown in Fig. 28. As you can see, there is a capacity C_{GSG} between the grid and the screen grid, and another between the screen grid and the

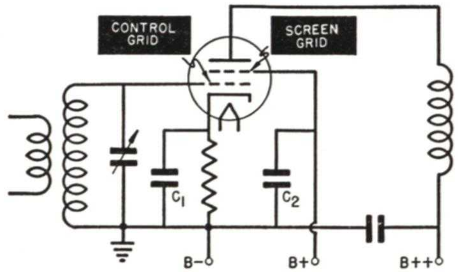


FIG. 27. The early screen-grid tube.

plate (C_{PSG}). However, since the screen grid and the cathode are effectively tied together, the feedback voltage from the load R_L travels through C_{PSG} to the screen grid but is then returned to the cathode (through C_2 and C_1 in Fig. 27) and thus to the load. It does not get to the control grid.

There is still a very small amount of capacity between the grid and plate, represented by the shaded lines and condenser C_{GP} in Fig. 28. However, this capacity is so much smaller than

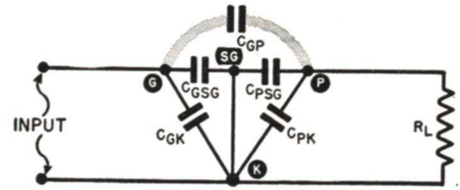


FIG. 28. The internal capacities of a screen-grid tube. Here, (G) is the grid, (SG) the screen grid, (P) the plate, and (K) the cathode.

it is in the triode tube that the amount of feedback through this path is negligible by comparison: the grid-plate capacity of triode tubes ranges from 3 to 10 mmfd., whereas it is less than .01 mmfd. in the screen-grid tube.

Although the grid-plate capacity is reduced to negligible proportions, the input and output capacities of the screen-grid tube are somewhat higher than those of the triode because the grid-to-screen-grid capacity (C_{GSG}) and the screen-grid-to-plate capacity (C_{PSG}) are added to the input and output capacities respectively. However, these are fixed capacities that are not changed by the load voltage as are those of the triode.

Of course, the parts used in r.f. amplifier circuits must be laid out to keep the grid and plate circuits well separated. Otherwise, there can be coupling between the leads outside the tube that will cause oscillation just as readily as will grid-plate capacity within the tube. In the early screen-grid tubes, the control-grid lead was brought out through the top of the tube and the plate lead was brought out through the socket to keep these leads as far apart as possible. Modern wiring methods and the decreased size of parts make it possible today to use tubes in which both the plate and the grid leads come out through the bottom. As a serviceman, you will find it necessary to observe precautions in replacing parts so as to keep leads separated and as short as possible. Disturb r.f. wiring as little as you can.

FURTHER FACTS ABOUT SCREEN-GRID TUBES

Since a positive voltage (with respect to the cathode) of somewhere between one-third and one-half the plate voltage is always applied to the screen grid, and since this grid is considerably nearer the cathode than the plate is, the plate of this tube exerts far less control on the plate current than the screen grid does. As a matter of fact, if the screen-grid voltage is kept constant, the plate voltage can be changed considerably without affecting the plate current much. This fact is shown by the family of E_p-I_p curves in Fig. 29. Here, each curve is for a fixed bias value and represents the plate-current changes for different plate voltages. As an example, notice that with a control grid bias of —3 volts, the plate current (marked I_p) remains almost constant at 4 milliamperes for plate voltages ranging from 160 to 400 volts. Therefore, changing the plate voltage has but little effect on the plate current in this range. You will recall that our definition of amplification factor is: *the plate-voltage change necessary for a plate-current change, divided by the grid-voltage change that will produce the same plate-current change.* Therefore, since the control grid has its normal effect on the plate current, but the plate voltage has little effect on the plate current, the amplification factor is far greater.

