Limiters and Discriminators for F.M. Receivers

By G. G. JOHNSTONE,* B.Sc.

4.—SOME LESSER-KNOWN DISCRIMINATOR CIRCUITS

THE discriminators most commonly employed in f.m. receivers, the Foster-Seeley circuit and the ratio detector, were discussed in earlier parts of this series. In the present article it is proposed to discuss some of the lesser-known types of discriminator. The chief of these is the gated-beam discriminator. Whilst it is not widely used in f.m. receivers, it is, however, used extensively in television receivers in the U.S. for the demodulation of the frequency-modulated sound carrier.

Gated-Beam Discriminator.—This class of discriminator relies for its action upon the phase relationships between the voltages developed across two loosely coupled circuits. The circuit is given

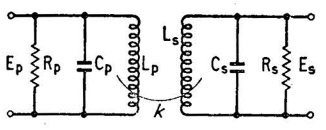


Fig. 1. Two coupled circuits; coupling coefficient k.

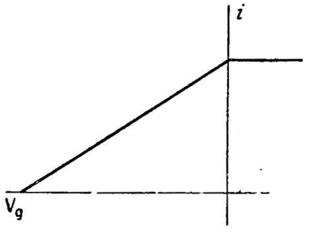


Fig. 2. Anode-current/grid-voltage curve of idealized valve.

in Fig. 1, and it was shown in the Appendix to Part 2 that the primary and secondary voltages are related by the expression

$$E_s = \frac{-jkQ_s\sqrt{L_s/L_p}}{1+jQ_s y} \cdot E_p$$

where $y=2\delta f/f_o$, f_o is the resonance frequency of the secondary circuit, and δf is the difference between the frequency of the applied signal and the resonant frequency. This relationship is true whether the primary circuit is tuned or not.

At the resonance frequency the secondary voltage lags on the primary voltage by 90°. At a signal frequency displaced by δf from resonance the phase shift increases to 90° plus an angle given by $\tan^{-1} -Q_s y$. This suggests that if it is possible to produce a signal with a magnitude dependent upon the phase angle between the two signals, a detector for f.m. signals will result. This is the principle embodied in the gated-beam discriminator. There is an additional complication in that both voltages tend to vary in amplitude with δf, so the detector must be insensitive to such variations. If this condition is met, the detector is similarly insensitive to a.m. of the original signal and no separate limiter stage will be required.

The properties required in the detector can be realized by utilizing two input electrodes of a multi-element valve, such as a pentode. Ideally, such a pentode should have a control grid and a suppressor grid which have characteristics of the type shown in Fig. 2. The grid base should be short, and in the positive region the anode current should not vary with the bias; additionally, grid current should be small, to minimize damping of the input circuit. In an ideal pentode, the control grid determines the space current (anode and screen) through the valve, whilst the suppressor grid controls the ratio in which this space current divides between anode and screen. As the suppressor grid is biased negatively, a retarding field is set up in front of the anode, and an increasing proportion of the space current is reflected to the screen grid. When the suppressor grid is driven positive, the anode current does not increase appreciably above its value for zero suppressor bias, because all the electrons which pass the screen grid mesh must travel to the anode; the total current is not affected since this is determined solely by the control-grid and screen-grid potentials. Thus the ideal characteristic is approached fairly closely by a practical anode-current/suppressor-bias characteristic. The ideal characteristic is difficult to realize at the control grid because of the grid current which flows when the grid is driven positive. To obtain the desired performance a special form of construction has to be adopted, as in the valve type 6BN6.

An alternative way of avoiding this difficulty is to employ a multi-electrode valve with two "suppressor" grids, neither of which is immediately adjacent to the cathode. Such a valve is the nonode type EQ80, which has nine electrodes, as shown in Fig. 3.

In addition to the control grid proper, there are

*B.B.C. Engineering Training Department.

