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CIRCUIT COMBINATIONS THAT PROVIDE SUBSTANTIALLY UNIFORM SIGNAL SELECTION†

Band-Pass Filter Circuits Combined With Tuned Resonance Circuits in Cascade

By E. A. UEHLING*

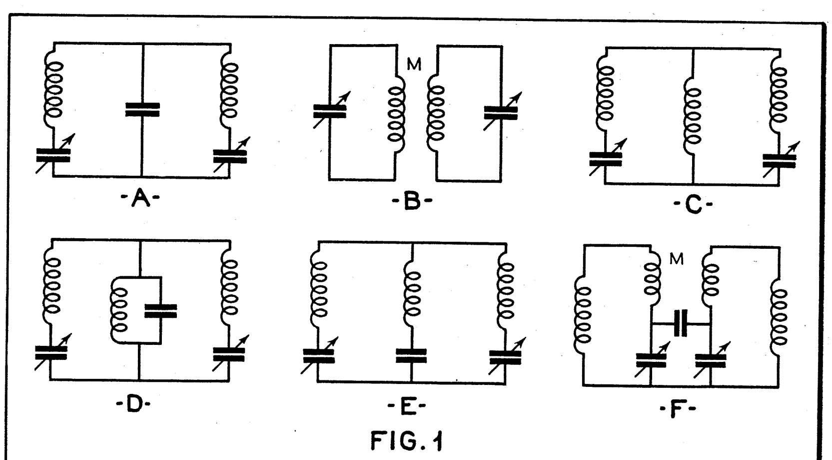
THERE has been a marked tendency in recent years in the design of radio broadcast receivers toward the separation of the functions of signal selection and radio-frequency amplification. This trend began a few years ago, when the majority of manufacturers discarded the coupling tube method of supplying the signal in the antenna to the receiver, and has now gone so far that laboratory models have been constructed and commercial receivers have been built in which the complete function of signal selection is accomplished before any amplification of the signal is permitted to take place. This separation of functions has been accomplished through the use of the band-pass filter. Whether or not this separation of functions should be complete remains to be seen after a more thorough study of the band-pass filter has been made. Such filters as are generally used have certain inherent disadvantages which in a practical radio receiver must be eliminated and which in general are most easily eliminated by combining the band-pass filter with other circuits.

"Phantom Stations"

The desirability of adequate signal selection and filtering before amplification is well known to everyone. Due to the non-linear characteristics of amplifier tubes, a certain amount of rectification of the signal takes place in the first stage of amplification as well as in the following stages resulting in a small rectified current available for the modulation of other carriers if they are present. In the neighborhood of a strong broadcast station, as a result, a carrier to which the receiver is tuned will sometimes be modulated by the audio-frequency signal of the partially rectified carrier of the local broadcast station, and both signals will be heard together. This modulation can take place in another way. Two stations whose carrier frequencies are separated by such an amount that their difference falls in the broadcast band can often be heard together when the radio receiver are not impaired, and they quency. The result is the well-known phenomenon of "Phantom Stations." In order to avoid this condition, a more thorough separation of signals is required before any amplification takes place.

The band-pass filter is well known for its selective properties. It has other advantages as well, among them being the flat top characteristic which so adequately permits the transmission of sidebands. As a consequence, sideband cutting is reduced to a minimum, yet the selective properties of the receiver are not impaired, and they may, in fact, be improved. A broadcast receiver must, however, be capable of receiving carriers over a wide range of frequencies. It is here that trouble is encountered in the use of the band-pass filter. Its characteristics at one end of the frequency range may be entirely unlike its characteristics at the other end. A variation in band width with change in frequency is the most evident of these varying characteristics. There are others as well, among them being a marked change in the ratio of the input and output voltages if the coupling between the circuits of the band-pass filter is permitted to have a value less than the critical value at any frequency in the broadcast range. Occasionally there are conditions under which it is desirable to design the circuits in this way.

A band-pass filter in its most simple form consists of two circuits of positive and negative reactance coupled together by means of either a positive or negative reactance or a combination of one or both. Two such circuits, coupled by a negative reactance, are shown in Fig. 1-A. In Fig. 1-B and Fig. 1-C, respectively, similar circuits coupled by a positive reactance are shown; in Fig. 1-D and 1-E the same circuits having a coupling between them consisting of a combination of reactances, and in Fig. 1-F the same circuits again with capacitative and magnetic coupling. The discussion in this paper relating to band-pass circuits will be confined entirely to the two-section type of structure.



Typical band-pass filter circuits.

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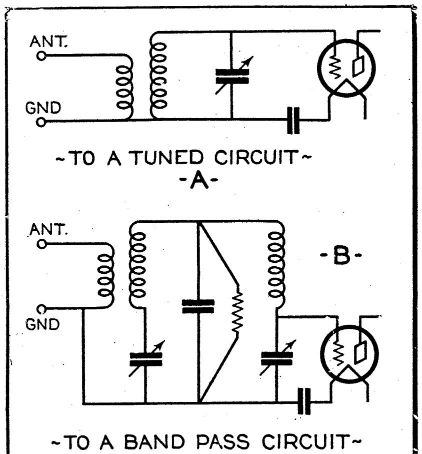


FIG.3

Two methods of antenna coupling.

