Limiters and Discriminators for F.M. Receivers

3 (cont'd)—Practical Ratio Detector Circuits: Comparison of Foster-Seeley and Ratio Detectors

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To extend the treatment to a practical ratio detector we shall next consider the case when the parallel resistance of the tuned secondary circuit R_s is not infinite.

As before, the fundamental-frequency currents (I_{ac}) flowing through the diodes are equal in magnitude, and the current in each diode is in phase with its applied voltage. Additionally, the current flowing through the resistive components $R_s/2$ are also in phase with the applied voltage. These currents are given by $2E_1/R_s$ and $2E_2/R_s$ respectively. The equations relating the magnitude of the currents flowing must thus be modified as follows:

$$(I/2)^2 = E_1^2Y_1^2 + (I_{ac} + 2E_1/R_s)^2$$
$$(I/2)^2 = E_2^2Y_2^2 + (I_{ac} + 2E_2/R_s)^2$$

where Y_1 and Y_2 have the same meaning as formerly, i.e. they are the admittances of the reactive elements of the tuned circuits. Inserting the values for Y_1 and Y_2 and writing $g_s = 1/R_s$

$$E_1^2\,[4C_s(\Delta\omega - \Delta\Omega)]^2 - E_2[4C_s(\Delta\omega + \Delta\Omega)]^2 + 4I_{ac}\,g_s(E_1 - E_2) + 4g_s^2(E_1^2 - E_2^2) = 0$$

As the diode rectification efficiency is assumed 100 per cent, $E_1 = E_b + E$ and $E_2 = E_b - E$, where E is the a.f. output voltage.

Combining the expressions above gives

$$\frac{E}{E_b} = \frac{-\Delta\omega}{\Delta\Omega} \times \frac{1 + (E/E_b)^2}{1+(g_s^2/4C_s^2\Delta\Omega^2)+I_{ac}g_s/(8C_s^2\Delta\Omega^2E_b)+(\Delta\omega/\Omega\Delta)^2}$$

In this expression, we can replace I_{ac} by $2I_{dc}$, and we can simplify the expression appreciably for initial examination by restricting consideration to the region where $\Delta\omega/\Delta\Omega$ is appreciably less than unity, i.e. to the working region near the centre frequency. The expression for E then becomes

$$E = -E_b\frac{\Delta\omega}{\Delta\Omega} \times \frac{1}{1+g_s^2/(4C_s^2\Delta\Omega^2)+g_sI_{dc}/(4C_s^2\Delta\Omega^2E_b)}$$

To a first degree of approximation therefore, the output is linearly proportional to the frequency of the input signal. However, it can be shown analytically that if the input current (I) increases, then so does the direct current in the load, I_{dc}. This is apparent also from an inspection of the circuit. Thus we can draw the important conclusion that the audio output voltage *decreases*, as the input current *increases*, and conversely. In this form the circuit is over-compensated. An examination of the expression above suggests a way out of this difficulty. If only part of the load voltage is "stabilized," i.e. shunted by a large capacitor, E_b will be no longer constant, and variations of numerator and denominator may be made to cancel. If a resistor R_m is inserted in series with each "battery" as shown in Fig. 10, R_L becomes $R_m + R'_L$, where R_L' is the resistance shunted by the large-value capacitor. E_b in the foregoing expressions must then be replaced by $E_b' + R_mI_{dc}$, where E_b' is the new battery voltage, i.e. that developed across R_L', no longer equal to the total load voltage. Inserting this value for E_b' in the expression above gives

$$E = -E_b'\frac{\Delta\omega}{\Delta\Omega} \times \frac{1 + R_m\,I_{dc}/E_b'}{(1 + g_s^2/4C_s^2\Delta\Omega^2)\;\;[1 + g_sI_{dc}/(g_s^2 + 4C_s^2\Delta\Omega^2)E_b]}$$

The factor $1/E_b$ is the denominator can also be replaced by $1/E_b'(1+R_mI_{dc}/E'_b)$, and provided the voltage R_mI_{dc} is small compared with E_b', this can then be replaced by the first two terms of the series expansion, viz.

$$\frac{1}{E_b'(1 + R_mI_{dc}/E_b')} = \frac{1}{E_b'}(1 - R_mI_{dc}/E_b')$$

Then

$$E = \frac{-E_b'(1 + R_mI_{dc}/E_b')\,(\Delta\omega/\Delta\Omega)}{\{1 + g_s^2/4C_s^2\Delta\Omega^2\}\left[1 + \dfrac{g_sI_{dc}(1 - R_mI_{dc}/E_b')}{E_b'\;(g_s^2 + 4C_s^2\Delta\Omega^2)}\right]}$$