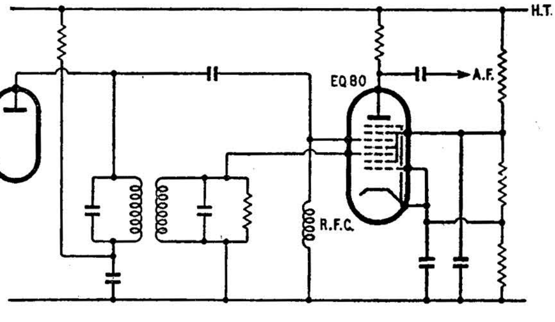


Fig. 3. Circuit for use with nonode discriminator.

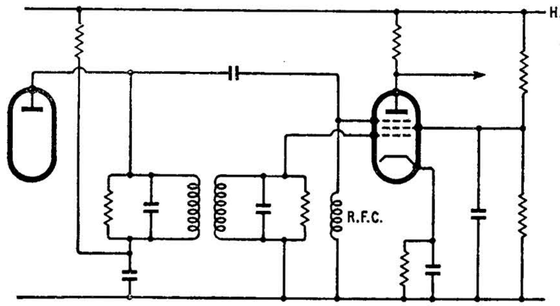


Fig. 4. Circuit for use with gated-beam discriminator.

two short-base grids, to which the input signals are applied. There are also three "screen" grids which serve to maintain the potential gradient through the valve and screen the input circuits from one another. The control grid may be biased to set the quiescent current through the valve.

The simple circuit employing an "ideal" pentode will serve to illustrate the method of operation of this type of detector; the circuit arrangement is as shown in Fig. 4. The quiescent bias at each grid is adjusted so that each is at the mid-point of its characteristic. The anode current of the valve is then one-quarter of the maximum value, which occurs when both grids are simultaneously at zero bias. The input signals applied to the two grids are taken from the primary and secondary circuits of the coupled pair. The coupling factor (kQ) is usually in the region of unity, so that approximately equal primary and secondary voltages exist at resonance. If the signal voltages are sufficiently large, both grids are heavily overdriven. Consider now an input applied to one grid alone; anode current will flow in pulses, having a mark/space ratio of unity, as shown in Fig. 5. However, with an input to both grids, anode current can flow only when the signal at each grid is within the grid base. This is shown in Fig. 6, which shows the effect of applying each signal separately and together. The period of anode current flow is proportional to the overlap of the two sets of pulses. The amplitude of each resultant pulse is constant and hence independent of the input signal magnitude, so long as the condition of overdriving at each grid is maintained. The period of overlap of the pulses is proportional to the phase angle between the two sine waves giving rise to the pulses. At resonance, the phase difference is 90°, i.e., one-quarter of the wave period. Hence the mean anode current is one-quarter of the maximum current, i.e., it is equal to the anode current in the absence of input signals.

When the frequency of the input signal changes, the period of overlap changes, and hence the mean anode current varies with the signal frequency. Thus the audio output is directly proportional to the departure of the phase angle between the two input signals from the 90-degree condition at resonance. It was shown earlier that this phase change is equal to $\tan^{-1} - Q_s y$, where $y = 2\delta f/f_o$. The graph of audio output plotted against frequency shift thus has the form shown in Fig. 7. In practice, the curve has turnover points, due to the selectivity of the tuned circuits, which reduces the drive to the grids. Typical turnover points are shown dotted in Fig. 7. From Fig. 7 it will be seen that the input signal frequency shift/output characteristic is not truly linear anywhere, but offers a fair approximation to linearity in the region near the centre frequency. For a fixed frequency deviation, improved linearity can be obtained if the value of Q_s is lowered. However, this process cannot be carried too far, or difficulties arise in obtaining sufficient input signal for satisfactory limiting.

The expression for the audio output may be expanded as a power series as follows.

$$E \propto \frac{-Q_s}{f_o}\,\delta f + \frac{1}{3}\left(\frac{Q_s}{f_o}\right)^3 \delta f^3 \ldots$$

With the EQ80 type of gated-beam discriminator an input of some 8 volts r.m.s. is required at each grid for satisfactory limiting to commence. This somewhat low sensitivity is probably one of the major reasons why this type of valve has not been more widely used. The audio output is of the order of 10 volts r.m.s. for a deviation of 75 kc/s; this is usually sufficient to drive an output stage directly without an intervening audio amplifying stage. The a.m. suppression ratio is between 25 and 30 dB, and this falls below the desirable limit of 35-40 dB. (The a.m. suppression ratio was defined in Part 3 as the ratio of the audio outputs due to the f.m. and a.m. components of an input signal simultaneously modulated by a.m. and f.m. to a modulation depth of 30 or 40 per cent.)