Two-Section Band-Pass Filter

One of the properties of a two-section band-pass filter as shown in these figures is that to an e.m.f. placed in one of the circuits there will be two frequencies at which the alternating e.m.f. and current will be in phase. These two frequencies occur when the reactance of the first circuit X_1 and of the second circuit X_2 is expressed by

$$X_1 = \pm \sqrt{\frac{R_1}{R_2}(B^2 - R_1 R_2)}$$

and

$$X_2 = \pm \sqrt{\frac{R_2}{R_1}(B^2 - R_1 R_2)}$$

where B is a real quantity and is equal to the coupling reactance, and R_1 and R_2 are the resistances of the first and second circuits, respectively. The coupling between the circuits must be sufficient in which case

$$B^2 > R_1 R_2$$

It is at these two values of reactance given by the above equation, the positive and the negative value, that the two peaks of the band-pass transmission characteristic occur. The radio engineer is interested in the separation of these peaks; for on this separation the selectivity of the receiver and the degree of side-band cutting largely depends. In general, R_1 and R_2 are equal, or nearly so, when loosely coupled to the energizing circuit, so that

$$X_1 = X_2 = \pm \sqrt{B^2 - R^2}$$

The frequency separation of the two peaks can be easily determined. One peak occurs when the reactances X_1 and X_2 are given the positive values of the above equation, and the other peak occurs when the reactances X_1 and X_2 take on the negative value. To determine the frequency separation of the peaks it is necessary only to determine the rate of variation of the reactances with changes in frequency. This is done for an ordinary series circuit which is a generalized form of each of the circuits of the band-pass circuit.

$$X = \omega L - \frac{1}{\omega C}$$

$$\frac{dx}{d\omega} = L + \frac{1}{\omega^2 C}$$

$$\frac{dx}{df} = \frac{dx\, d\omega}{d\omega\, df} = 2\pi L + \frac{2\pi}{\omega^2 C}$$

at resonance

$$\frac{1}{\omega^2 C} = L$$

Then

$$\frac{dx}{df} = 4\pi L$$

$$\frac{df}{dx} = \frac{1}{4\pi L} \qquad (1)$$

Thus, for a small change in frequency near the resonant point of the circuit, the change in frequency per unit change in reactance is equal to $\frac{1}{4\pi L}$. The change in frequency per unit change in reactance is independent of the original value of ω.

As the generated frequency is varied from the value it has at one peak of the transmitted band to its value at the second peak of the transmitted band, the reactance of the individual circuits X_1 and X_2 changes from $\pm\sqrt{B^2 - R^2}$ through zero to $\mp\sqrt{B^2 - R^2}$. The total change in reactance is then $2\sqrt{B^2 - R^2}$. Integrating equation (1)

$$\frac{df}{dx} = \frac{1}{4\pi L}$$

we have

$$f = \frac{X}{4\pi L}$$

Therefore, the frequency separation of the points of maximum transmission is

$$f = \frac{X}{4\pi L} = \frac{2\sqrt{B^2 - R^2}}{4\pi L} = \frac{\sqrt{B^2 - R^2}}{2\pi L} \qquad (2)$$

The narrower the band width, or the smaller the quantity $\sqrt{B^2 - R^2}$ the greater is the accuracy of this equation. The error introduced by spreading the value of the derivative over the width of the band is, however, very small when radio circuits are considered because of the relatively small value of f as compared with f_r the resonance frequency of the individual circuits.

The frequency width of the transmitted band is usually given by equations which neglect the circuit resistances. According to these formulas the lower frequency peak is given by

$$f_1 = \frac{f_r}{\sqrt{1 + K}}$$

and the higher frequency by

$$f_2 = \frac{f_r}{\sqrt{1 - K}}$$

where f_r is the resonant frequency of one circuit taken alone and K is the coefficient of coupling

$$K = \frac{M}{\sqrt{L_1 L_2}}$$

If

$$L_1 = L_2 = L$$

$$K = \frac{M}{L}$$

Then f, the frequency width of the band is

$$f = f_2 - f_1 = f_r \left[\frac{1}{\sqrt{1 - K}} - \frac{1}{\sqrt{1 + K}}\right] \qquad (3)$$

Circuit Resistance

But the radio engineer cannot afford to neglect circuit resistances when dealing with band-pass filters. This is especially true considering the relatively narrow width of the transmitted band of frequencies and the high value of the radio frequency to which it is referred. Considered in this way, the coupling appears very close to the critical value and in this range the circuit resistances play an

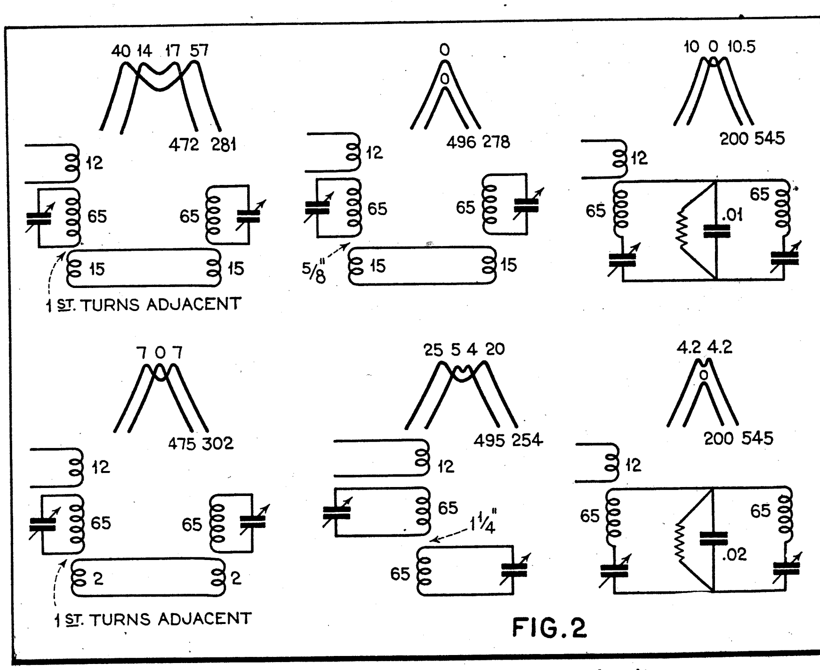


FIG.2

Transmission characteristics of coupled circuits.

extremely important role. It will be evident that for zero circuit resistances all values of coupling are greater than the critical value for $(B^2–R^2)$ would always be greater than zero.