At the same time, the a.c. plate resistance of the screen grid tube is extremely high. The plate resistance

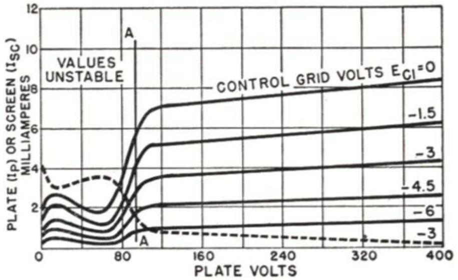


FIG. 29. A family of plate-voltage, plate-current curves for a screen-grid tube.

of the average triode is about 10,000 ohms, but the r_p of screen-grid tubes ranges around 400,000 ohms. The μ of triodes is from 3 to 20, that of screen grid tubes about 400. Surprisingly, the mutual conductance (G_m) of both is about the same value—near 1000 micromhos.*

With resonant circuits, the maximum gain is reached when the load impedance equals the tube plate resistance. This limit can be reached with triodes, which puts a limit on the Q value we can use. However, the very high plate resistance of a screen-grid tube cannot be reached, so we can use higher Q's (higher loads) and hence can get a gain of from 50 to 100 from such a tube, which is many times higher than that of the average triode tube. Thus, the screen-grid tube gives far greater gain than the triode and does not have the feedback troubles that the triode has.

Secondary Emission. Although the screen-grid tube appears to be a good solution to the problem of getting a practical r.f. amplifier, it does have a disadvantage. At low plate voltages, particularly when the plate voltage falls below the screen-grid voltage, the plate-current values are very erratic. In fact, at some plate voltages, the plate current can even decrease as the plate voltage is increased!

This effect is caused by the screen grid. It pulls electrons out of the electron cloud at a terrific rate of speed. These electrons fly to the plate and hit it so hard that they knock other electrons out of the plate. In other words, the electron bombardment of the plate forces it to emit electrons also, a phenomenon known as *secondary emission.* If the plate voltage is much higher than the screen voltage, all these secondarily emitted electrons are attracted right back to the plate; but if the plate voltage is low, some of them are attracted to the screen grid. Naturally, any such loss of electrons from the plate causes a reduction in the plate current.

This effect is illustrated in Fig. 29. Notice that in the region to the left of the vertical line A-A, the chart is marked "values unstable." As you check the plate current lines (the solid-line curves), you will find that there are a number of dips and humps to the left of the line A-A and that these lines straighten out only to the right of this point.

The line A-A represents the screen-grid voltage for the particular tube to which Fig. 29 refers. On this same chart you will find a dashed curve that represents the screen-grid current. As you will notice, to the left of the screen-grid voltage line, the screen current is high because the screen attracts secondary electrons. Then, as the plate voltage is increased, the screen-grid current drops gradually to a lower and lower value, because the plate is then capable of holding all its electrons.

Of course, this secondary emission is not troublesome if we maintain the plate voltage well above the screen-grid voltage. Remember, however, that the actual plate voltage is equal to the supply voltage minus the voltage dropped across the plate load. This drop may be enough to bring the actual plate voltage below the screen-grid voltage, causing secondary emission and a distorted signal wave form.

PENTODE TUBES

Tube designers eliminated the effects of the secondary emission characteristics of the tetrode tube by the introduction of the pentode. The pentode has all the advantages of the tetrode without the secondary emission characteristic, which is eliminated by inserting another grid between the screen grid and the plate. Since this grid suppresses the effects of secondary emission, it is called the *suppressor grid.*

$$*G_m = \frac{\mu}{r_p} \times 1{,}000{,}000.$$

$$\text{triode } G_m = \frac{10}{10{,}000} \times 1{,}000{,}000 = 1000.$$

$$\text{tetrode } G_m = \frac{400}{400{,}000} \times 1{,}000{,}000 = 1000.$$

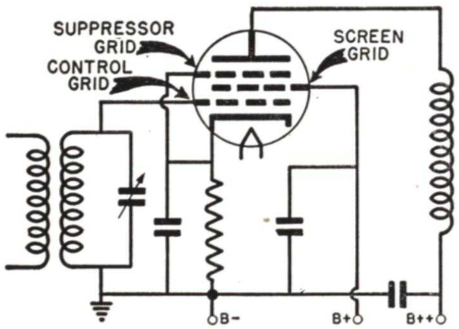


FIG. 30. The pentode tube.

This new grid is almost always connected to the cathode as shown in Fig. 30. In fact, this connection is used so universally that some tubes have the connection made internally—the suppressor grid then does not come to a prong on the tube base. Since this grid is at the cathode potential, it will always be negative with respect to the plate. Therefore, any secondary-emission electrons that try to leave the plate (moving toward the screen grid) are repelled by the suppressor grid and forced to return to the plate.