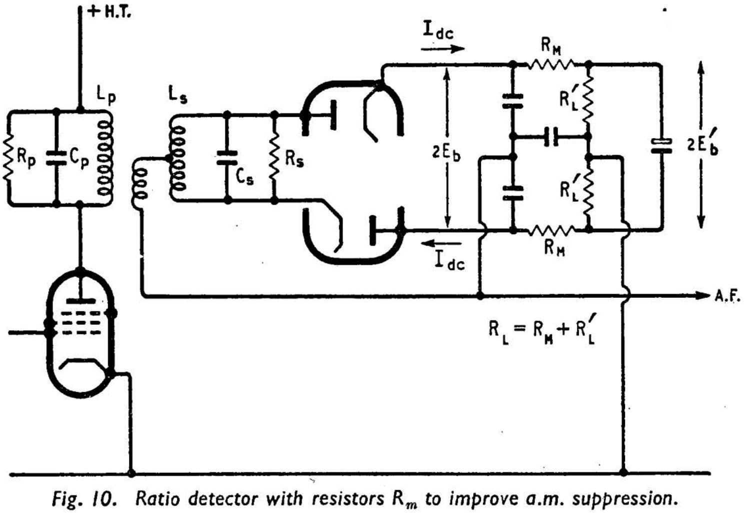

Fig. 10. Ratio detector with resistors R_m to improve a.m. suppression.

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The output E is independent of I_{dc} to a good degree of approximation if

$$R_m = g_s(1 - R_m I_{dc}/E_b') \,/\, (g_s^2 + 4C_s^2 \Delta\Omega^2)$$

$$\text{i.e. } R_m = \frac{g_s}{g_s^2 + 4C_s^2\Delta\Omega^2} \cdot \frac{1}{1 + g_s I_{dc}/\{E_b'(g_s^2 + 4C_s^2\Delta\Omega^2)\}}$$

$$= \frac{g_s}{g_s^2 + 4C_s^2\Delta\Omega^2 + g_s I_{dc}/E_b'}$$

This expression can be simplified by introducing the undamped Q value of the secondary circuit $Q_s = R_s\omega_o C_s = \omega_o C_s/g_s$. Then

$R_m = R_s/\{1 + (2Q_s\,\Delta F/f_o)^2 + R_s I_{dc}/E_b'\}$

This expression shows that complete a.m. rejection cannot be achieved, since the optimum value of R_m depends on I_{dc}, which varies during the a.m. cycle. We can make the output due to a.m. zero over a limited range about a selected value of I_{dc}. It is usual to do this about the working point, when E_b'/I_{dc} is equal to R_L', the resistance in parallel with the stabilizing capacitor. Then

$R_m = R_{m\,opt} = R_s/\{1 + (2Q_s\Delta F/f_o)^2 + R_s/R_L'\}$

The effect of varying R_m whilst the total diode load remains constant is shown in Fig. 11, which shows how the output varies with an a.m. input when the signal frequency is constant at a value near the centre frequency. The value of R_m is expressed in terms of $U = R_m/R_L$, where $R_L = R_L' + R_m$. The factor U is equal to the fraction of the output voltage which is not stabilized. The optimum value of U corresponding to $R_{m\,opt}$ calculated above is designated U_{opt}.

At this point it is convenient to introduce the "a.m. suppression ratio." This is a measure of the effectiveness of a f.m. detector in rejecting a.m. of the input signal. It is the ratio of the a.f. output due to f.m. to that due to a.m. when the input signal is modulated equally in depth by a.m. and f.m. The value of modulation depth employed is usually

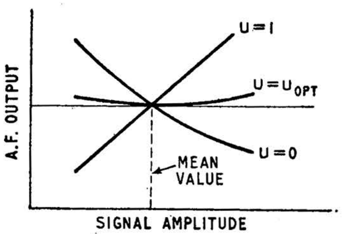

Fig. 11. Showing variation of a.f. output with input signal amplitude for a fixed frequency input signal (Δf small).

30 or 40 per cent. Different frequencies are usually employed for the a.m. and f.m. components to facilitate measurement. Typical values for these frequencies are 100 c/s for the f.m. component, and 2 kc/s for the a.m. component. In a practical ratio detector the a.m. suppression ratio is between 20 and 30 db.

If now we return to the full expression derived earlier for E the a.f. output voltage, we can evaluate the distortion terms in the output. Substituting for the value of E_b leads to the following expression for the a.f. output voltage E expressed as a fraction of E_b'

$$y = -Ax\,\frac{1 + B^2y^2}{1 + ABx^2}$$

where $y = E/E_b'$

$x = \Delta\omega/\Delta\Omega = \Delta f/\Delta F$

$A = (2Q_s\Delta F/f_o)^2/\{1 + (2Q_s\,\Delta F/f_o)^2\}$

$B = R_L'/(R_L' + R_m)$. This is the fraction of the direct voltage at the diode output "stabilized" by the electrolytic capacitor, and is equal to E_b'/E_b.