It will be interesting, nevertheless, to show that equation (3) is approximately equal to equation (2) when the resistance term in equation (2) is dropped. The two fractions of equation (3) can be expanded into an infinite series of which all but the first few terms are negligible.

$$\frac{1}{\sqrt{1-K}} = 1 + 1/2\ K + 3/8\ K^2 + \frac{15}{48} K^3 + \frac{105}{384} K^4 + \ldots\ldots$$

$$\frac{1}{\sqrt{1+K}} = 1 - 1/2\ K + 3/8\ K^2 - \frac{15}{48} K^3 + \frac{105}{384} K^4 - \ldots\ldots$$

and

$$f_r = \frac{\omega_r}{2\pi}$$

Substituting these values in equation (3) and neglecting all terms in which K appears to the second and higher degrees

$$f = \frac{\omega_r}{2\pi}\left[1 + \frac{K}{2} + \ldots\ldots - 1 + \frac{K}{2} \ldots\ldots\right]$$

$$= \frac{\omega}{2\pi} K = \frac{\omega M}{2\pi L}$$

Neglecting the resistance term in equation (2)

$$f = \frac{\sqrt{B^2 - R^2}}{2\pi L}$$

we have

$$f = \frac{B}{2\pi L}$$

and if inductive coupling is used $B = \omega M$ and $f = \frac{\omega M}{2\pi L}$ establishing the approximate equality of the two equations.

Having determined the width of the transmitted band of frequencies and the factors affecting this width, it remains to determine what variations exist when the circuits are tuned to different carrier frequencies in the broadcast range without introducing any change in these circuits except to vary the tuning condensers as is ordinarily done in tuning a receiver made up alone of tuned circuits in cascade. In the process of tuning, L remains unchanged. B, the coupling impedance, will, however, vary, and R will vary. If the quantity $(B^2–R^2)$ can be maintained constant, the transmitted band width will remain constant. R in general increases in value with frequency. B increases in value with frequency if the coupling is inductive and decreases if the coupling is capacitative. It might appear on first consideration that inductive coupling would give more uniform width of band than capacitative coupling. It must be remembered, however, that we are interested not only in the direction in which the coupling impedance varies, but the rate at which it varies

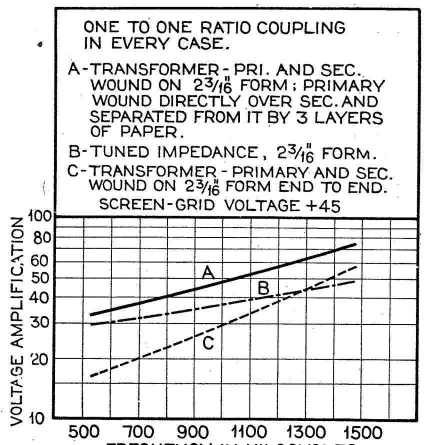


FIG. 5

Amplification of one screen-grid tube with transformer and tuned impedance coupling.

with frequency. The value of quantity $(B^2–R^2)$ may be substantially constant though the individual quantities are varying in opposite directions, provided that the rate of variation of these quantities is actually low. Such is the case under certain conditions.

The rate of variation of B for inductive coupling is

$$\frac{d\,B}{d\,\omega} = \frac{d\,[\omega\,M]}{d\,\omega} = M \qquad (4)$$

The rate of variation of B for capacitative coupling is

$$\frac{d\,B}{d\,\omega} = \frac{d\left[\frac{-1}{\omega\,C}\right]}{d\,\omega} = \frac{1}{\omega^2\,C} \qquad (5)$$

In an average radio circuit of fairly low resistance coils and for 10,000-cycle band width at 500 kilocycles, M for inductive coupling, and C for capacitative coupling will have a value such that the values of the two derivatives just given will be approximately the same. As the frequency is increased, the value of equation (4) remains unchanged, which means that for inductive coupling the band width increases constantly as the frequency is increased, the resistance term being more nearly negligible as the peaks of the transmitted band separate from one another. On the other hand, when capacitative coupling is used, we must use equation (5), the value of which decreases with increasing frequency, which means that the rate of decrease of the width of the band as the frequency increases becomes less and less.

Capacity Coupling

While neither type of coupling is wholly desirable from the point of view of changes in band width with changes in frequency, capacity coupling under certain conditions offers the greater possibilities. This is especially true if the band-pass filter is to be combined with other resonant circuits in cascade. Such a combination first of all improves the characteristics of the band-pass filter by flattening the portion of the transmitted curve between the two peaks. In this way the advantages of both types of circuits can be utilized and the combination in most respects is very desirable. A little further study of the characteristics of a combined band-pass filter and resonant circuits will disclose the important advantages which exist in favor of capacity coupling. As already shown, the transmission band becomes more narrow as the frequency is increased. The selectivity of ordinary resonant circuits, however, decreases as the frequency is increased because of the increasing resistance of the circuits. To offset this decrease in selectivity, a band-pass filter that has a very narrow band at high frequencies and a 10,000-cycle band at low frequencies will give an over-all selectivity band that is of almost constant width.

