A NUMBER of problems arise in the design of f-m receivers as compared with those for a-m only. Because the f-m system is capable of handling greater dynamic range, the power output of the audio amplifier must usually be about twice that for a-m, and the extended frequency range requires up to 15,000 cycles frequency response.

Distortion must be held to less than 5 percent and efforts must be made to eliminate high-order distortion. A de-emphasis circuit must be switched in for f-m to compensate for the predistortion introduced at the transmitter to improve the high-frequency signal-to-noise ratio. The de-emphasis element takes the form of an R-C low-pass filter with a 75-microsecond time constant for the requisite 6-db-per-octave audio attenuation.

Audio and modulation hum may arise in the filament circuit. Heater-to-cathode leakage in detectors employing balanced discriminators or ratio detectors may cause hum problems in production because one of the cathodes is above ground for audio frequencies. Insofar as a-c/d-c receivers are concerned, the detector must be placed at the low end of the filament string.

Filament-to-grid emission in the converter and local oscillator tubes of a-c/d-c receivers has been found to be a possible cause of frequency-modulation hum in the local oscillator. The use of a low value of grid leak in the order of 15,000 ohms has been found quite helpful. The converter tube should be located next to the second detector in the filament string in order to minimize the a-c filament-to-grid potential. The presence of frequency modulation in the local oscillator is readily detected by applying to the converter grid first an unmodulated i-f signal and then an unmodulated r-f signal and noting the increase in audible hum.

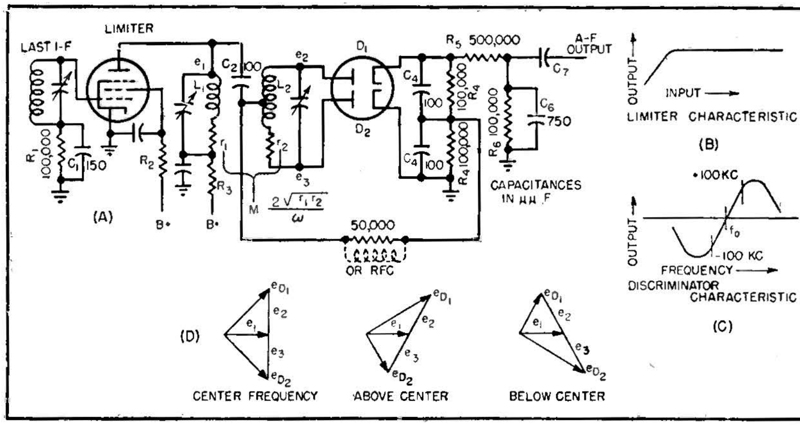

FIG. 1—Limiter-discriminator f-m detector circuit. The graphs show operational characteristics; the vector diagrams explain instantaneous effects

F-M Receiver

A survey of design and production techniques, including an evaluation of limiter-discriminator, ratio, and synchronized-oscillator detectors. Hum reduction and the tracing and elimination of regenerative effects in i-f and r-f stages, particularly for a-c/d-c receivers, are described

By E. C. FREELAND

Engineering Department
Philco Corp.
Philadelphia, Pa.

Detector Systems

There are three types of f-m detectors generally used in commercial receivers. They are: the balanced discriminator which is usually preceded by a limiter; the ratio detector which uses a balanced discriminator in a circuit arrangement which accomplishes noise reduction without the use of a limiter; and the synchronized-oscillator type of frequency detector, the commercial form of which is known as the Bradley detector, giving noise reduction without a limiter.

Figure 1A shows a schematic for the limiter and balanced discriminator form of detector. Amplitude-modulation limiter action is obtained by the use of a low time constant circuit of the order of 15 microseconds in the grid of the limiter tube and by the proper adjustment of plate and screen voltages to obtain a flat saturation characteristic, as shown in Fig. 1B. The opposing noise voltage appears across the low time constant circuit R_1C_1, with the result that the a-m noise modulation is materially reduced. Resistors R_2 and R_3 are chosen to give screen and plate voltages in the order of 5 to 40 volts to obtain the desired limiter characteristic.

The desired discriminator characteristic shown in Fig. 1C is obtainable with approximate transformer constants such that L_1 equals L_2, M is twice that for critical coupling, and Q is 50 for an i-f frequency of 10 mc. The magnitude of L_1 and L_2 determines the voltage output and is about 5 to 10 mh in commercial design.