There are two reasons why this grid does not affect the electron flow from the cathode to the plate. First, the grid is coarse—it has but few turns of wire, and these are placed in the "electrical shadow" of the screen grid. That is, the suppressor wires are directly behind the screen-grid wires and are therefore not directly in the electron path. Second, electrons traveling from the cathode toward the plate are going at a very fast rate, too fast to be bent out of their paths by this wide-meshed grid. On the other hand, the secondary emission electrons are moving slowly, with the result that they feel the full effect of the suppressor grid.

We might compare the suppressor grid to a thin sheet of steel, which is easily pierced by a bullet from a high-powered gun but which is able to stop a slow-speed bullet. This one-way action of the suppressor grid is so efficient that it is even possible to operate the screen grid up to the same potential as the plate. Such an arrangement is used particularly in a.c.-d.c. receivers, where the maximum plate supply voltage is low; this improves the rate of electron flow to the plate without affecting secondary emission.

Typical characteristics of a pentode tube are shown in Fig. 31. As you can see, these characteristics do not change noticeably until the plate voltage of the tube becomes extremely low. Therefore, the plate voltage of the pentode can be made to swing over a very wide range without distortion from secondary emission. For this reason, the pentode tube has replaced the screen-grid tube as an r.f. amplifier in modern receivers, just as the screen-grid tube earlier replaced the triode.

VARIABLE-MU OR SUPER-CONTROL TUBES

The early screen-grid and pentode tubes all had a relatively sharp cut-off characteristic like that shown by curve A in Fig. 32. Notice that a bias of just a few volts will cut off the plate current. The operating point is on the straight portion of the curve, but the amount of signal input is limited. Strong signals easily produce an interference known as cross modulation.

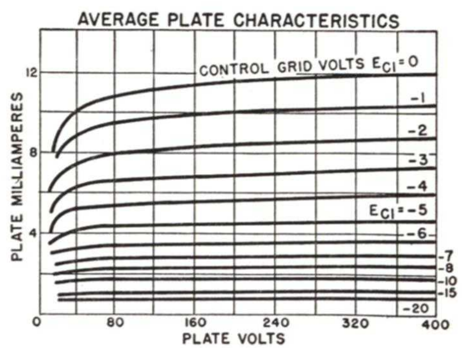


FIG. 31. The curves for an r.f. pentode tube.

In addition, it is desirable to control the signal level (the volume) before the signal reaches the demodulator. The sharp cut-off characteristic interferes with any volume control method involving a change in operating voltage. Let's investigate these problems further and learn the solution worked out by tube manufacturers.

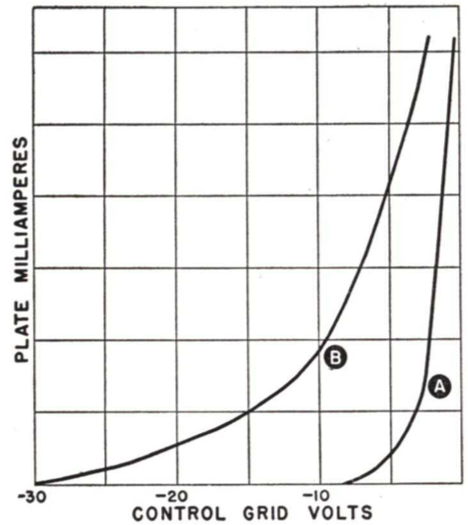


FIG. 32. A comparison between sharp cut-off and remote cut-off characteristics.

Cross Modulation. Cross modulation is said to occur when two modulated signals combine in a stage to produce one signal that contains both modulations. Obviously, this can happen only if the two signals get into the same stage. In addition, either the stage (or more accurately, the tube used in it) must have a curve in the part of its characteristic over which the signals swing, or one of the two signals must be strong enough to drive the tube to cut-off on the negative signal alternations. In other words, the tube must operate in some non-linear manner for cross modulation to occur.

This kind of defective operation occurs most frequently in locations where there is a nearby station so powerful that its signal can force its way through the tuned circuits when they are tuned to another signal.

A tube that has a sharp cut-off characteristic has a sharp bend in its curve and is easily overloaded. Cross modulation is therefore rather likely to occur in stages in which such tubes are used.

Volume Control. It is desirable to control the volume of the signal in the r.f. amplifier, because otherwise it is possible to overload the demodulator and the low-frequency amplifier. Many different kinds of volume-control circuits have been used, as you will learn in another Lesson. Some of these controls vary the amount of signal voltage fed from stage to stage, others work by changing the operating potentials of one or more tubes.