Selectivity curves showing the transmission characteristics of such a combination of circuits are shown in Figs. 9 and 10. In this combination of circuits a capacity-coupled type of band-pass circuit precedes the first amplifier tube and adequately performs the function of signal selection. In the amplifier, tuned circuits are used to couple the amplifier tubes and the detector. These circuits have selectivity characteristics of their own which tend to improve that of the band-pass circuit at the longer wavelengths. At the short wavelengths conditions are reversed. The tuned circuit characteristics are broad at these wavelengths, but in combination with the capacitative-coupled band-pass circuit of extremely sharp selectivity characteristics at these wavelengths, the transmission a few kilocycles off resonance is very materially reduced. The result is a selectivity characteristic at all wavelengths that is substantially uniform.

There are still other possibilities. Radio-frequency amplification tends to

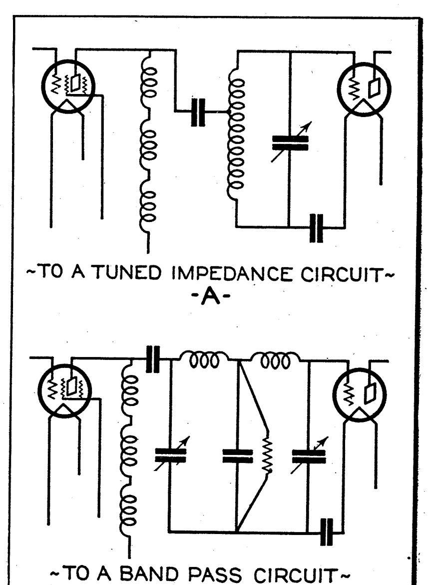


FIG. 4

Two methods of screen-grid coupling.

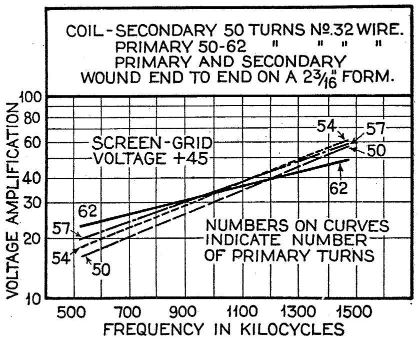


FIG. 6

Amplification of one screen-grid tube with transformer coupling.

increase as the frequency is increased. In some cases this change in amplification can be partially eliminated by the introduction of fixed regeneration that affects the higher wavelengths only, by careful design of the antenna coupling system, by self resonance in the primary circuits of the radio-frequency transformers, and by several other methods. If screen-grid tubes are used, however, this problem becomes more difficult, and if the amplification is made to meet the requirements at 500 kc. it may be so large at 1,500 kc. that over-all regeneration will produce instability. This condition can be partially removed by using a capacity-coupled band-pass filter that has its coupling impedance so adjusted as to pass through the point of critical coupling at some point in the broadcast range. At all frequencies greater than this critical frequency, the gain will be gradually reduced as the frequency is increased with the result that the receiver will have a more nearly uniform amplification curve and the possibility of oscillation at short wavelengths will not be present even though the total amplification is further increased at the longer wavelengths than could otherwise have been done.

Position of Band-Pass Circuit

When the band-pass circuit is used in this way and loosely coupled to the antenna circuit, it will supply to the grid of the first amplifier tube for a given voltage in the antenna circuit, if the coupling between the circuits is sufficient, a voltage equal approximately to one-half the voltage that would have been supplied had an ordinary tuned circuit been used. It does not follow, however, that from the point of view of maximum voltage amplification a more desirable position for the band-pass circuit could have been found. If used in the output circuit of a screen-grid tube, the voltage amplification for the stage would be only one-half the amplification obtainable if a tuned circuit had been used instead. In fact, it is more or less general that when a band-pass circuit is supplied by means of an ideal generator, i. e., one in which the generated current is independent of the load impedance, the voltage amplification is approximately one-half that obtainable if an ordinary tuned circuit had been used instead of the band-pass circuit. If a band-pass circuit is used following a -27 type tube, the loss in amplification as compared with that obtainable with an ordinary tuned circuit, other conditions being identical, is approximately 30% under the best conditions for each type of circuit. This loss in voltage amplification can be shown very easily. Referring to Fig. 3-A where the ordinary antenna circuit is shown, the current in the tuned circuit is given by

$$i_2 = \frac{\omega M e_1}{Z_1' Z_2}$$

where Z_1' is the forward equivalent impedance of the antenna circuit, that is, the impedance of the antenna circuit as influenced by the tuned circuit. But the resonant frequency of the antenna circuit is usually much higher than the highest frequency of the broadcast band and the coupling impedance is usually very small. Therefore,

$$Z_1' = Z_1 \text{ approximately}$$

and

$$i_2 = \frac{\omega M e_1}{Z_1 Z_2} \tag{6}$$

In Fig. 3-B is shown an antenna circuit coupled to a band-pass circuit. The current in the first of the band-pass circuits, the one to which the antenna is coupled, is given by

$$i_2 = \frac{\omega M e_1}{Z_1' Z_2'}$$

where Z_2' is now the impedance of the first tuned circuit as influenced by the impedance of the following circuit to which it is coupled and not that of the tuned circuit alone as before. Again,

$$Z_1' = Z_1 \text{ approximately}$$

and

$$i_2 = \frac{\omega M e_1}{Z_1 Z_2'} \text{ nearly}$$

If $R_2 = R_3$

$$i_2 = i_3 \tag{7}$$

and the equation just given holds as well for the current in the final circuit, which is the one in which we are interested.