The vector diagrams in Fig. 1D show how the voltage output is developed as the signal goes through its frequency modulation cycle. Voltages e_2 and e_3 are added at 90 degrees to the primary voltage e_1 through C_2 when at the center frequency. As we move off resonance, the phase of the secondary voltage shifts with respect to that of the primary, a difference in diode voltage is obtained and an incremental

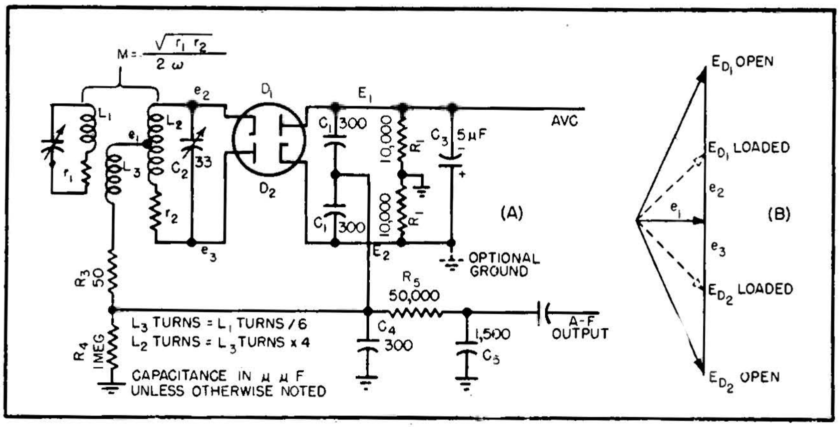

FIG. 2—Ratio detector for f-m. Transformer turns ratios are given and the vector diagram indicates optimum transformer performance

Design Problems

d-c voltage is produced across resistors R_4. Circuit R_6C_6 in conjunction with R_5 serves as an attenuator and de-emphasis network. The values given are approximately as required for the average audio amplifier of a radio receiver.

Ratio Detector

Figure 2A shows the schematic diagram for the ratio detector. Here, a-m rejection is obtained by a double diode circuit in conjunction with a balanced discriminator transformer which has special electrical constants. The addition of a capacitor C_3 for storage of stabilizing voltage and the reversal of the diode D_1 are the most significant circuit changes over that of the ordinary balanced discriminator. In addition, the diode conductance should be high and the diode load resistors low so as to load the secondary windings L_2 to a point where the secondary Q is approximately one-fourth the unloaded Q.

The open circuit voltage across L_2 is made large compared to that across L_3. The application of the diode load should then reduce the voltage across L_2 such that it is equal to about 75 percent of the voltage across L_3. This effect is illustrated in the vector diagram in Fig. 2B, which also gives the approximate turns relation required to simulate the conditions of this vector relation. Coil L_1 is made large in comparison to L_3 so as to match more nearly the plate resistance of the i-f amplifier and to minimize the effect of the diode load on the primary Q. The value of M is made less than critical so that E_2, within limits, will be a direct function of the diode load.

The application of a carrier signal to the primary of the discriminator transformer will now charge C_3 to a d-c voltage equal to $E_{D1} + E_{D2}$ and this charge will follow low rates of change in carrier level. However, suppose that the carrier level is suddenly increased as by a burst of noise. Capacitor C_3 will essentially act as a short-circuit, consequently the diodes will impose a heavy load on L_2 and the voltages E_{D1} and E_{D2} will increase only by a small fraction of the carrier increase caused by the noise. Conversely, if the carrier level is modulated downward by the noise, C_3 remains charged and reduces the diode current, which results in an increase in E_{D1} and E_{D2}. If the downward modulation drives the diode current to zero, then L_2 is open-circuited and no increase in opposition noise voltage is possible.

The lower the value of R_1 and the greater the Q of L_2 the more downward modulation the detector can handle. However, the lower the value of R_1 the less sensitive is the detector and the greater is the possibility of distortion due to incorrect phase relations in the transformer. A good compromise design using approximately the constants shown will handle up to 60 percent of downward modulation.

The audio voltage appears across R_4 which, so far as operation is concerned, may be an open circuit and is shown only as a means of clarifying the functioning of the circuit. When the carrier frequency is at its center value the voltages E_{D1} and E_{D2} are equal, and since the two diode currents are in opposition, no voltage will be developed across R_4.

Now as modulation moves the carrier off center frequency such that E_{D1} decreases and E_{D2} increases, the current through R_4 due to E_{D1} is decreased. This is equivalent to an increase being caused by a voltage in opposition to that due to E_{D1}. Since E_{D2} represents such an opposition voltage, the two changes in diode currents produce a resultant change in voltage across R_4 equivalent to connecting the two diode voltage increments in parallel, while a balanced discriminator adds these voltages. Thus the output of the ratio detector is given as one-half E_{D2} minus E_{D1}.

Since no audio voltage can appear across the stabilizing capacitor, the ground can be shifted to the optional location and thus permit the use of a cathode common to that of the first audio tube. However, this arrangement does not permit the effective use of balancing resistors between the diode and the junction of the stabilizing capacitor and load resistor which are sometimes required for the best results insofar as noise rejection is concerned. These balancing resistors are of the order of 1,000 ohms.