As you know, changing the grid bias or plate voltage will change the slope of the operating curve of the tube. This will change the mutual conductance of the tube. In other words, the stage gain can be varied by changing the grid bias or plate voltage to a sufficient degree. However, if the tube has a sharp cut-off characteristic, even a slight increase in the bias will cut off plate current altogether. This gives a very sharp-acting control that is not desirable.

The Solution. From what we have just said, you can see that it is undesirable to use tubes having sharp cut-off characteristics as r.f. amplifiers. Tube manufacturers have therefore devised special pentodes for r.f. and i.f. stages. In these tubes, the grid structure is changed so that different sections of it will have different amounts of control over the plate current. Several constructions have been tried, but the one shown in Fig. 33 has replaced all others.

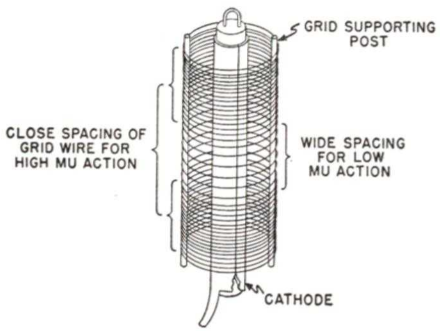


FIG. 33. The basic variable-mu grid structure.

When the control-grid wires are close together, only a small voltage is needed to block all electron flow. This means that a tube having such a grid will have a sharp cut-off, and, since the grid has a great control over the plate current, the amplification factor can be made high. On the other hand, wide spacing between the grid wires provides less control; a high bias is necessary for plate current cut-off, and a lower amplification factor results.

The construction shown in Fig. 33 provides both high and low amplification. As the grid bias is increased, the closely spaced sections tend to cut down the plate current sharply. This gives a high amplification factor (high-mu) action within this bias range. However, in the center section of the grid, we have relatively wide spacing. It takes a high bias to cut off the plate current through this section. Therefore, the high-mu section cannot cut off plate current completely. Instead, there is a gradual change-over from a high-mu action to a low-mu action as the bias is increased.

Curve B in Fig. 32 shows how the plate current of a typical variable-mu (or remote cut-off) pentode varies with the bias voltage. Compare curve B with the sharp cut-off curve A. Notice that the sharp cut-off tube has practically zero plate current at a bias of —8 volts. On the other hand, the variable-mu tube characteristic curve B shows considerable plate current at this bias; in fact, it takes a bias of —30 volts to cut off the plate current of this tube. Notice, also, that the slope of the characteristic curve changes gradually from nearly vertical to nearly horizontal. The more nearly vertical the curve, the greater the plate-current change for a given grid-voltage change, and hence the higher the G_m of the tube. Therefore, this curve represents a tube that changes in G_m with bias voltage changes.

The gradual curve changes and the remote cut-off value make this tube less susceptible to cross modulation. Furthermore, this tube is well suited for use in a volume-control circuit that changes the bias; if the bias is increased, the tube has lower G_m and hence less output.

This particular characteristic has permitted the development of the modern "automatic volume control" circuit. (This circuit is poorly named, because it is actually an automatic gain control; the volume of sound coming from the speaker must still be adjusted manually.) In the automatic volume control (a.v.c.) circuit, the signal voltage fed to the demodulator develops a d.c. voltage that is used to bias the r.f. and i.f. amplifier tubes. If the circuit parts are properly chosen, this bias will keep the input to the demodulator almost constant. If the signal is stronger than average, the bias will be increased to cut it back to an average level; if it is weaker than average, the bias will be reduced to permit it to come up to an average level. You will learn more about this circuit and other methods of volume control in later Lessons.