If the graph of y against x is plotted, it has the form shown in Fig. 12 for the portion of the curve in which we are interested.

The expression for y can be expanded as a power series, giving

$y = E/E_b' = d_1(\Delta f/\Delta F) + d_3(\Delta f/\Delta F)^3 \ldots\ldots$

where $d_1 = -A$

and $d_3 = A^2B\,(1 - AB)$

In order to use this expression to determine distortion terms, the value of ΔF must be known. Except for the special case of R_s infinite, this is not equal to the half-bandwidth (ΔF_p) measured to the turn-over point of the practical characteristic. However, in the process of deriving the expansion above, it emerges that the turn-over points of the characteristics occur at the values given by $x_p^2 = 1/AB$. As A and B are both less than unity, the measured half-bandwidth ΔF_p is greater than ΔF. In a practical circuit, if the measured half-bandwidth is found, ΔF can be found from

$$\Delta F = \Delta F_p\,\sqrt{AB}$$

The value of B can be calculated from the circuit constants. The value of A depends up ΔF, and hence requires a knowledge of the answer. However, ΔF can be found by successive approximations if the value of ΔF_p is used instead of ΔF to calculate A. This gives an approximate value of ΔF, which can be used to determine A more accurately, and so on. In fact, the error introduced by using the first approximation only is generally small.

Alternatively, the value of ΔF can be calculated from a knowledge of circuit values. Thus the resonance frequencies of the two tuned circuits are given by $1/2\pi\sqrt{(1 + M/2L_p)L_sC_s}$ and $1/2\pi\sqrt{(1 - M/2L_p)L_sC_s}$ respectively. From these expressions $\Delta F/f_o = M/4L_p$, whence

$$\Delta F = Mf_o/4L_p$$

If a tertiary winding or tapped primary circuit is employed, then L_p must be replaced by a^2L_p, and M by aM, giving

$$\Delta F = Mf_o/4aL_p$$

The expression for the output voltage is given in terms of E_b' and ΔF. If it is required to determine the sensitivity of the circuit, we require an expression relating E_b' to I_{in}. With a tertiary winding or tapped primary circuit, the equivalent circuit must be drawn with L_p replaced by a^2L_p, C_p by C_p/a^2, R_p by a^2R_p and I_{in} by I_{in}/a. We shall consider the signal frequency to be near the centre frequency. We can then ignore the effect of the reactive components of the tuned circuit connected between terminals 1 and 2 of the equivalent circuit. The dynamic resistance of this tuned circuit (R′) is that which in parallel with $R_s/4$ is equal to a^2R_p (see Part 1), i.e.

$$\frac{1}{R'} + \frac{4}{R_s} = \frac{1}{a^2R_p}$$

The impedance presented at the centre tap of the transformer T by the two tuned circuits and diode loads can be shown by an extension of the argument employed earlier to be

$$R_1 = \left(\frac{R_D}{4}\right) \Big/ \{1 + (2Q_w\,\Delta F/f_o)^2\}$$

where $R_D/2$ is equal to $R_L/2$ in parallel with $R_s/2$, i.e. the total damping applied to each of the two tuned circuits considered previously. The current flowing

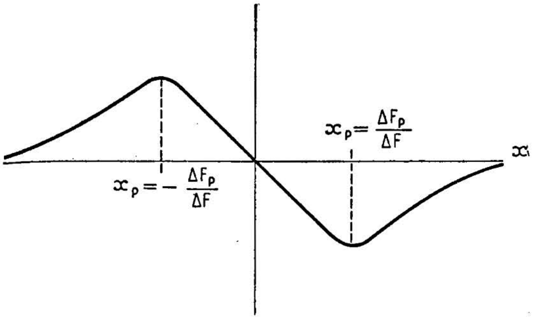

Fig. 12. Showing variations of output ($y = E/E_b'$) against input signal frequency ($x = \Delta f/\Delta F$).

to the centre tap of the transformer T is then given by

$$I = \frac{I_{in}}{a}\,\frac{R'}{R_1 + R'}$$

The power delivered to the centre tap of the transformer is thus

$$P_{in} = \tfrac{1}{2}I^2R_1$$
$$= \tfrac{1}{2}\left(\frac{I_{in}}{a}\,\frac{R'}{R_1 + R'}\right)^2 R_1$$

This power appears in the diode load circuit and in the dynamic resistances ($R_s/2$) of the two tuned circuits. Near the centre frequency, the voltage across each dynamic resistance is approximately E_b, and hence the power dissipated in each resistance is $\frac{1}{2}E_b^2/(R_s/2)$. The power in each diode load is E_b^2/R_L, and hence

$$\frac{2E_b^2}{R_L} + \frac{2E_b^2}{R_s} = \tfrac{1}{2}\,\frac{I_{in}^2}{a^2}\,\frac{R'^2}{(R_1 + R')^2}\,R_1$$