Equivalent Circuit of Ratio Detector

Figure 3 may help further to clarify the operation of the ratio type of detector. Figure 3A shows the equivalent circuit under center-frequency conditions, with batteries substituted for rectified and applied voltages. Resistors are substituted for the various impedances. The values shown are only relative for the purpose of illustration and bear no absolute relation to the actual detector circuit. Batteries B_1 and B_2 are fictitious voltages which replace the i-f plate current of the driver stage. The diode load has also been relocated so as to represent the load across the transformer and a one-megohm resistor occupies its conventional location so as to simulate as nearly as possible the actual detector.

The diagram shows the detector under a stabilized condition, wherein B_3 may be removed without changing the circuit conditions. Now if, with B_3 connected, B_1 plus B_2 is increased or decreased the change in current will flow through B_2 plus R_{D1} and R_{D2} and the change in E_{D1} and E_{D2} will be 1/100th of that which would occur if the stabilizing voltage B_3 were not present. It is seen that a-m has little effect on E_{D1} and E_{D2}.

Figure 3B shows the relations at off center frequency for a change in diode voltages of one volt, Fig. 3D shows voltages in parallel.

Figure 3C shows the equivalent of a circuit used with a grounded cathode. Since the ground has effectively moved from zero to plus four volts with respect to point *a* and the voltage from *a* to *d* is proportional to the carrier strength, point *a* delivers an avc voltage equal to half the charge on the stabilizing capacitor.

The importance of electrical balance of the secondary of the discriminator transformer cannot be overstressed. The transformer parameters, particularly the impedance of the secondary and its coupling to the primary, are also deserving of careful consideration for the best possible a-m rejection.

A signal generator capable of delivering a signal with simultaneous amplitude and frequency modulation will be of great help. The frequency modulation should be of the order of 100 cycles at $\pm$75-kc deviation and the a-m of the order of 1,000 cycles at 50-percent modulation so that it is possible, by the use of a high-pass filter, to measure the a-m component of the audio output voltage in the presence of the output due to frequency modulation. This filter should have sufficient attenuation to reduce the reading on the output meter, due to f-m, to a negligible value compared to that due to a-m.

This check for attenuation can be made by switching off the amplitude modulation and reading the output through the filter when frequency modulation is applied. A synchronized scope pattern of the discriminator characteristic will also show the presence of a-m by giving a wide line. The proper interpretation of this pattern will be of value in determining the f-m/a-m ratio. A ratio of 30 to 1 in voltage, which is about 30 db, is considered satisfactory for field performance. The ratio should be checked at various levels of input.

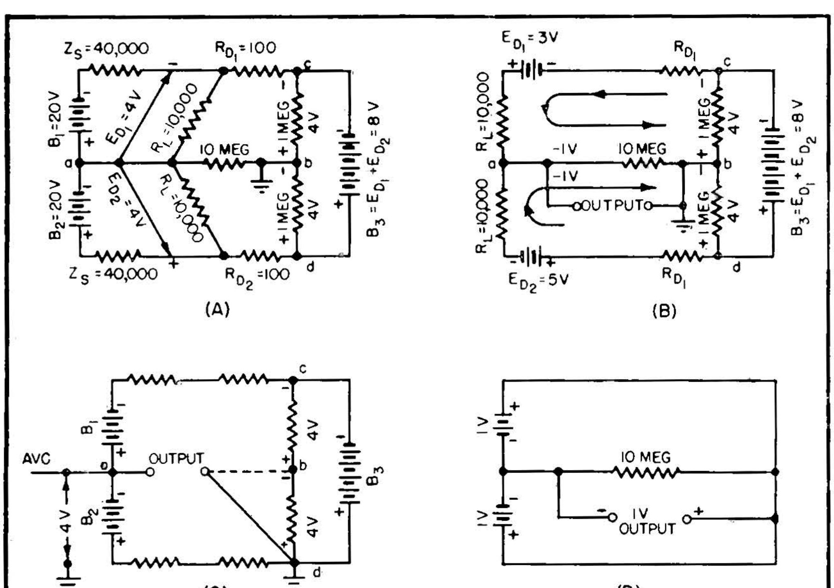

FIG. 3—Simplified diagrams of the ratio detector, with batteries representing voltages and resistances representing impedances

Figure 4 shows a circuit arrangement which is applicable to the Bradley synchronized-oscillator type of detector. The oscillator circuit may be of the Colpitts or Hartley type, the particular requirement being that it run under class C conditions. The tube is a pentagrid type with the element structure of such design as to give the special characteristics required for best operation as a synchronized-oscillator type of detector and noise rejection.