METAL AND SINGLE-ENDED TUBES

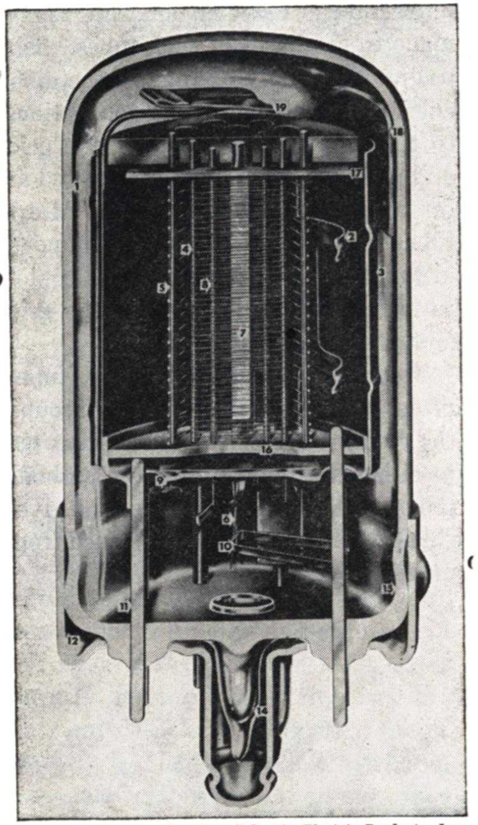
As we have said, grid-plate capacity (which screen-grid and pentode tubes largely eliminate) is not the only undesirable capacity that may exist in radio circuits. Stray capacitive coupling between parts and even stages can be troublesome. To reduce such coupling, r.f. tubes in glass envelopes are enclosed in shield cans; sometimes even the grid leads have to be shielded. A tube with a metal envelope does not need a shield because the metal envelope can be grounded to act as one.

Now that more is known about proper positioning of tube and circuit elements, single-ended tubes (which were at one time all but discarded in favor of tubes having a top-cap grid connection) have come back into almost universal use. Since all connections are made to the bases of such tubes, their use automatically eliminates the difficulties involved in making connections to tube top caps. Of course, the grid and plate connections are separated as far as possible on the socket. Many tubes with octal bases have a grounded metal shield located in the center "locating" pin that is between the grid and plate terminals.

Modern single-ended glass tubes often have a built-in shield around the elements. This shield is connected to either a ground pin or the cathode pin on the base. This construction gives shielding as good as that obtained from metal tubes.

Introduction to V. H. F. Stages

In the early days of radio, the serviceman had only the radio frequencies in the broadcast band to deal with. Then receivers were built that received short-wave bands also. The use of these bands brought up few new problems; the receivers were the same as they had been before except that they contained new coils that allowed the new tuning ranges to be covered.



Courtesy Sylvania Electric Products, Inc.

An interior view of a modern loktal tube. The parts identified by numbers are: 1, bulb; 2, plate; 3, metal shield enclosing the elements; 4, grid #2; 5, grid #3; 6, heater; 7, cathode; 8, grid #1; 9, bottom shield; 10, connectors; 11, pins; 12, base; 14, exhaust tube; 15, stem header; 16, bottom mica; 17, top mica; 18, dome pads; 19, getter. The base of this tube is a metal shell that acts as a shield for the leads. The leads do not touch the base when they emerge from the tube; they come out through the glass bosses that protrude through holes in the base.

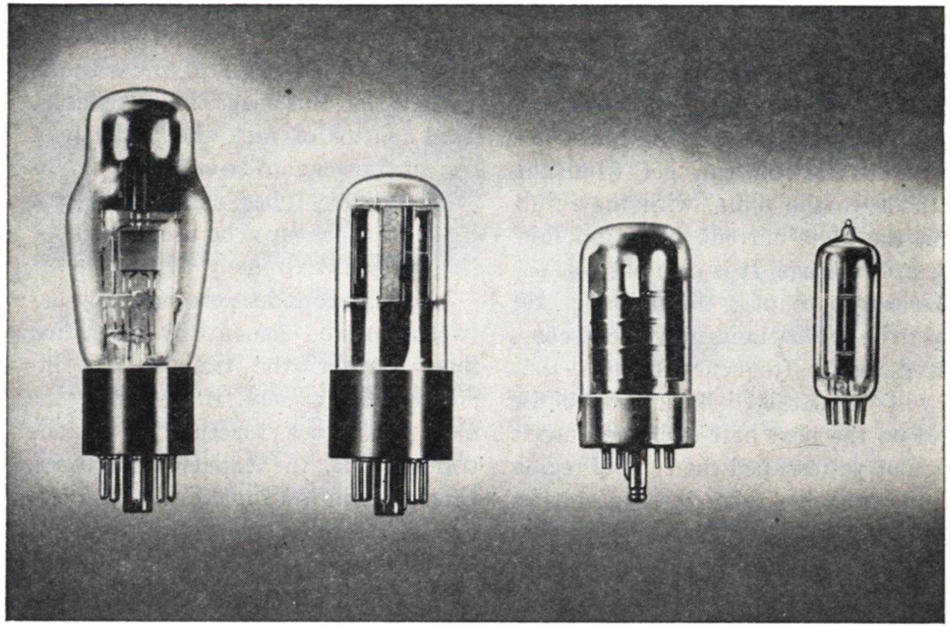
Today, however, the serviceman must also be prepared to handle frequency modulation and television as well. Both these services transmit in the very-high-frequency (v.h.f.) band, the use of which introduces many new problems. Let's see what some of these are.