But $$\frac{2}{R_L} + \frac{2}{R_s} = \frac{2}{R_D}$$

$$E_b^2 = \tfrac{1}{4}\,\frac{I_{in}^2}{a^2}\,\frac{R'^2}{(R_1 + R')^2}\,R_1R_D$$

$$E_b = \tfrac{1}{2}\,\frac{I_{in}}{a}\,\frac{R'}{R_1 + R'}\,\sqrt{R_1R_D}$$

This expression has its maximum value when $a = a_{opt}$; this value of a is given by

$$1/a^2_{opt} = R_p\{1/R_1 - 4/R_s\}$$

Using this value for a, and $E_b' = E_bR_L'/R_L$, gives the maximum value for $E_b' = E_{b\,max}'$

$$E_{b\,max}' = \tfrac{1}{4}\,\frac{R_L'}{R_L}\,I_{in}\,\sqrt{R_pR_D/\{1 - (4R_1/R_s)\}}$$

If as is usual, $R_s/4$ appreciably greater than R_1 then

$$E_{b\,max}' \approx \tfrac{1}{4}\,\frac{R_L'}{R_L}\,I_{in}\sqrt{R_pR_D}$$

and $$a^2_{opt} \approx R_1/R_p$$

Near the centre frequency, the a.f. output voltage is given approximately by

$$E = -AE_b'\Delta f/\Delta F$$

For maximum sensitivity, E_b' should be large and ΔF small. For E_b' to be large R_p and R_D should be large. However, as we shall show next, the condition for good "downward" a.m. handling capacity requires R_D small. Thus a practical design represents a compromise between these requirements.

To complete the investigation of the circuit, we shall evaluate the maximum "downward" a.m. handling capacity. With no a.m. present, the relationship between the peak value of the r.f. input current to each of the tuned circuits ($I/2$), when the signal frequency is near the centre frequency, is given by

$$(I/2)^2 = E_b^2(16C_s^2\Delta\Omega^2 + 4/R_D^2)$$
$$= E_b^2 16C_s^2\Delta\Omega^2\{1 + 1/(2Q_w\Delta F/f_o)^2\}$$

When the diode current falls to zero, the value of the r.f. input current ($I'/2$) is given by

$$(I'/2)^2 = E_b'^2\ 16C_s\ \Delta\Omega^2\{1 + 1/(2Q_s\Delta F/f_o)^2\}$$

whence

$$m_{max} = 1 - I'/I = 1 - \frac{E_b'}{E_b}\sqrt{\frac{1+1/(2Q_s\,\Delta F/f_o)^2}{1+1/(2Q_w\Delta F/f_o)^2}}$$

$$= 1 - \frac{R_L'}{R_L}\sqrt{\frac{1 + 1/(2Q_s\,\Delta\,F/f_o)^2}{1 + 1/(2Q_w\Delta F/f_o)^2}}$$

If $2Q_s\,\Delta F/f_o$ is large, the expression simplifies to one similar to that given earlier for the case of R_s infinite.

The value of m_{max} calculated above ignores the effect of the primary circuit, and in general this is not negligible. To evaluate this, consider the input impedance presented at the centre tap of the transformer T of the equivalent circuit. With no a.m. present, this is

$$R_1 = (R_D/4)/\{1 + 2Q_w\ \Delta F/f_o)^2$$

Similarly, the impedance when the diode current falls to zero is

$$R_2 = (R_s/4)/\{1 + (2Q_s\Delta\,F/f_o)^2\}$$

If the values of R_1 and R_2 differ, then m_{max} is modified. This happens because the proportion of the input current (I_{in}/a) fed to the centre tap of the transformer T differs in the two cases. The effective impedance R' of the current source was shown earlier to be

$$1/R' = 1/a^2R_p - 4/R_s$$

Thus the proportion of the input current flowing to the centre tap when no a.m. is present is given by

$$I = \frac{I_{in}}{a}\,\frac{R'}{R_1 + R'}$$

and when the diode current falls to zero by

$$I' = \frac{I_{in}'}{a}\,\frac{R'}{R_2 + R'}$$

Thus m_{max} is changed from the value calculated above to

$$m'_{max} = 1 - \frac{I_{in}'}{I_{in}} = 1 - \frac{I'}{I}\,\frac{R' + R_2}{R' + R_1}$$

$$= 1 - \frac{R' + R_2}{R' + R_1}\,\frac{R_L'}{R_L}\sqrt{\frac{1+1/(2Q_s\Delta\,F/f_o)^2}{1+1/(2Q_w\Delta F/f_o)^2}}$$