Frequency modulation as required to maintain lock-in is obtained from a 90-degree component of the oscillator signal injected across L_1 by L_2. The magnitude is controlled by the change in oscillator plate current I_p. The control of magnitude of the 90-degree component is accomplished by changing the total effective bias of E_{G3} plus E_{G1}. This 90-degree component will appear as a capacitance or an inductance across L_1C_1, dependent upon the polarity of L_2. Therefore the reversal of L_2 reverses the phase of the audio output in reference to the carrier modulation.

In most cases the capacitive polarity for L_2 gives the best results and is determined by observing the oscillator frequency when changing the bias E_{G3}. If the phasing is correct, the oscillator frequency will increase with more negative bias.

Load resistance R_L damps L_2 such that its Q is approximately 10. This damping prevents changes in phase of the 90-degree component during the application of frequency modulation to the oscillator under lock-in conditions. When the circuit parameters are properly adjusted, a straight line is obtained between the break-out points and the output is independent of the input.

Figure 4B illustrates the method by which the effective bias of $E_{G3} + E_{G1}$ is made to vary with modulation. This method produces an audio component of plate current through R_3C_3, the amplitude of which is a direct function of the carrier deviation. It is to be noted that the time constant of R_3C_3 is 75 microseconds, as required to supply the proper de-emphasis correction to the audio response curve.

When the signal voltage is at 90 degrees to the oscillator grid pulse

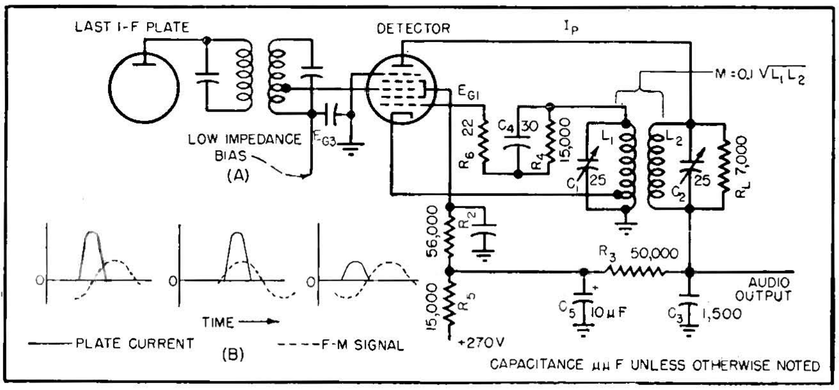

FIG. 4—Typical synchronized-oscillator f-m detector. Plate-current and signal relations are shown at B

no change in plate current will take place. It should be remembered, however, that a fictitious capacitance or inductance is present across L_1C_1 under this steady-state condition. The oscillator frequency is in part determined by this 90-degree component as well as L_1C_1 and the padding of C_1 must be such as to compensate properly.

Now let us apply modulation to the carrier such that its frequency is shifted by a small incremental change, and such that the phase shifts in such a direction as to approach the in-phase condition. A small increase in plate current takes place, resulting in a corresponding increase in the 90-degree component of oscillator current across L_1C_1. This change in turn produces enough change in oscillator frequency to satisfy the conditions around the loop.

The converse takes place when the modulation is such as to move the frequency in the opposite direction, as indicated by the out-of-phase condition shown in Fig. 4B. When the change in plate current caused by this phase shift is not sufficient to provide the magnitude of 90-degree component required to deviate the oscillator so that it is in step with the carrier, break-out occurs and the audio output goes through a point of discontinuity which will be observed as a ragged type of distortion. When break-out occurs, either the phase shift has gone through 90 degrees or the tube has reached saturation.

While the lock-in sensitivity of this type of detector is as low as 0.3 volt, full advantage of this sensitivity cannot be realized because of the fact that it is necessary to drive G_3 from a source impedance of a few thousand ohms in order to reduce stray oscillator voltage on G_3 due to capacitance coupling in the tube. Experience has shown that excessive oscillator voltage on G_3 introduces objectionable distortion. A more complete description appeared in the October 1946 issue of ELECTRONICS, p 88.

It is seen from Table I that the choice of the f-m detector has a direct bearing on the requirements of the i-f amplifier, particularly so far as gain is concerned. Whether or not an r-f stage is used also affects the gain requirements and overall stability of the i-f system.

The antenna sensitivity may vary between two and 75 microvolts, depending on the price class and performance requirements. It is possible to realize a gain of 2 in the antenna stage, and for the r-f stage a gain of 10. Although the theoretical maximum is considerably higher, it is difficult to realize, because of tube loading and other circuit losses that are difficult to control. Keeping these factors in mind, we can estimate the gain requirements and the number of stages to be used in the i-f system.