CAPACITY REDUCTION

The interelectrode capacities within the tube are across the resonant circuits, and are a serious problem at very-high frequencies. When the frequency is high, this capacity may be equal to or greater than the capacity needed for tuning to the frequency desired. Reducing the resonant circuit inductance will permit a higher frequency to be reached, but there are definite limits to the amount by which the inductance can be reduced. Once our "coil" is only a short, straight piece of wire, the only way we can use ordinary circuits and go to higher frequencies is to reduce the internal tube capacities.

The capacity between the tube elements can be reduced by increasing the separation between the elements or reducing their size. This applies not only to the tube elements themselves, but also to their connecting leads.

In the ordinary tube, there is considerable capacity between the prongs and between the long leads from these prongs up to the tube elements. (These leads are two inches long in some tubes.) The first step in reducing ca-



Courtesy Sylvania Electric Products, Inc.

FIG. 34. This illustration shows the same tube type in four different forms. At the left is the standard octal-base glass tube; next to it is the GT tube; next is the loktal tube; and at the right is the miniature tube.

pacity is to bring the elements down closer to the tube base. The modern GT tubes have shorter leads from the prongs to the elements than do some of the earlier tubes.

The next step is to eliminate the base itself. This removes some of the solid dielectric material between the tube prongs and reduces the capacity to what it is with air as the dielectric. The loktal tube is an example of a tube from which the base has been removed; the prongs of this tube are also reduced greatly in size, which has the same effect as spacing them farther apart, although the distance between their centers remains the same.

A further step in this direction is the miniature tube shown in Fig. 34. This tube has no base attached to the glass; the glass itself is pressed and holds the extension of the leads to form the prongs to fit the socket. In these tubes, the elements are reduced physically in size and are brought closer to the connecting pins.

TRANSIT TIME

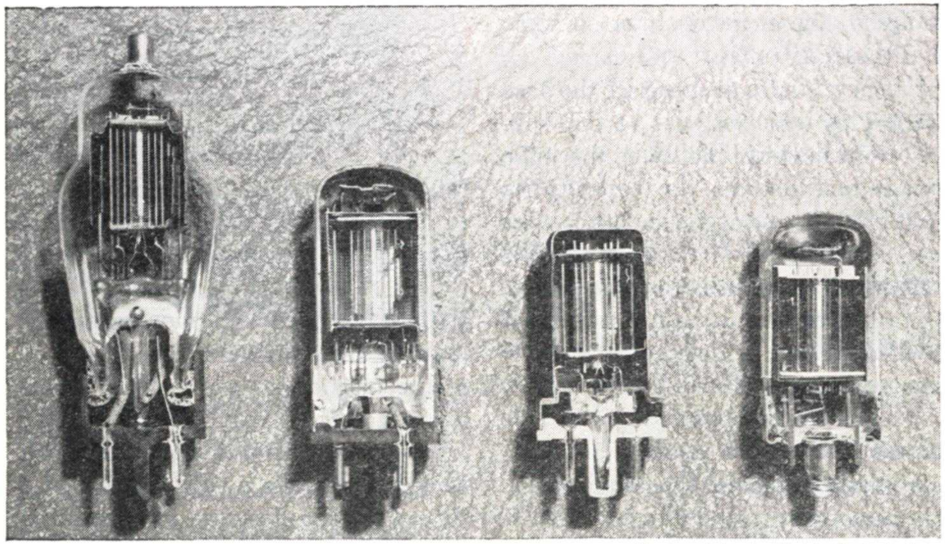
Another problem at very-high frequencies is the *transit time* within the tube. (The time needed for an electron to travel from the cathode to the plate is called the "transit time.") Ordinarily, we think of the electrons as leaving the cathode and moving to the plate at a rapid rate of speed. In fact, at the frequencies with which we ordinarily deal, we can consider that the trip is practically instantaneous. However, it does take a tiny fraction of a second for the electrons to move from the cathode to the plate; the farther these elements are apart, the longer it will take.