Ratio Detector with Practical Diodes.—The analysis of the circuit operation with practical diodes, i.e. those with a rectification efficiency of less than 100 per cent is very complex, because the direct current component in the diode I_{dc} is no longer equal to half the peak value of the fundamental frequency a.c. component I_{ac}. If the diode efficiency is high, however, the assumption that $I_{dc} = \frac{1}{2}I_{ac}$ may still be made with fair accuracy. The diode itself must then be regarded as a perfect diode in series with a resistor R_{di}. This resistance then forms part of the resistance R_m calculated earlier, and represents a minimum value below which R_m cannot fall. To calculate its value, we may note that given an input peak voltage E the "perfect" diode delivers an output voltage E. A fraction of this voltage, ηE, appears across the true load resistance R_L, where η is the diode rectification efficiency.

The remainder of the voltage $(1-\eta)$ E, appears across the fictitious resistor R_{di}. Hence

$$\eta \frac{E}{R_L} = \frac{1-\eta}{R_{di}} \cdot E$$

$$R_{di} = \frac{1-\eta}{\eta} R_L$$

It is the fact that η varies with the input signal level which limits the useful range of input signal levels which the detector can handle satisfactorily.

Normally, η tends to a constant value as the input signal increases, and the circuit constants are adjusted for this value of η. At low input signal levels, however, η decreases appreciably, and the a.m. suppression ratio is seriously impaired. Thus there is a lower limit of input signal which the detector can handle satisfactorily. It is apparent that, for best performance high-efficiency diodes should be employed.

Unbalanced Effects.—In the presence of amplitude modulation, a ratio detector exhibits an "unbalance effect." This is an output due to the a.m. which is constant at all frequencies in the working range. This effect has a number of causes, which are not indicated by the preceding anaysis because of the simplifying assumptions made. The causes include variations of diode input capacitance with signal amplitude, the finite impedance of the tuned circuits to harmonics of the current flowing in the diodes, and inadequate decoupling of the diode loads at r.f. The last cause produces an effect which opposes that due to the first, and hence some reduction of the a.m. output may be obtained by using relatively small decoupling capacitors. The second cause can be minimized by using a large value of secondary circuit tuning capacitance. This tends to

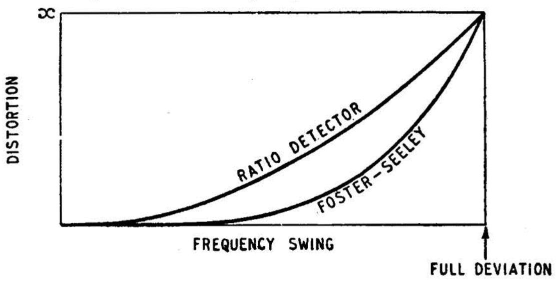

Fig. 13. Showing how distortion varies for ratio detector and Foster-Seeley circuits having equal distortion at full deviation.

produce low values of R_s, the dynamic resistance, and hence tends to lead to low sensitivity. A compromise is thus necessary, and a value of 50pF is usually employed. The a.m. output can, however, most easily be reduced by unbalancing the two diode-load circuit resistors, R_m. The foregoing analysis suggests that these should be equal; if, however, their sum is kept constant, while they are altered individually, substantial reduction of the a.m. output is then obtained.

Time Constant of Load Circuit.—The foregoing analysis was based upon the assumption that the time constant of the load circuit was very large, so that the load circuit could be replaced by a battery for the purposes of analysis. Unfortunately, if the time constant of the circuit is made very large, the tuning characteristics of the receiver are affected. The receiver then behaves like a conventional a.m. receiver in which the a.g.c. time constant is too long; there is an appreciable lag between adjustment of the tuning control and the return to stable operating conditions. The tuning has then to be adjusted very slowly. To avoid this effect, the load time-constant has to be shorter than is desirable. In practice a compromise value of time constant of the order of 0.1 to 0.2 seconds is usually employed. Such a circuit ceases to behave like a constant-voltage battery when the input is varying at a slow rate, and there is a slow variation of the output signal in accordance with the signal variation. This is especially noticeable if "flutter" due to signal reflections from an aircraft occurs. This flutter generally begins to be noticeable when the flutter rate is about 0.5 c/s; the flutter rate increases, as does also the amplitude of the "flutter," until the flutter rate rises to value when the load time-constant is sufficient to suppress the variations. To counter this effect, an effective fast-acting a.g.c. system is required. A suitable control voltage is available from the load circuit itself, and this is usually employed. The a.g.c. system also has the desirable effect of equalizing the audio output from input signals of unequal amplitude.