Figure 5 shows a typical double-channel i-f stage capable of handling either f-m or a-m signals. The a-m trimmers C_5, C_6, C_7 and C_8 act as bypasses for the 10.7-mc f-m signal, while L_1, L_2, L_3 and L_4 are of negligible impedance at the a-m i-f.

The stage gain at optimum coupling is given by

$$A = E_{G2}/E_{G1} = \frac{G_M\,(Z_3Z_4)^{1/2}}{2 + (Z_3/R_p)} \qquad (1)$$

when Z is ωLQ, L_3L_4 is in the order of 8 to 12 microhenrys and Q is about 50 for the average receiver. In most cases the term Z_3/R_p can be neglected. The attenuation at plus and minus 100 kc in a stage employing a transformer with a Q of 50 and adjusted for critical or optimum coupling is 1.2 to 1. It is desirable, from the standpoint of facilitating production padding and field operation in any one of the previously discussed detector systems, to maintain the coupling at slightly less than the critical value.

Regeneration

In many cases the chief problem pertaining to the i-f amplifier is that of regeneration. The cause of regeneration is difficult to locate because in many cases no analytical method seems to be available by which its source can be located.

It is helpful to consider a regenerative or degenerative amplifier as one having a portion of the output signal coupled back to the input in some particular phase relation to the original. Rather than feed the original signal in at the first stage of the amplifier, let it be applied to the last stage in such a manner that the regenerative as well as the original signal will be amplified. This effect is accomplished as shown in Fig. 5 by applying the so-called original signal from the generator through a small capacitor about 3 μμf to the terminal of L_3, C_3 being adjusted to compensate for

Table I—Approximate Performance Data for ±22-KC Deviation

Type of F-M Detector	Location of Measurement	Sensitivity in μV	Output Voltage
Ratio	Driver Grid	100,000	0.3
Bradley	Driver Grid	75,000*	2.0
Limiter and Balanced Discriminator	Limiter Grid	10×10^6	5.0

* Lock-in sensitivity for ± 75 kc

the additional capacitance across L_3. The front end of the amplifier has previously been tuned and the detector converted to an a-m type by opening one diode or stopping the oscillator. When the diode is removed, an equivalent capacitance should be substituted in its place to maintain correct tuning. With an amplitude-modulated signal, it is now possible for the audio amplifier to indicate relative signal amplitudes. The feedback signal, if present, is readily removed by shorting L_2 with a 0.01-μf capacitor. The presence of feedback will be indicated by a change in amplitude of the detected audio output.

When the amplifier circuits are tuned exactly to the frequency of the applied signal, the phase angle between the feedback signal voltage and that of the applied signal is usually some multiple of 90 degrees. If under these conditions the feedback is not in phase with the applied signal, the tuned circuits will seek a new frequency such that the several small phase shifts in each circuit will add up to bring the feedback voltage almost in phase with the applied signal. Under this condition the apparent maximum gain is not at the true resonant frequency of the tuned circuits and the selectivity curve becomes unsymmetrical, or if the feedback is of sufficient magnitude the amplifier will oscillate at some frequency, usually within $(1 \pm 2/Q)$ times the resonant frequency of the tuned circuits. If the feedback is small, and at 90 degrees to the applied signal, it may only change the symmetry of the selectivity curve, and it will then be necessary to check for the presence of feedback at frequencies slightly off resonance. The check is still made by observing the change in output due to shorting L_2 with a 0.01-μf capacitor.

The application of this method to the solution of a regenerative problem is relatively straightforward. Since there is no longer dependence upon the front end of the amplifier to provide a source of signal, it is possible to disconnect or short-circuit any point ahead of L_3 without affecting anything other than the regeneration.

To locate the source of feedback, start at the first stage and, with a 0.01-μf capacitor, bypass successively the grids and plates of each stage until a change in output signal level is noted. Such a point is a source of feedback, but not necessarily the only one.

If the amplifier is oscillating it will be necessary to proceed down the line until a point is reached where the bypass kills the oscillation. Having located the point at which regeneration occurs, it then becomes only a matter of inserting the necessary filtering, providing the feedback is in the low-potential end of the circuit. If it is in the high-potential end, other methods of correction must be applied.

The most familiar type of regeneration, which occurs in the high-potential end of i-f amplifier impedances, is that resulting from grid-to-plate capacitance. The advent of the screen grid tube eliminated, for a time, this type of regeneration, but as better i-f transformers were developed and the individual stage performance improved, it again became the limiting factor so far as stage gain is concerned. This style of feedback can be found in most low-cost broadcast receivers and manifests itself as an unsymmetrical selectivity curve.