This equation differs from equation (6) only in the term Z_2' which has replaced Z_2. In the term Z_2'

$$X_2' = X_2 - \frac{\omega^2 M^2}{Z_3^2} X_3 = O$$

and

$$R_2' = R_2 + \frac{\omega^2 M^2}{Z_3^2} R_3 = 2 R_2$$

Therefore, the denominator of the equation (7) is twice as large as that of equation (6) and the voltage ratio when the band-pass circuit is used is one-half that obtainable with an ordinary tuned circuit. Now assume that the band-pass filter had been placed after one of the screen-grid tubes. The circuit is shown in Fig. 4-B. The load impedance would be that of a tuned circuit and is equal to

$$R_L = \frac{L}{R_1' C}$$

where L is the inductance of the tuning coil, C the capacity of the tuning condenser, and R_1' the resistance of the first circuit as influenced by that of the circuit to which it is coupled. This resistance R_1' is equal to

$$R_1' = R_1 + \frac{\omega^2 M^2}{Z_2^2} R_2 = R_1 + \frac{R_1}{R_2} R_2 = 2 R_1$$

Therefore, the resistance term in the equation for R_L is twice as large for a band-pass circuit as for an ordinary tuned circuit. Therefore, R_L is reduced one-half when the band-pass circuit is used, and the amplification which is equal to

$$g = g_m R_L$$

is also reduced one-half.

R-F. With Screen-Grid Tube

Having considered the problem of signal selection and the selectivity characteristics of a particular type of band-pass filter combined with tuned radio-frequency circuits, we will now turn our attention to the problem of radio-frequency amplification using the screen-grid tube. Both impedance coupling and transformer coupling will be considered. Because of certain characteristics of the screen-grid tube, and in particular the high plate impedance of these tubes, the problem of high amplification per stage is reduced to one of obtaining a high load impedance.

The elementary equation for amplification with a single three- or four-element tube is

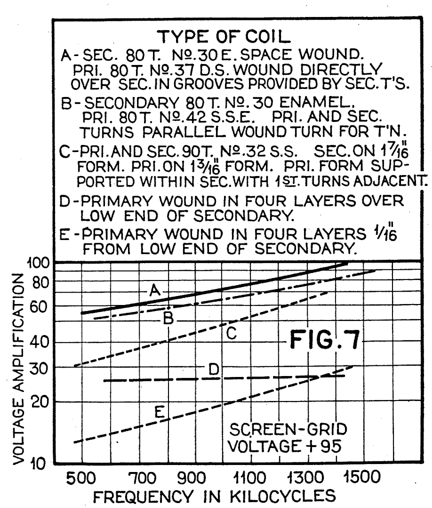
$$g = \frac{\mu R_L}{r_P + R_L}$$

This can be transformed and reduced to more simple terms by neglecting the value of R_L whenever it appears with r_P in comparison with which it is small as follows:

$$g = \frac{\mu R_L}{r_P + R_L} = \frac{\mu}{r_P} \frac{R_L r_P}{r_P + R_L}$$

$$= G_m R_L \frac{r_P}{r_P + R_L} = G_m R_L \text{ approximately.}$$

The external load impedance is usually obtained by means of a resonant circuit. If an anti-resonant circuit appears directly in the plate circuit, the



Amplification of one screen-grid tube with transformer coupling.

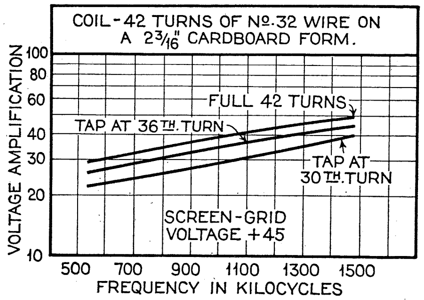


FIG. 8

Amplification of one screen-grid tube with tuned impedance coupling.

load impedance R_L is given by

$$R_L = \frac{L}{R\,C}$$

where L is the inductance of the resonant circuit, C is the capacity and R is the resistance.

If the resonant circuit is coupled inductively to a coil which is located in the external plate circuit the load impedance is given by

$$R_L = \frac{\omega^2 M^2}{R}$$

where ωM is the mutual impedance of the radio-frequency transformer and R is the radio-frequency resistance of the secondary circuit. It will be understood from this equation why ωM must be made so much larger in value when screen-grid tubes are used instead of the —27 type.

A little calculation will show that if M is made so large that it is equal in value to the inductance L of the tuning coil, the value of R_L obtained with transformer coupling is equal to that obtained with impedance coupling. To obtain this value of M the primary inductance will have to be greater in value than the secondary inductance because the coefficient of coupling K cannot be made equal to unity.

A little study of these equations will give us some idea of the possible amplification per stage and the factors that affect this amplification. If the inductance L is equal to 200 microhenrys, the maximum capacity of the tuning condensers will be 419 mmf. or slightly more. The resistance R of the circuit may be of the order of 10 ohms at 550 kc. If impedance coupling is used, the load impedance at 550 kc. will be

$$R_L = \frac{L}{RC} = \frac{200\,.\,10^{-6}}{10\,.\,419\,.\,10^{-12}} = 47700 \text{ ohms,}$$

and at 1500 kc. the load impedance, assuming that the resistance R has increased from 10 to 30 ohms, will be

$$R_L = \frac{L}{RC} = \frac{200\,.\,10^{-6}}{30\,.\,56\,.\,10^{-12}} = 119000 \text{ ohms.}$$

The screen-grid tube has a mutual conductance of approximately 1000 micromhos. The amplification per stage is then equal to

$$G = G^m R_L = 1000.10^{-6}\,.\,47{,}700 = 47.7$$

at 550 kc.