Variants of the Ratio Detector Circuit.—A number of variants of the ratio detector circuit have been described from time to time. The most common of these employ two tuned circuits instead of the phase-difference transformer. Two such circuits, shown in Figs. 14 and 15, are described by Paananen. In the circuit of Fig. 14, two tuned circuits are driven from two valves with the input grids connected in parallel, to supply equal currents to the tuned circuits. A battery is employed instead of the self-biasing circuit. In the circuit of Fig. 15, a low impedance source (a cathode follower) is used to drive two series-tuned circuits; the "battery" voltage is provided by the cathode bias of the cathode follower. This circuit may be described as the dual of that of Fig. 14, in that a constant voltage is fed to two series-tuned circuits instead of a constant current to two parallel-tuned circuits.

Comparison of the Foster-Seeley and Ratio Detector Circuits

A comparison of these two circuits depends critically upon the requirements of the detector in a receiver. These may be stated as (a) low distortion (b) good "downward" a.m. handling capacity (c) good a.m. suppression ratio (d) driving voltage required and (e) wide-band characteristics. On the score of low distortion, the Foster-Seeley circuit is better than a ratio detector of comparable bandwidth, although not necessarily better than a wide-band ratio detector.

The two circuits differ appreciably in the way in which the distortion varies with the signal frequency swing. In the Foster-Seeley circuit, the distortion at optimum adjustment increases with the fourth power of the swing; in the ratio detector it increases with the square of the swing. Thus if both circuits are adjusted to give equal amounts of third-harmonic distortion at full deviation, their characteristics at smaller frequency swings will be different. This is shown in Fig. 13.

In respect of "downward" a.m. handling capacity, the two are not strictly comparable, since in the Foster-Seeley circuit this maximum "downward"

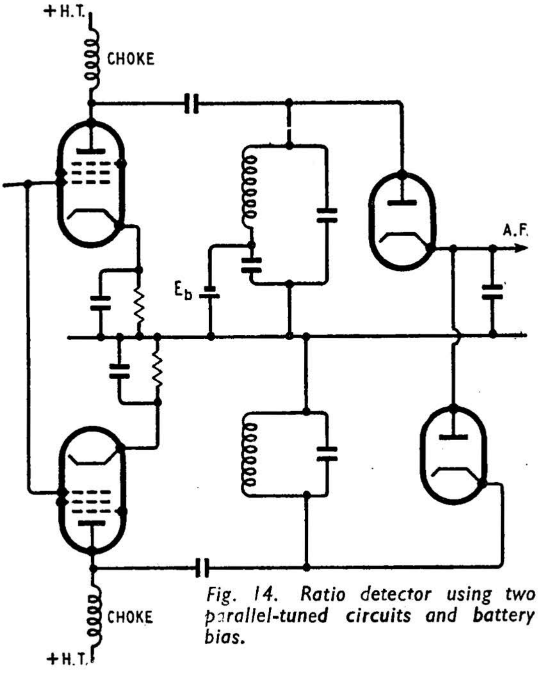

Fig. 14. Ratio detector using two parallel-tuned circuits and battery bias.

a.m. handling capacity is proportional to the amount by which the input signal exceeds the " threshold " input required by the limiter. In the ratio detector, the maximum " downward " a.m. handling capacity is a fixed quantity. In general a detector should be capable of handling " downward " modulation of the order of 50 per cent, and both circuits can normally achieve this.

The a.m. suppression ratio required in a receiver depends upon the type of interference encountered. To deal satisfactorily with all types of interference a ratio of 35–40 dB would appear to be necessary. The Foster-Seeley circuit preceded by a limiter has a ratio of the order of 40 dB. The ratio detector, in practice, appears to have a ratio of the order of 20–30 dB. This is not sufficient for all types of interference, and is perhaps the most serious limitation of the circuit. The ratio can be increased by employing a limiter preceding the detector, but if this is done a relatively large input signal to the stage is required. This offsets one of the major attractions of the ratio detector, the smaller number of valves required in a receiver. A hybrid arrangement, in which the preceding valve functions as a high-level limiter, may go some way to improving performance, but there may be difficulties in areas of low field strength. Alternatively, a diode dynamic limiter may be added to increase the a.m. suppression ratio, but this generally leads to some loss of overall gain.

The driving voltage quoted differs for the two cases. With a Foster-Seeley circuit, there is a " threshold " at which the limiter commences to function satisfactorily. Below this threshold the a.m. suppression ratio is poor, the " downward " a.m. handling capacity zero, and the a.f. output varies approximately linearly with the input signal amplitude. Above the threshold, the a.m. suppression ratio rises rapidly to an approximately constant value, and the a.f. output tends to a constant value. The " downward " a.m. handling capacity rises linearly with the ratio of input signal amplitude to threshold amplitude. In a ratio detector there is a " threshold " of a different type. This occurs when the input signal falls to the point where the diode efficiency begins to fall off. Below this threshold the a.m. suppression ratio and " downward " a.m. handling capacity decrease steadily. Above this threshold the a.m. suppression ratio and the " downward " a.m. handling capacity tend to constant values. The a.f. output, however, is proportional to input signal mean amplitude at all amplitudes except below the " threshold " where it falls somewhat more rapidly than the input signal mean amplitude.