The 6BN6 gated-beam discriminator was discussed in detail in *Wireless World* (January 1957) by Lawrence W. Johnson, and reference should be made to this article for circuit details, operating conditions, etc. The a.f. output obtainable from this valve is of the order of 15 volts r.m.s. for a deviation of 75 kc/s. The input signal amplitude required at the control grid for limiting is 2 to 3 volts r.m.s. The a.m. suppression ratio is between 25 and 30 dB. This is below the desirable limit, and it would appear that the 6BN6 should be preceded by a further limiter. This reduces the attractiveness of the circuit, since its chief merit lies in its simplicity and cheapness.

The harmonic distortion can be evaluated approximately by means of the expansion for the a.f. output given previously. If the modulating signal is $f_d \cos \omega t$, the output is given by

$$E \propto \frac{-Q_s}{f_o} f_d \cos \omega t + \frac{1}{3}\left(\frac{Q_s}{f_o}\right)^3 f_d^3 \cos^3 \omega t$$

$\cos^3 \omega t$ may be expanded in terms of $\cos \omega t$ and

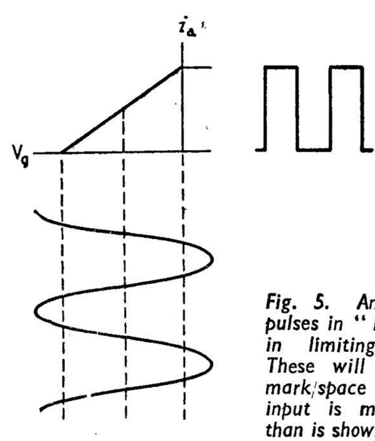


Fig. 5. Anode current pulses in "ideal" valve in limiting condition. These will have unity mark/space ratio if the input is much greater than is shown here.

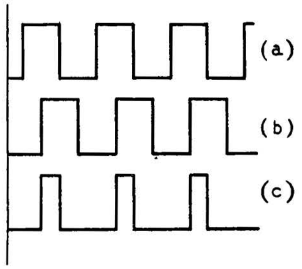


Fig. 6. (a) Anode current pulses due to input at control grid alone. (b) Anode current pulses due to input at suppressor grid alone. (c) Anode current pulses with inputs at both control and suppressor grids.

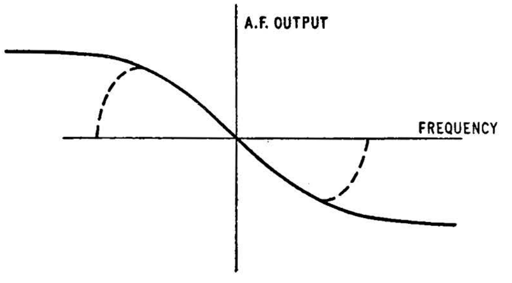


Fig. 7. A.F. output against frequency for gated-beam discriminator, assuming limiting at all frequencies. The dotted curve is that obtained in practice due to falling-off of signal amplitude with circuit selectivity.

cos $3\omega t$, and the percentage of third harmonic distortion shown to be

$$\frac{1}{12}\left(\frac{Q_s f_d}{f_o}\right)^2 . 100$$

With $Q_s = 35$, $f_d = 75$ kc/s and $f_o = 10.7$ Mc/s, the third harmonic distortion is approximately 2 per cent.

There is one feature of the 6BN6 circuit given by L. W. Johnson which is not immediately apparent: this is the mechanism of coupling between the primary and secondary circuits. The circuit arrangement is as shown in Fig. 8, and at first sight there is apparently no coupling between the two circuits. In fact, there is the equivalent of top-end capacitance coupling, with the somewhat unusual feature that the coupling capacitor is a negative capacitor, i.e., it has positive reactance, like an inductor, but the magnitude of the reactance decreases with increasing frequency, as with a capacitor.