The amplification available is found to be directly proportional to the tuning coil inductance and inversely proportional to the circuit resistance. The latter consideraton calls for coils of good shape factor, shielding that does not appreciably intercept the magnetic lines of force of the coil, large size copper wire, low-loss materials in the constructon of coil forms and compensating condenser dielectrics and supports, and good condensers. The value of L should be made as large as practical not only because of the increase in amplification obtainable, but also because the selectivity is improved thereby as is indicated by the equation for the decrement of the tuned circuit, given by

$$\delta = \frac{R}{2\,f\,L}$$

R-F. Choke

When the stages are impedance coupled an r-f. choke coil must be used, to avoid an excessive use of blocking condensers in the tuned circuit, and a grid resistor. The use of such a choke has several advantages. The gain characteristic of the amplifier can be varied considerably by choosing the characteristics of the choke coil. The choke coil impedance must, however, be extremely high at all frequencies to prevent it from adding considerable reactance to the tuned circuit. The characteristics of r-f. choke coils vary considerably with the conditions under which they are used, as for example, the circuit to which they are connected, their proximity to the metal chassis of the receiver, and the iron plate on which this chassis rests in the console and the character of the bolt used to hold the choke coil to the chassis.

Because of the characteristics of the choke coil and because of the effectively closer coupling obtained with tuned impedance circuits, the amplification characteristic in the r-f. stages is in general more uniform than is obtained with transformer coupling. Comparative results for the two types of coupling are shown in Fig. 5. If a different design of r-f. choke coil had been used, a different curve for impedance coupling would have been obtained.

The same latitude in design is permissable with transformer coupling but there are greater practical limitations. In order to obtain a reasonably low ratio of amplification at short waves to that obtained at long waves, the primary inductance must be very large. The effect of increasing primary turns is shown in Fig 6. One reason for the decrease in amplification at the short wavelengths when the primary inductance is increased above a certain value can be understood by considering the equation for transformer coupling. The load impedance we have found equal to:

$$R_L = \frac{\omega^2 M^2}{R_2}$$

The ratio of the voltage across the primary to that impressed on the grid of the preceding tube is then

$$\frac{E_1}{E_g} = G_m R_L = G_m \frac{\omega^2 M^2}{R_2}$$

If the coefficient of coupling between primary and secondary is nearly unity the voltage step-up or step-down between primary and secondary is substantially equal to $\frac{L}{M}$ and the voltage ratio between the grids of successive tubes is then

$$g = G_m \frac{\omega^2 M^2}{R_2}\,\frac{L}{M}$$

As ωM is increased, the value of the term R_L increases and may reach a value where it is no longer possible to express the equation for amplification without including the term $\frac{r_P}{R_L + r_P}$ which may become considerably less than unity in value. Then the gain will have to be expressed as follows:

$$g = G_m \frac{\omega^2 M^2}{R_2}\,\frac{r_P}{r_P + \frac{\omega^2 M^2}{R_2}}\,\frac{L}{M}$$

$$= G_m \frac{r_P}{\frac{r_P R_2}{\omega^2 M^2} + 1}\,\frac{L}{M}$$

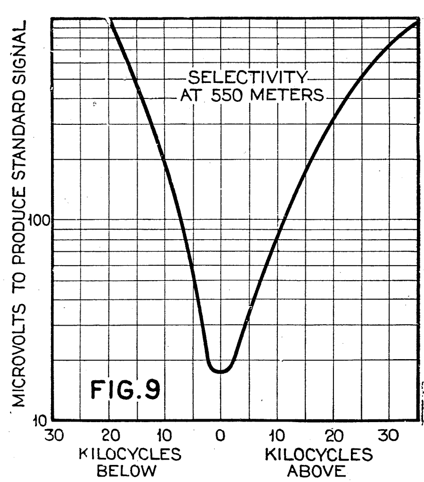
$$= \mu \frac{\omega^2 M^2}{r_P R_2 + \omega^2 M^2}\,\frac{\omega L}{\omega M}$$

$$= \mu \frac{\omega M}{r_P R_2 + \omega^2 M^2}\,\omega L$$

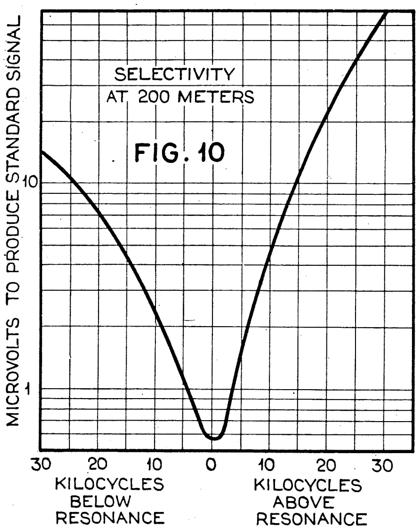
With respect to variations in ωM this value of g is a maximum when:

$$\omega^2 M^2 = r_P R_2$$

When $\omega^2 M^2$ is increased in value by varying M to the point where it becomes equal to or greater than $r_p R_2$ there is a decrease in amplification. The decrease in amplification at 200 meters with increasing primary turns as shown by the curves of Fig. 6 may not be explained in this way, however, for it is doubtful whether the conditions for optimum coupling have been satisfied in these tests. These results are more likely due to capacity coupling between primary and secondary which reduces the load impedance of the primary circuit when the num-



Microvolts required in a .00025 mf. antenna to produce a standard signal of 50 milliwatts.



Microvolts required in a .00025 mf. antenna to produce a standard signal of 50 milliwatts.

ber of primary turns exceeds that of the secondary.

Effect of Capacity

The effect of capacity between the primary and secondary windings of an r-f. transformer designed for screen-grid tubes is more important than similar capacities have been considered in the past. This capacity has its effect on the amplification per stage and on the frequency range that a given tuning condenser will cover. The latter effect is probably the more important, and is especially noticeable if the primary winding is wound close to the secondary and covering only a portion of the secondary winding. In general, coils with an equal number of turns for both primary and secondary have been found most satisfactory. Especial care must be taken that the stages will tune to the same frequency over the entire broadcast range. Because of the very close coupling required, and the effect of the copper in the primary winding on the distributed capacity of the secondary, these coils may not tune to the same frequency over the broadcast band with the antenna coil or other coils in the receiver that differ from them in construction.