At the very-high frequencies, the transit time becomes important. It is possible for the frequency to be so high that the grid will reverse in phase before the electron can get from the cathode to the plate, with the result that the plate current will not follow the grid voltage. It is also possible for an electron to pass through the grid but to be so slow in its movement, compared to the frequency change, that it will be attracted back toward the grid on the next half-cycle because it has not yet escaped the control region of the grid. To solve these problems, some way must be found to reduce the transit time.

The electrons can be speeded up by using higher voltages on the plate, but there are definite limits to the plate voltage that can be used. A better solution is to reduce the space between the cathode and plate so that the electrons will have to move but a short distance. Naturally, there are also limits to the amount that this space can be reduced. As a matter of fact, the problem of transit time cannot be solved by modifying ordinary tubes: entirely different kinds of tubes must be used when the frequency is extremely high.

If we try to reduce the capacity between the tube elements by separating them farther, the transit time increases. Conversely, if we bring the elements closer together to reduce transit time, the capacity increases. One solution is to make the tube elements so small that they can be brought close together (thereby reducing transit time) without increasing the capacity unduly.

The factor of transit time sets a definite limit to the frequencies at which pentode and screen-grid tubes can be used, because, in these tubes, the plate



Courtesy Sylvania Electric Products, Inc.

Four different styles of the same tube type, cut away to show the internal structure. From left to right, they are: an octal-base glass tube (type 6J7G), a GT tube (type 6SJ7GT), a metal tube (type 6SJ7), and a loktal tube (type 7C7). Although the loktal tube has a different number, it is the electrical equivalent of the others.

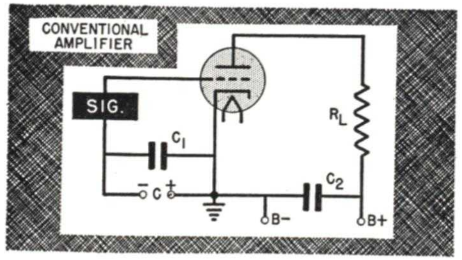


FIG. 35. A standard triode amplifier.

must be a fair distance from the cathode to allow room for the screen and suppressor grids. Triode tubes can be and are used at very-high and ultra-high frequencies because the closer spacing between the cathode and plate permits a reduction in transit time. Although, of course, the triode tube does not have the amplification of the screen-grid tube, it is the only type that can be used at these frequencies, and it does give at least fair gain.

Grounded-Grid Amplifier. The problem of oscillation returns with the triode, so neutrodyne circuits are back in use in v.h.f. circuits. A modern amplifier circuit known as a *grounded-grid* amplifier is also popular because this arrangement prevents oscillation without using the neutrodyne principle. Let's see how.

Fig. 35 shows a standard amplifier circuit in which a triode tube is used.

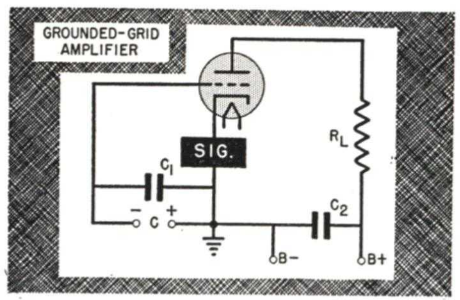


FIG. 36. The grounded-grid amplifier.

Notice that the signal is fed in between the grid and cathode. For simplicity, a separate C bias source is indicated.

This circuit works exactly like the triode circuit described earlier in this Lesson. Regeneration may occur, because part of the voltage developed across the load is fed back through the grid-plate capacity to the grid circuit.

For comparison, look at Fig. 36, which shows the circuit known as the *grounded-grid* amplifier. Notice that the signal source is inserted between the *cathode* and ground and that the

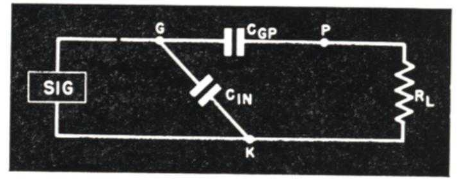


FIG. 37. The standard triode circuit has the feedback path shown here from load to input.

control grid of the tube is grounded directly through condenser C_1. (This, of course, still leaves the signal source connected between the cathode and the grid.) With this circuit arrangement, the voltage across R_L cannot be considered to be across the signal source at all. Let's see why.

Fig. 37 shows the equivalent of the standard triode amplifier of Fig. 35. The voltage across R_L is applied directly through C_{GP} across the signal source.

Fig. 38 shows the equivalent of the grounded-grid circuit. The signal source is still shunted by the input capacity of the tube, which is the C_{GK} capacity within the tube. However, the capacity C_{GP} does not feed back voltage from across the load.