An indication of the magnitude of feedback due to the grid-to-plate capacitance (C_{GP}) is given by

$$\frac{A_f}{A} = \frac{1}{1 - (C_{GP}\, Q_2\, A/C_2)} \qquad (2)$$

when A_f is gain with feedback owing to C_{GP}. It is approximate because it does not include the phase

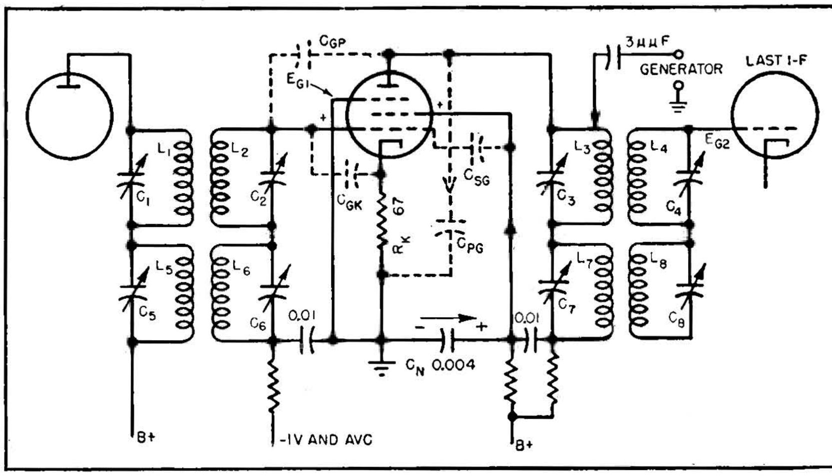

FIG. 5—Representative a-m and f-m intermediate-frequency stage with notations for discussion of regenerative effects

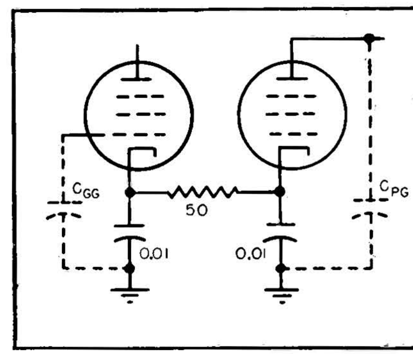

FIG. 6—B-minus decoupling circuit

angles which vary with the degree of feedback and approach 90 degrees as the magnitude of the feedback is reduced.

Consider a possible example where C_{GP} is 0.004 μμf, Q_2 is 50, C_2 is 24 and A is 60. By substitution it is found that the amplification with feedback present is twice that without feedback, while good design practice calls for this ratio to be less than about 1.3 to 1. If the overall design is such as to require the maximum possible i-f stage gain, the use of neutralization is a means of reducing the feedback due to grid-to-plate capacitance.

Neutralization

A convenient means of neutralizing the grid-to-plate capacitance involves obtaining an out-of-phase voltage on the screen with respect to that of the plate. The screen-to-control-grid capacitance C_{SG} then sets up a voltage across L_2 in opposition to that of the feedback voltage caused by the grid-to-plate capacitance. The polarity signs in Fig. 5 indicate the conditions of in-

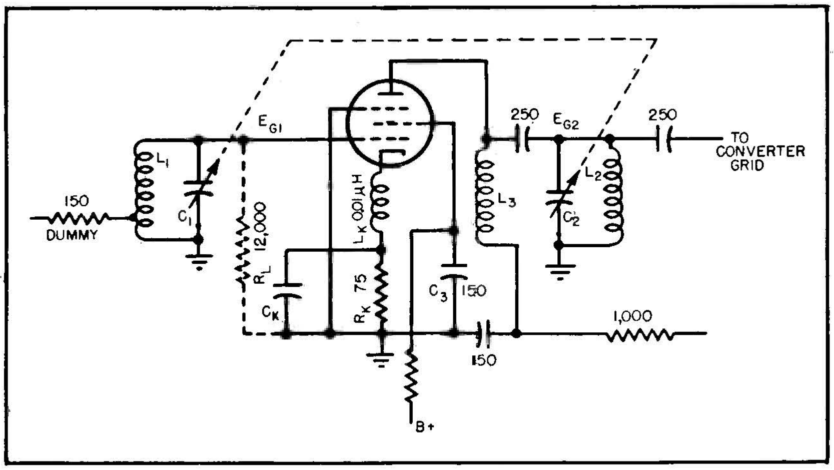

FIG. 7.—R-f stage for a-m and f-m receiver, arranged for discussion of regenerative effects

stantaneous polarity as required for conditions of neutralization.