Selectivity requirements may not, in general, permit the use of unity ratio in the design of the radio-frequency transformers. The resistance added to the tuned circuit and due to the tube plate impedance will be equal to

$$R' = \frac{\omega^2 M^2}{r_p}$$

This may amount to nine or ten ohms at 200 meters. If impedance coupling is used, the same value of resistance is added to the tuned circuit if the full number of turns on the tuning coil is used in the plate circuit of the amplifier tube.

Methods of Measurement

Two distinct methods of measurement have been used in obtaining the experimental data that was necessary for the verification of the theory. Most measurements on radio receivers and circuits require the use of a very small known input voltage. If amplification measurements on a single stage of radio frequency are to be obtained this voltage can be as large as .1 volt and its value must be accurately known. If several stages of radio-frequency amplification or an entire receiver are to be measured, this voltage may have to be less than a millivolt and if a very sensitive receiver is to be measured it may have to be as low as a microvolt, and again the voltage must be accurately known.

There are a number of well-known methods of obtaining a small known voltage of radio frequency. One of these is to pass a current of the desired frequency through a straight copper rod of negligible resistance and to use the reactive voltage drop of a portion of this rod to supply the input voltage. The inductance of the rod can be calculated by empirical formulas.

A second method is to use the resistance drop of a small non-inductive resistance. It is, however, more difficult to obtain a small accurately known resistance of negligible inductance than it is to obtain a small accurately known inductance of negligible resistance.

A third method is to use a mutual inductance. Empirical formulas for mutual inductance are not, however, as accurately known as for self-inductance, and measured mutual inductances would not be of a sufficiently small value to render them satisfactory for this purpose.

A fourth method is to generate a large voltage of the desired frequency, measure it with a thermocouple meter or vacuum tube meter, and then attenuate it with a radio-frequency attenuator of known characteristics to the desired value.

The first and fourth of these methods have been used to obtain the curves described in this paper. The first method was used exclusively until another method giving smaller voltages was necessary. Then the results obtained by the first method were checked by the fourth and they were found to be reasonably accurate.

The first method is valuable because it can be used conveniently without the need of elaborate apparatus. It is accurate, provided that considerable care has been taken in determining that the calculated impedance and the actual impedance across which the voltage drop is used are identical. For the purpose of these measurements a copper rod .15 cm. in diameter and 25 cm. long was used. The inductance of this rod is given by the formula

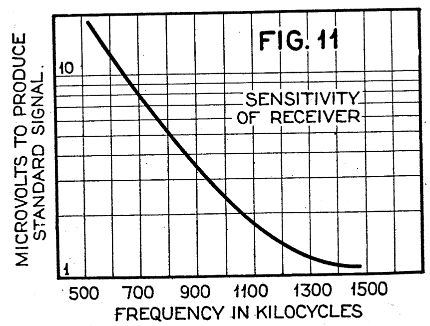
$$L = .002 \left[2.303 \log_{10} \frac{4l}{d} - .75\right]$$

where l is the length of the rod in cm., d is the diameter of the rod in cm., and L is the inductance in microhenrys. The inductance of the rod described is then equal to .287 microhenrys, and the reactance is equal to .991 ohm at 550 kc. and 2.703 ohms at 1500 kc. If two milliamperes of current are permitted to pass through the rod at either frequency, the voltage developed between the ends of the rod will be 1982 microvolts at 550 kc. and 5406 microvolts at 1500 kc.

The mechanical and electrical arrangement must be such that the thermocouple meter cannot read current that does not actually pass through the rod. In general, the meter should be connected to the circuit as near ground as possible and the circuit should be grounded at one point only and all capacities to ground for the rest of the circuit should be as low as possible. Sources of error that might be encountered are: a, mutual inductance between the portion of the rod used for supplying the reactive voltage and the ends of the rod whether the end lengths are in the same straight line as the useful portion or not; b, mutual inductance between the useful portion of the rod and the ground wires and metal chassis; c, charging currents in the circuit.

If the measuring equipment is well arranged mechanically and electrically, these errors can be made very small. The errors due to mutual inductances can be eliminated by using instead of a copper rod, a copper rod and sheath concentrically arranged, the sheath acting as a return circuit. The sheath will not eliminate mutual impedances but it will make them more definite and susceptible to calculation. Current flowing in the copper sheath does not produce a magnetic field inside the sheath, yet a mutual inductnace between these two branches of the circuit does exist due to the electrostatic fields existing between the electrons comprising the current. The mutual impedance can be calculated, whereupon the resultant impedance of the rod alone becomes known and can be used for supplying the reactive voltage drop.

Thus, having a source of known voltage, we need only an output voltage measuring device to measure the voltage after amplification. A vacuum tube voltmeter is most satisfactory for this purpose because of its slight effect on the apparatus to be measured.



Microvolts required in a .00025 mf. antenna to produce a standard signal of 50 milliwatts.

Amplification is then given as the quotient of the measured output voltage and the known input voltage. Selectivity measurements can be made in the same way, varying either the input frequency or the reactance of the circuit under test.