The minimum input signal for satisfactory operation with a Foster-Seeley circuit is thus fairly well defined. If the circuit is to handle " downward " a.m. to a modulation depth of 50 per cent, this requires the input signal to the limiter to be approximately twice the threshold input. In a practical circuit, this corresponds to an input signal of some two volts at the limiter grid. The " threshold " input signal with a ratio detector is usually stated as the input voltage required at the grid of the i.f. stage feeding the detector, and this may be of the order of 20 mV. At this figure, however, the a.f. output may be appreciably below that of the Foster-Seeley circuit, and as a basis of comparison the ratio detector driver input, for an output comparable to that of a Foster-Seeley circuit, is perhaps better. A typical practical figure for the ratio detector on this basis of comparison is some 100 mV, the a.f. output being then approximately 1 volt peak. The overall gain from the aerial input to the driver/limiter grid is thus less by a factor of approximately 20 in a receiver employing a ratio detector than that in a receiver employing a Foster-Seeley circuit, and may enable an i.f. stage to be omitted. However, this saving may not be possible if the a.m. suppression ratio of the ratio detector has to be supplemented.

If a wide-band detector is required, the ratio detector would appear to be most satisfactory. It was stated in Part 2 that the Foster-Seeley circuit may suffer from " diagonal clipping," which occurs if the time-constants of the diode loads are so great that the envelopes of the signals applied to the diodes can fall faster than the rectified output falls. With increasing bandwidth the input signal envelope can fall progressively more rapidly and hence with a Foster-Seeley circuit, the diode loads must be reduced progressively with increased bandwidth. This leads to appreciable design difficulties. The ratio detector, however, can be made free from " diagonal clipping " as in the circuits of Figs. 14

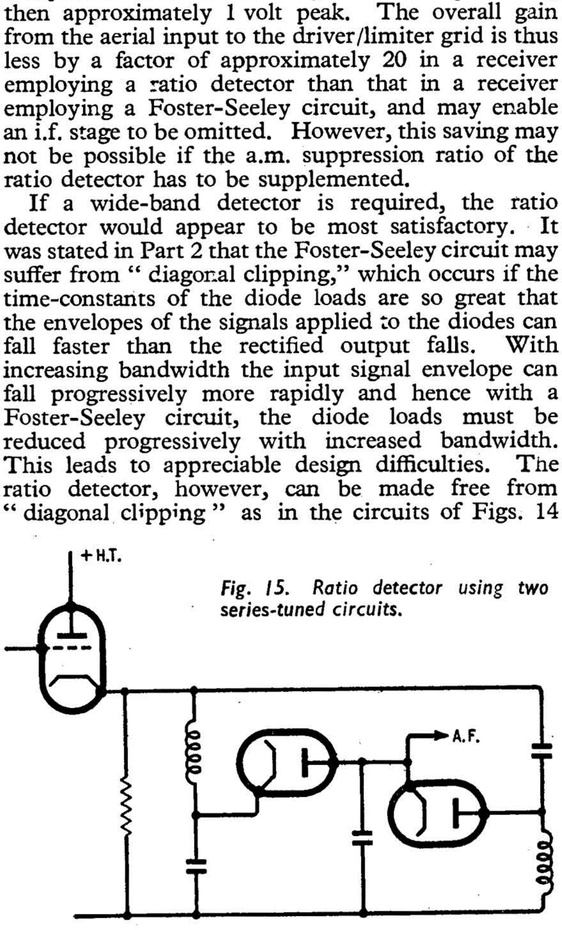

Fig. 15. Ratio detector using two series-tuned circuits.