The mechanism of coupling is as follows. The input " primary " circuit voltage controls the total electron stream through the valve, and hence the anode current flowing past the suppressor grid is modulated at the input signal frequency. Now if an electric charge is brought near a conductor connected to earth, there is a movement of charge to the face of the conductor tending to neutralize the field of the approaching charge. This is a familiar phenomenon in electrostatics. A positive change of grid potential increases the number of electrons flowing through the valve, and hence increases the number of electrons in the vicinity of the suppressor grid. There is then an increase of the positive charge on the suppressor grid, which is the conductor past which the electron stream is flowing; and there is a movement of electrons from the suppressor grid through the external circuit.

If a change of grid voltage dv produces a change of the charge dq, in the neighbourhood of the suppressor grid, we may write

$$dq = -\, a.dv$$

where a is a positive constant.

The reason for the negative sign is that a positive increment dv increases the number of electrons in the vicinity of the suppressor grid, and since these are negatively charged there is a negative increment of charge. The increase of the charge near the suppressor grid induces a proportional charge dq' flowing out of the suppressor grid, and we may thus write

$$dq' = b.dq$$

where b is a positive constant.

Thus

$$dq' = -\, a.b.\, dv$$

This may be compared with the relationship for a capacitor $Q = CV$.

From this it appears that the electron stream coupling is equivalent to a negative capacitor of magnitude $a.b$ connected directly between control grid and suppressor grid. This form of coupling occurs in all multi-electrode valves. In particular its effect has long been recognized in frequency changers where on short wavebands it may induce " pulling " of the local oscillaor frequency. In this application neutralizing is effected by means of a small (positive) capacitor connected externally between the electrodes.

The degree of coupling obtained by this means is insufficient in the 6BN6 to produce adequate voltage drive at the suppressor grid, and is supplemented by means of the undecoupled anode lead resistor R, shown in Fig. 8. This resistor is of low value, usually a few hundred ohms. Under working conditions,

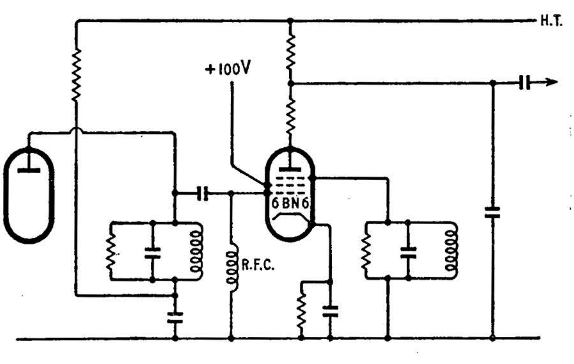


Fig. 8. Circuit used with 6BN6 gated-beam discriminator.

a voltage is produced across it which is in anti-phase with the control-grid voltage. We may write this anode voltage as

$$E_a = -\, c.\, dv$$

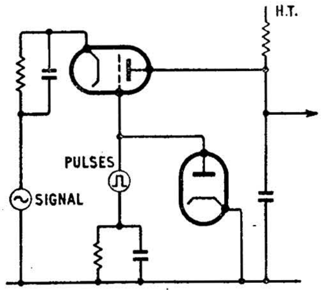
where c is the working gain of the valve at r.f. There is a physical capacitance which we may designate C_{a-su} between the anode and the suppressor grid, and hence current is fed through this capacitor to the " secondary " circuit. If the impedance of this circuit is Z, then the current i is given by