The curves shown in Figs. 5, 6, 7 and 8, were obtained by the use of the method just described. These results were checked at a later date using a standard signal generator supplying an accurately known voltage across a very small resistance. The latter method is the fourth one mentioned above. For single stage amplification measurements this voltage was applied directly to the grid of the amplifier tube. When making measurements on a complete receiver and

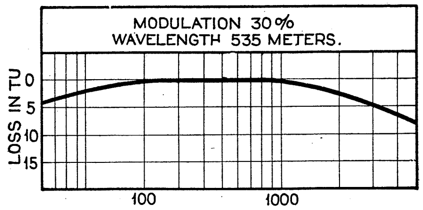


FIG. 12

Fidelity characteristic from antenna to loudspeaker.

when making measurements on the transmission characteristics of band-pass circuits, this voltage was applied to a dummy antenna circuit consisting of the 2-ohm resistance of the signal generator, a .00025 mf. condenser and the primary of the transformer used for coupling to the apparatus being measured. In addition to checking the results obtained by the first method, this method was used to obtain the results of Figs. 2, 9, 10, 11 and 12. The last four figures describe the performance of a laboratory receiver designed according to the principles discussed in this paper. The selectivity curves are of special interest. They illustrate what can be done in the way of maintaining uniform selectivity characteristics in a receiver. Though the selectivity curve at 200 meters is slightly broader than at 545 meters, it is believed that there is an improvement in results over that which is usually achieved with four tuned circuits in the space available for the radio-frequency circuits.

Book Review

ELEMENTS OF RADIO COMMUNICATION, by John H. Morecroft, x + 269 pages, John Wiley & Sons, 1929, $3.00.

RADIO TELEGRAPHY AND TELEPHONY, by Rudolph L. Duncan and Charles E. Drew, x + 950 pages, John Wiley & Sons, 1929, $7.50.

Here are two new volumes, both broadly concerned with radio technology, both issued by the firm of John Wiley & Sons, and both presenting a valid claim for admission to the progressive radio engineer's library, yet differing considerably in their contents and mode of presentation. The difference in contents would follow naturally, of course, from the fact that the two books are issued by the same publisher at about the same time. Another point of similarity is that all the authors are well known as radio educators, although in different fields. Morecroft is Professor of Electrical Engineering at Columbia University, and author of the classic *Principles of Radio Communication*. Duncan is Director of the Radio Institute of America, in which Drew occupies the position of Instructor in Radio.

The Duncan-Drew work is almost as long as Morecroft's *Principles of Radio Communication*. The difference between the two books may be inferred from the statement that a professional radio engineer, if he could secure only one of them, would unhesitatingly choose the *Principles*, while a radio serviceman or operator would be likely to pick the work of Duncan and Drew. This is consistent with the object of the authors, and the fields in which they have done their work. Similarly, a student radio engineer would naturally take to Morecroft's *Elements*, although he might also get valuable material out of *Radio Telegraphy and Telephony*. He would find in all of Morecroft's writings a firm, highly evolved theoretical grasp of the type which forms the only reliable basis for practical results in engineering, and less concern with the details of contemporary equipment. The tendency of Duncan and Drew is to present the elements of the subject in much detail and to break down all subjects into easily comprehended fundamentals, and then to leap directly to lengthy apparatus description.

Radio Telegraphy and Telephony starts off with a brief introductory chapter which is rather badly arranged. The following chapters are concerned with the elements of magnetism and electricity, motor-generators, meters, storage batteries, etc. Chapter IX, a comprehensive review of the elements of alternating-current theory, is followed by a treatment of "Condensers—Electrostatic Capacity—Capacity Measurements," preparatory to a 100-page chapter on "Vacuum Tubes" and a 126-page chapter on "Receiving Circuits." Considerable text is devoted after this to alternating-current receivers and tubes, and receiving accessories, especially loud speakers. Various commercial types, from the venerable 106-D to modern tube receivers in the communication field, are described at length.

With Chapter XVII the discussion turns to transmitting equipment. High voltage condensers, antennas, the phenomena of resonance, transmitter adjustment, and the characteristics of commercial broadcast and telegraph tube transmitters are considered in turn. Spark transmission is relegated to a place behind short-wave transmission and reception. The arc transmitter and the radio compass have later chapters of their own. The last chapter, XXVI, is concerned with "Radio Telephone Broadcast Transmitter Equipment," although much material on this topic is included previously in Chapter XX ("Commercial Broadcast and Telegraph Transmitters"). An appendix and index complete the text.

The somewhat confusing arrangement of broadcast transmitter material mentioned above is a characteristic fault of Messrs. Duncan and Drew's otherwise meritorious effort. The descriptions are in places badly arranged and give an appearance of imperfect digestion of the material. As a specific instance, the carbon microphone, including the broadcast type, is discussed on pages 610–614 of Chapter XX, while the condenser transmitter is described on pages 719–720 of Chapter XXI and again on page 894 of Chapter XXVI.

The fault of illogical arrangement is not found in the Morecroft text. At times the terminology is open to criticism, as when the author refers on page 10, to "distorted waves," when he means complex or non-sinusoidal waves. Otherwise, the book sustains throughout the impression of mature reflection on the author's part. The first three chapters present the underlying laws governing the behavior of audio and radio-frequency circuits and the principles of radiation. "The Vacuum Tube and Its Uses" is the title of Chapter IV. Then follow chapters on radio telegraphy and radio telephony, and a final chapter, VII, on "Receiving Sets." Pages 257–266 contain, in small type, problems arranged by chapters. The index is brief.

Morecroft's *Elements of Radio Communication* contains no plethora of material, but at every turn Morecroft's wide physical knowledge is exhibited, to the profit of the student and even the experienced engineer. Such points as the calculation of the capacity of the earth on page 35, the fine range of comparative data in the discussion of "What Is a Good Vacuum?" on page 104, and the illustration, beginning on page 248, of how ordinary alternating-current equations may be used to solve simple filter problems, are examples of this invaluable trait.

Carl Dreher.