The out-of-phase voltage to be applied to the screen is best obtained by making the screen bypass C_N common to the plate bypass. Now C_N, in combination with the plate-to-ground capacitance C_{PG}, forms a voltage divider across L_3 and the current is in such direction as to obtain the necessary phase reversal across C_N. For purposes of clarity, C_{SG} is assumed small in comparison to C_2 and hence shows only the approximate relation required for balance

$$C_N = C_{PG}\ C_{SG}/C_{GP} \qquad (3)$$

Since C_{PG} is not readily determined and the lead inductance of C_N is also a factor, the conditions for balance are more exactly determined by experimental methods than by calculation from the theoretical equation. Because the 3-μμf coupling capacitor adds to the plate-to-ground capacitance, a more accurate balance will be obtained if this capacitor is reduced to about 1 μμf and its leads dressed down towards the chassis to keep the grid-to-plate capacitance at a minimum. It will then be necessary to readjust C_3 for maximum output to compensate for the change in coupling capacitance. Now with L_1 shorted, observe the effect of C_2 on the output. If C_N is too large the circuit will be under-neutralized and the output will be observed to increase and then decrease as the phase of the feedback is changed by tuning C_2 through resonance from the high-frequency side. The converse is true if the circuit is in the underneutralized condition.

When the correct value of C_N is inserted no change in output will be observed as C_2 is tuned through resonance. The value of C_N may vary from minus 25 to plus 50 percent without increasing the feedback ratio beyond about 1.3 to 1 providing the ratio without neutralization does not exceed 2 to 1, but since allowance must be made for other variables, a tolerance on C_N of minus 10 to plus 25 percent is preferred.

Feedback in F-M Sets

The common type of overall regeneration due to common coupling in the B-plus circuits is familiar to most engineers and need not be discussed here. There is, however, a new problem in the B-minus circuits which will be encountered when designing an f-m receiver incorporating the familiar a-c/d-c circuit in which the chassis is isolated from the power circuits. Figure 6 shows the circuit elements involved in this type of feedback and a filter for eliminating it. The plate-to-ground capacitance C_{PG} sets up a voltage across the 0.01-μf capacitor connected between ground and cathode of this stage. In the absence of the 50-ohm filter resistor, this voltage would be applied between the grid and cathode of the previous stage through its grid-to-ground capacitance C_{GG}.

Experience would lead one to think that the value of the capacitor between cathode and ground in Fig. 6 should be increased, but a little further investigation will show that a 0.01-μf capacitor will resonate with half-inch leads at about 10 mc. The choice of this particular value of capacitor for most of the bypass requirements in the i-f amplifier of f-m receivers is then apparent. It is necessary to depart from conventional a-m technique so far as the B-minus circuit is concerned, by adding filter sections between the cathodes of successive amplifier stages.

Filament Feedback

Similar problems to that of the B-minus circuit also exist in the series filament string of a-c/d-c receivers, the feedback being due to the capacitance between the filament and cathode and the filament and grid. In the solution of this problem, series chokes are used and the bypasses are returned to the cathode rather than ground in order to avoid modulation hum due to a-c potential on the floating chassis. Filters of this type are also often required in a-c receivers to reduce feedback currents carried by the parallel filament circuits.

The combination of the low capacitance used in the tuned circuits of the i-f amplifier and the construction used in high G_m tubes increases to a marked degree the effect of the change in input capacitance when avc is applied to the grid. The proper choice of cathode resistor R_K will introduce an apparent negative capacitance designated as ΔC in the expression

$$R_K = \Delta C / C_{GK}\ G_M \quad (4)$$

This negative capacitance diminishes with G_m at approximately the same absolute rate as that of the positive input capacitance. The value shown in Fig. 5 is approximately that required for correction of input capacitance change of the average i-f stage employing a high G_m tube.

The choice of a converter is somewhat dependent upon overall design considerations. The triode is known to have a much lower equivalent noise resistance than a pentagrid type and therefore will be the most likely choice if no r-f stage is contemplated for one or both bands. If an r-f stage is included as part of the design, a pentagrid converter may give slightly better performance in f-m gain and r-f input resistance.

The chief regeneration problems associated with the converter are the so-called hot spots which are identified as highly regenerative portions of the band. They apparently arise from the oscillator coupling to the i-f through the filament or B-plus leads to produce an antenna frequency, which in turn is coupled back to the antenna or r-f circuits. Bypassing of the hot filament or B-plus leads to chassis through a small self-resonant capacitor of about 100 $\mu\mu$f has been found to be effective in removing this type of regeneration.

R-F Amplification

If an r-f stage is used, degenerative as well as regenerative problems will be encountered. Because the phase relations cannot be maintained so that the two effects cancel, it is necessary to treat each as an individual problem and apply independent solutions.

The acorn tube makes the most reliable type of diode voltmeter, but still imposes a loss across the tuned circuit. A small series capacitor will reduce the loss, but at the same time reduces the voltmeter sensitivity. Keep leads short.