$$i = -\, c.\, dv/(Z + 1/j\omega C_{a=su})$$

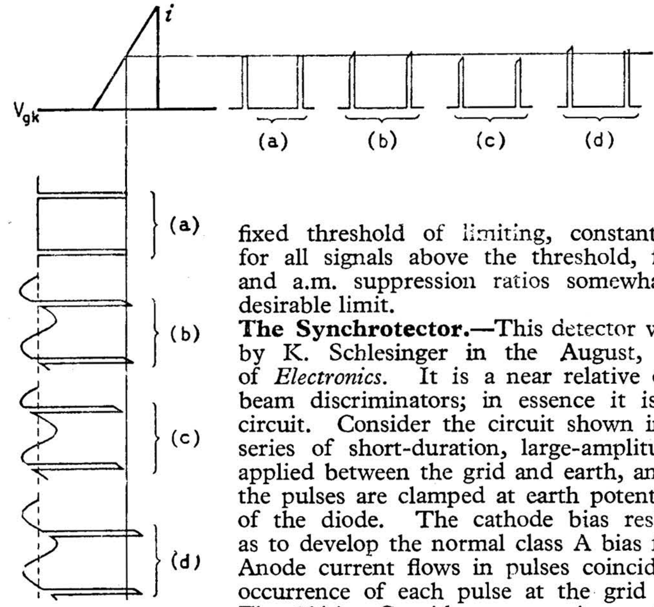
If the reactance of the capacitor is appreciably greater than Z, we may use the approximation

$$i = -\, c.dv.j\omega C_{a-su} = c.dv.j\omega(-C_{a-su})$$

i.e. this coupling also behaves like a negative capacitor connected directly between control and suppressor grids, and hence supplements the equivalent capacitance existing already.



Above: Fig. 9. Basic circuit of the synchrotector.



Right: Fig. 10. Showing how a.f. output from the synchrotector varies with phase angle between applied signal and sampling pulses; (a) no input signal (b) 90-degree phase shift (c) phase shift greater than 90 degrees (d) phase shift less than 90 degrees.

An important distinction exists between two circuits coupled together physically, and the two circuits coupled together by the electron stream of a valve, as in the 6BN6. In the former case, the energy in the secondary circuit is supplied from the primary circuit, and consequently the shape of the resonance curve of the primary circuit is affected by the coupling to the secondary circuit. The response exhibits "rabbit's ears" similar to those obtained across the secondary circuit when the coupling factor (kQ) exceeds the critical value. However, the primary circuit "rabbit's ears" are more widely spaced in frequency, and more pronounced. When the coupling is via the valve electron stream, the energy in the "secondary" circuit is supplied by the electron stream, and consequently the "primary" circuit resonance curve is unaffected by the presence of the "secondary" circuit. As a corollary of this, the resonance curve of the secondary does not develop "rabbit's ears" as the coupling factor increases, but remains single-peaked.

An interesting variant of the 6BN6 has recently appeared in the U.S. This valve is the 6DT6. This has sufficient internal coupling via the electron stream to produce adequate drive at the suppressor grid. At low input signal levels there is a gain from control grid to suppressor grid, and this fact is utilized to make the circuit self-oscillating at small signal inputs. The physical capacitance between the control and suppressor grids is made sufficiently large to maintain oscillation in the absence of an input signal, the suppressor grid functioning as an "anode." The oscillator is then of the tuned-anode, tuned-grid type. With a small input signal, the detector functions as a locked-oscillator limiter, as well as a detector. This action lowers the threshold value of input signal with which the detector will work; the 6BN6 requires an input of the order of 1 to 2 volts, whilst the 6DT6 requires an input of only 0.3 to 0.5 volts.

The properties of the gated-beam discriminators as a class may be summarized as good sensitivity, fixed threshold of limiting, constant a.f. output for all signals above the threshold, fair linearity, and a.m. suppression ratios somewhat below the desirable limit.

The Synchrotector.—This detector was described by K. Schlesinger in the August, 1956, issue of *Electronics.* It is a near relative of the gated-beam discriminators; in essence it is a sampling circuit. Consider the circuit shown in Fig. 9. A series of short-duration, large-amplitude pulses is applied between the grid and earth, and the tips of the pulses are clamped at earth potential by means of the diode. The cathode bias resistor is such as to develop the normal class A bias for the valve. Anode current flows in pulses coincident with the occurrence of each pulse at the grid as shown in Fig. 10(a). Consider now an input signal applied to the cathode, the frequency of the signal being the same as that of the grid pulses. The mean anode current will now vary according to the phase relationships between the applied pulses and the signal.