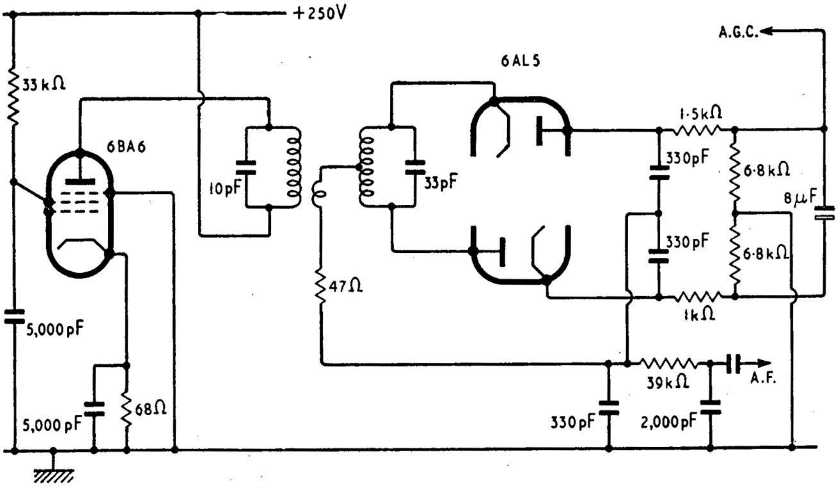

Fig. 16. Circuit values used in ratio detector described by Seeley and Avins (RCA Review, *June 1947*).

and 15. With increased bandwidth, the "downward" a.m. handling capacity decreases, and hence a wide-band ratio detector must be preceded by an efficient limiter.

Theoretical and Practical Results

To illustrate the order of accuracy of the preceding analysis, the practical results can be compared with those given for a published design. The design chosen is that due to Seeley and Avins, described in the *RCA Review*, June 1947. The circuit diagram is shown in Fig. 16.

The rectification efficiency of the diodes was estimated from published data to be 0.8 approximately. The value of R_{di}, the equivalent inherent resistance of each diode, is thus found to be 2 kΩ. This gives the mean value of R_m of the practical circuit as 3.25 kΩ. The calculated value is 3.2 kΩ.

The measured a.g.c. voltage is 2.5 volts for an r.f. input of 100 mV to the grid of the 6BA6. This is equal to E_b'. Assuming the 6BA6 to have a mutual conductance of 4.3 mA/V, the calculated value of E_b' is 2.3 volts. The measured value of the maximum "downward" a.m. capacity is 70 per cent; the calculated value is 70 per cent also.

To evaluate sensitivity and distortion, the values of A and B are required. From circuit values B = 0.68. The value of A is calculated to be 0.83, given the measured half-bandwidth (ΔF_p) of 175 kc/s. The value of ΔF is then 135 kc/s approximately.

Taking E_b' as 2.5 volts, the a.f. output voltage is given by $\quad E = 0.0154(\Delta f) + 2.7 \times 10^{-7}(\Delta f)^3 \ldots$

where Δf is measured in kc/s.

With an input signal of 75 kc/s deviation, i.e., $\Delta f = 75 \cos\omega t$

$$E = 1.06 \cos \omega t + 0.03 \cos 3\, \omega t \ldots$$

The calculated r.m.s. fundamental frequency a.f. output is 0.75 volt; the measured value is 0.7 volt. The calculated third harmonic distortion is 2.0 per cent (r.m.s.); the measured total harmonic distortion is 2.5 per cent (r.m.s.).

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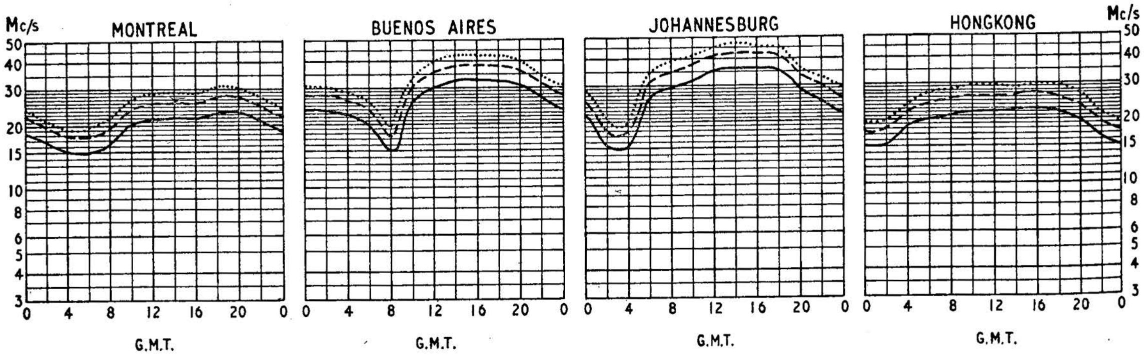
SHORT-WAVE CONDITIONS

Prediction for May



•••••••••• FREQUENCY BELOW WHICH COMMUNICATION SHOULD BE POSSIBLE FOR 25% OF THE TOTAL TIME

— — — PREDICTED AVERAGE MAXIMUM USABLE FREQUENCY

——— FREQUENCY BELOW WHICH COMMUNICATION SHOULD BE POSSIBLE ON ALL UNDISTURBED DAYS

THE full curves given here indicate the highest frequencies likely to be usable at any time of the day or night for reliable communications over four long-distance paths from this country during May.

Broken-line curves give the highest frequencies that will sustain a partial service throughout the same period.