To see this, trace the load connections in Fig. 36. One end is connected to the plate, and the other end is connected to ground through condenser C_2. Therefore, since the control grid is effectively grounded by C_1, the load

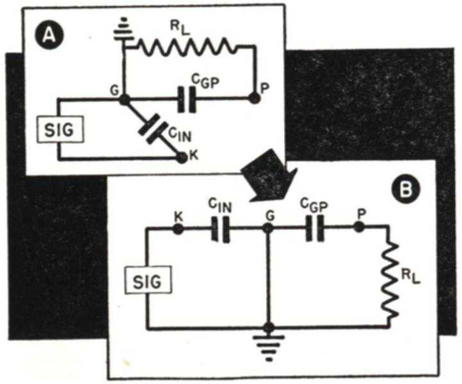


FIG. 38. The equivalent circuit of the grounded-grid amplifier.

connection is equivalent to that shown in Fig. 38A; with this arrangement, the voltage developed across R_L cannot be considered to be across the input signal source.

Rearranging the circuit somewhat (without making any real change in it) gives us Fig. 38B. As you can see from this, the grid now acts as a shield between the cathode and plate, producing much the same effect as a screen grid.

Naturally, we do not have the increased amplification of the screen-grid tube in this circuit, but we do reduce the input capacity and eliminate the normal triode regeneration effect. Therefore, it is possible to use a triode tube in this circuit at high frequencies.

Actually, of course, it is possible to use the triode with this circuit even at ordinary frequencies. However, at ordinary frequencies, we can use the pentode tube and take advantage of its higher gain. Therefore, this triode tube circuit is not used except where the pentode circuit has some definite disadvantage.

We'll leave other v.h.f. circuits for later Lessons, where you take up the problems of f.m. and television amplifiers more fully.

Lesson Questions

Be sure to number your Answer Sheet 16 FR-2.

Place your Student Number on every Answer Sheet.

Send in your set of answers for this Lesson immediately after you finish them, **as instructed in the Study Schedule. This will give you the greatest possible benefit from our speedy personal grading service.**

1. In addition to amplification, what important task is performed by an r.f. stage? TO SELECT A DESIRED FREQ.

2. What happens if the r.f. amplifier cuts off some of the side-band frequencies?
FIDELITY WILL BE SACRIFICED

3. If an L-C circuit has a very high Q, will the pass band be: *very narrow; of medium width;* or *very broad?*

4. If two r.f. stages having resonant gains of 15 each are connected in cascade, will the resulting gain at resonance be: 15; 30; 75; or 225?
15 X 15 = 225

5. If two r.f. stages having similar response curves are connected in cascade, is the resulting pass band: *wider; narrower;* or *the same as* for one stage?

6. If a double-tuned circuit is adjusted to give a double-humped band-pass response, are the fidelity and selectivity: *worse than; the same as;* or *better than* they are for a single-peaked response?

7. What drawback do triode tubes have that makes them undesirable for r.f. amplifiers at broadcast frequencies?
THE INTERELECTRODE CAPACITY THAT EXITS BETWEEN THE GRID + PLATE WITHIN THE TUBE

8. Why has the pentode replaced the screen-grid tube as an r.f. amplifier in modern receivers? THE PENTODE CAN BE MADE TO SWING OVER A WIDE RANGE WITHOUT DISTORTION FROM SECONDARY EMMISSION

9. What two problems make it difficult to use ordinary tubes at very-high frequencies? THE CAPACITY IS GREAT WITHIN THE TUBE AND THE TIME ALLOWED FOR ELECTRONS TO FLOW FROM CATHODE TO PLATE.

10. Why are triode tubes better than pentodes at extremely high frequencies?
BECAUSE THERE IS A CLOSER SPACING BETWEEN THE CATHODE + PLATE.

FEAR LEADS TO FAILURE!

No matter how hard a person may work for success, there is nothing which can help him if he is always doubting his own ability — if he is always thinking about failure.

To be ambitious for wealth yet always expecting to be poor is like trying to get past a vicious dog when afraid of the dog and uncertain of your ability to make friends with him — in each case, fear of failure is almost certain to result in failure. Success, on the other hand, is won most often *by those who believe in winning.*

Never doubt for a moment that you are going to succeed. Look forward to that success with just as much assurance as you look forward to the dawn of another day, *then work — with all that's in you — for success.*

J.E. Smith