If the pulses occur at the instants when the signal amplitude is passing through zero, the anode current pulse is of the same amplitude as it is in the absence of the signal, as shown in Fig. 10(b). If the pulses occur when the signal is positive with respect to earth, the anode current pulses will be smaller in amplitude, because this condition is equivalent to a negative signal in the grid circuit, as shown in Fig. 10(c). Conversely, if the pulses occur when the signal is negative with respect to earth, the anode current pulses will be larger in amplitude, as shown in Fig. 10(d). Thus it is possible to construct a discriminator if the phase angle between the pulses and the applied signal can be varied with the signal frequency. A suitable circuit arrangement is that shown in Fig. 11. The grid pulses are now sine waves generated across a tuned circuit, fed by a small top-end capacitance from the applied signal source. As shown earlier, the phase relationship between the "secondary" circuit signal and the applied signal varies with the signal frequency, being 90° at a frequency near the secondary circuit resonance frequency. This can be shown simply for the top-end capacitor coupling circuit by Thévenin's theorem. The circuit of Fig. 12(a) is equivalent to that of Fig. 12(b), and it can be seen that the "secondary" voltage E_s is at 90° with respect to the "primary" voltage E_p when the secondary inductance is resonant with the capacitors C_s and C_t in parallel, i.e., at a frequency slightly below the resonance frequency of L_s and C_s alone.

In the practical circuit the voltage at the grid is about 3 to 4 times that at the cathode, so that the periods when the valve is conducting are relatively short. By assuming the pulses to be very short, it is possible to derive an approximation for the variation of anode current with signal frequency. The amplitude of the anode current pulses is proportional to the sine of the angle between the zero value of the applied signal and peak value of the sampling pulse. Using $y = 2\delta f/f_0$ where δf is the difference between the signal frequency and the resonance frequency of the secondary circuit, this angle θ is given by

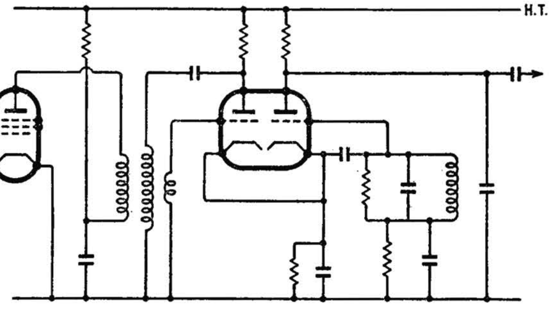
$$\theta = \tan^{-1} - Q_s y$$

Hence the amplitude of the anode current pulses is proportional to $\sin(\tan^{-1} - Q_s y) = -Q_s y/(1 + Q_s y)^{\frac{1}{2}}$. Provided that Q_s is small, the anode current is approximately linearly related to the signal frequency shift in a small region near the resonance frequency of the secondary circuit.

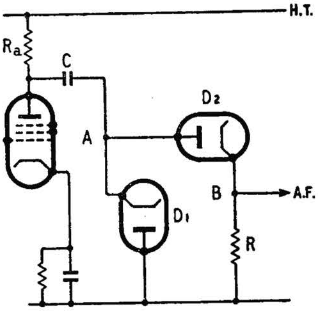
There is some degree of limiting action, since an increase of signal amplitude produces an increased amplitude of the "sampling" pulses. This results in the conduction period being shortened, which tends to reduce the increase of anode current which would otherwise occur. The circuit described by Schlesinger is shown in Fig. 13. The discriminator proper is driven from a locked-oscillator limiter. The circuit was designed for use with the U.S. television inter-carrier sound system, which employs a deviation of 25 kc/s at a carrier frequency of 4.5 Mc/s. The circuit is stated to give an audio output of 25 volts peak-to-peak, for an r.f. input to the driver stage of 6 millivolts. This represents a high conversion efficiency, being better than a comparable ratio detector circuit employing a driver stage and a double-diode-triode, the latter valve providing the detector diodes and a.f. amplifier. The circuit is claimed to have an a.m. suppression ratio greater than 40 dB.