The gain of the r-f stage shown in Fig. 7 is given by

$$A = G_M \, \omega \, L_2 \, Q_2 \qquad (5)$$

Special design precautions must be taken even to approach this theoretical value. An amplifier tube working at these frequencies has a low input resistance caused by transit time and the voltage across the cathode-lead inductance L_K. The grid-to-cathode capacitance C_{GK} couples this cathode voltage to the grid circuit in degenerative relation to the applied signal. If the cathode lead is assumed to be one inch long with a diameter of 0.05 inch, the relation

$$L_K = 0.005\, l \left[2.3 \log_{10}\left(\frac{4l}{d} + \frac{d}{4l} - 1\right)\right] \qquad (6)$$

shows it to have an inductance of 0.01 microhenry. Inserting this value in the equation

$$R_L = 1/\omega^2 \, G_M \, L_K \, C_{GK} \qquad (7)$$

and assuming that the grid-to-cathode capacitance of the tube is 5 $\mu\mu$f, while the G_m is assumed to be 4,000 micromhos, gives an apparent load resistance of 12,000 ohms. The shunt impedance of L_1C_1 is found to be about 10,000 ohms and hence R_L has the effect of reducing the Q of the tuned circuit by a factor of approximately two to one.

The effect of the input admittance on the circuit Q of L_1C_1 is readily verified by observing the voltage change across L_1 as the tube is turned on with the signal generator loosely coupled to L_1C_1 through a 2-$\mu\mu$f capacitor. If the loading is of the magnitude indicated by the previous calculations, the voltage will drop about two to one with the r-f tube hot as compared to that with the tube cold. Part of this loading depends upon the transit time of the electrons between the grid and cathode.

The net effect of the tube on the Q of the input circuit can be largely compensated by inserting, in series with the cathode, a small resistor R_K. If C_K is very small and R_K large in comparison to L_K, the phase angle of the feedback will be shifted such that R_L is replaced by a negative capacitance which affects only the value of C_1. If the impedance between cathode and ground has a capacitive phase angle, R_L will be negative, and if sufficiently so, the circuit will oscillate. When a proper adjustment of the cathode impedance is made, the generator voltage across L_1C_1 will not be affected by turning the heater of the r-f tube on or off.

The importance of short leads and low inductance in common circuit paths is important in r-f stages that employ the conventional type of f-m tuning capacitor. The circuit elements involved are arranged in Fig. 8, where L_{M1} is the mutual inductance to be considered as replacing the coupling in the capacitor owing to common currents through the rotor shaft and frame of the tuning capacitor. The inductance of a ground lead represented by L_{M2} indicates how coupling exists due to the stray ground capacitances of the grid and plate circuits. While this inductance, for the same degree of feedback, can be about five times that of L_{M1}, it will be shown that it is important to use the best possible grounding on the rotor shaft and frame of the tuning capacitor.

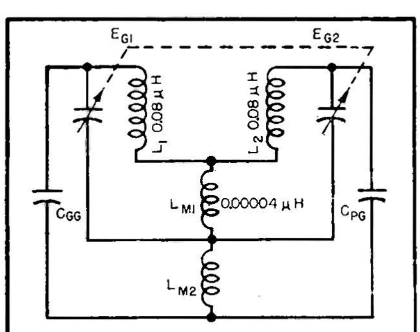

FIG. 8—Circuit to indicate capacitor-shaft coupling

Suppose that L_{M1} is physically represented by a copper rod 1/32 inch in diameter and 1/64 inch long. While Eq. 6 cannot be rigorously applied to such small dimensions, we find by substitution that the inductance of L_{M1} is approximately 0.00004 microhenry. Using the relation

$$A_f/A = 1/[1 - (L_M \, Q_1 \, A/L_2)] \qquad (8)$$

if the stage gain A is assumed to be 20 and Q_1 is 100, it is found that the right-hand denominator becomes zero and a condition of oscillation exists. This analysis, like that of the previous problem relative to i-f regeneration, is only approximate because it does not include the phase angle of the feedback which varies with the degree of feedback and approaches 90 degrees as the feedback approaches zero.

This type of regeneration might be analyzed by a similar procedure to that described for feedback in the i-f amplifier stage. The signal from the generator is applied across L_2 through a small capacitor and its amplitude observed on a high-frequency diode voltmeter connected directly across L_2. If regeneration is present the deflection of the diode voltmeter will change with the tuning of a trimmer across C_1 while C_2 is adjusted for resonance. Regeneration may be due to mutual coupling between the coils, capacitance coupling between the circuits, or mutual coupling owing to common currents in the tuning capacitor. Correction of these various conditions is straightforward. To eliminate the coupling in the tuning capacitor entirely, it may be necessary either to use insulated rotor sections in the gang capacitor or replace it with permeability tuned coils.