Counter Circuit.—If the incoming signal can be converted to a train of constant-amplitude pulses, demodulation can be effected by means of a "counter" circuit, which gives an output proportional to the repetition rate of the pulses. This type of circuit was discussed in some detail in the April, 1956, issue of *Wireless World* by M. G. Scroggie. The basic circuit considered by Scroggie is shown in Fig. 14. The incoming signal is heterodyned to produce an intermediate frequency signal at 200 kc/s approximately. After amplification, the signal is applied to a limiter stage, which gives a square wave output. The pulses are then applied to the diodes D1 and D2.

Consider first the quiescent condition with the limiter stage cut-off. The anode potential is that of h.t., and there is no voltage across either diode, or the diode load resistor R. If now the limiter anode potential falls as its grid is driven positively, diode D1 conducts; diode D2 remains cut-off. Because of the high ratio of the resistance R_a to that (R_{di}) of the diode D1 when conducting, the cathode of D1 is not driven appreciably negative with respect to earth,



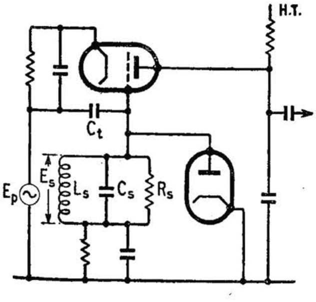
Above: Fig. 13. Practical circuit of synchrotector preceded by locked-oscillator limiter.



Right: Fig. 14. Basic "counter" circuit.

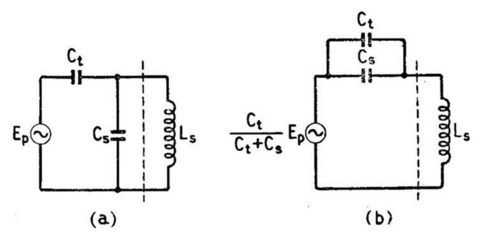
and capacitor C discharges through R_a and diode D1 is series until the cathode of D1 returns to earth potential. This is shown in Fig. 15(a). The discharging curve is exponential, and thus an infinite time is required theoretically for the cathode of D1 to reach earth potential; in practice the time constant $(R_a + R_{di})C$ is sufficiently small for the potential of D1 cathode to be indistinguishable from earth before the next part of the cycle. After a period equal to half the signal period, the anode of the limiter is driven positive, as its grid is driven negative to beyond cut-off. The anode potential then commences to rise to h.t. potential, and current flows through R_a, C, D2 and R in series; the voltage across R is shown in Fig. 15(b). The time constant of the combination is such that the voltage pulse developed across R has virtually disappeared before the next change of limiter anode potential occurs, when the cycle is repeated. There is thus a train of pulses developed across R, the area (volt-secs) of which is independent of the magnitude and frequency of the input pulses. However, the mean voltage output is equal to the area of these pulses multiplied by the rate at which they occur, and this rate is, of course, equal to the input signal frequency. Thus the output voltage is apparently linearly related to the input signal frequency.

The linearity is, however, not perfect because the capacitor C cannot charge completely through R and



Left: Fig. 11. Circuit of synchrotector with sampling pulses derived from input signal.

Below: Fig. 12. Thévenin's theorem applied to circuit (a) to give circuit (b).



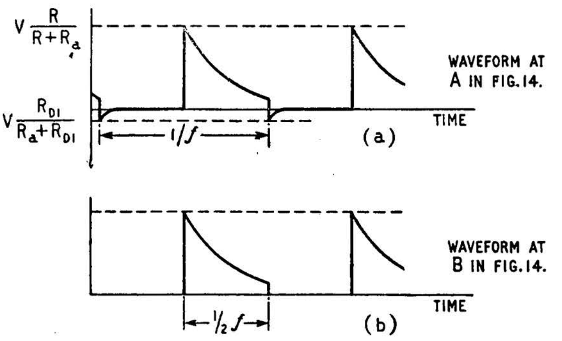


Fig. 15. Waveforms at points A and B of Fig. 14.

R_a in the half-cycle period, as required. If the time constant is made very short to approach this condition of perfection, the area of the pulses becomes smaller and the a.f. output decreases. In a practical circuit, the component values adopted must be a compromise between the requirements of good linearity and sensitivity. In addition, in the circuit described by Scroggie, there is a limitation placed upon the value of R_a by the limiter requirements.

The degree of non-linearity can be calculated as follows. If the signal frequency is f, then the time of one pulse cycle is $1/f$. The combination of R_a, C and R is thus charging for a period $1/2f$. If the amplitude of the voltage step at the limiter anode is V volts, then if the diode forward resistance is negligible, the voltage across R at the beginning of the charging period is $VR/(R + R_a)$. At the end of the period this voltage has fallen to

$$\frac{VR\, e^{-1/2fCR'}}{R + R_a}$$

where $R' = R + R_a$

The area of the pulse is given by

$$\frac{VR}{R+R_a} \int_o^{1/2f} e^{-1/CR't}\, dt$$

which is equal to

$$VCR(1 - e^{-1/2fCR'})$$

The a.f. output is equal to the product of this area and the signal frequency, i.e., $VCRf\ (1-e^{-1/2fCR'})$. This may be compared with the "ideal" output, $VCRf$, obtained if the time constant CR' is very small. Thus the second term within the bracket represents the departure from linearity. It is minimized if f, C and R' are small. However, the expression for the output voltage shows that if V, C, R and f are reduced to minimize non-linearity, the a.f. output will fall. Thus a compromise is required. The minimum value is further determined by the consideration that the signal frequency should not be allowed within the a.f. spectrum; thus with a deviation of 75 kc/s, the centre operating frequency must be above 90 kc/s, and preferably higher, to allow some margin for mistuning, drift, etc. Thus a centre signal frequency of 150–200 kc/s is usually employed. The use of a low-value intermediate frequency such as this brings other difficulties in its train, notably those of obtaining adequate i.f. selectivity, and the maintenance of second channel protection, since the second channel is only 300–400 kc/s removed from the wanted carrier frequency. The circuit response curve is markedly unsymmetrical, having a comparatively large linear portion in the direction of increasing frequency, and a comparatively small linear portion in the direction of decreasing frequency.

In the circuit described by Scroggie, the value of $R_a = R = 4.7\ k\Omega$, $C = 50\ pF$ and $V = 60$ volts. The centre frequency is 150 kc/s. With these values, the r.m.s. a.f. output for 75 kc/s deviation is 0.8 volt. Scroggie also plotted the departure from linearity against frequency; the curve obtained agrees well with the curve obtained from the calculation given previously. The distortion, computed by Scroggie, was about 0.5 per cent second harmonic at maximum deviation.

The degree of a.m. rejection cannot be specified, since it is a function of the limiter performance; in general, it should be possible to realize a satisfactory performance in this respect. With regard to sensitivity, the circuit compares closely with the Foster-Seeley circuit, requiring about 2 volts input at the limiter grid for an output of about 1 volt. It has a fixed threshold of limiting, and constant audio output for all input signals above this threshold. As with the Foster-Seeley circuit, the maximum degree of "downward" a.m. handling capacity is dependent upon the margin by which the signal at the limiter grid exceeds the limiter threshold.

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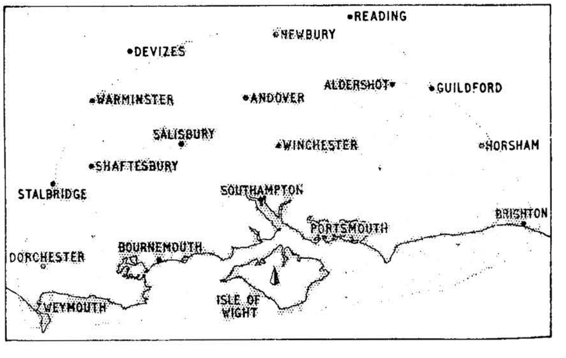
Southern I.T.A. Station



APPROXIMATE service area of the I.T.A.'s seventh transmitter, to be built at Chillerton Down, Isle of Wight, is shown shaded on this sketch map. It will probably come into service in the late spring of next year. No announcement has yet been made by the Post Office regarding the channel in which the station will operate. It is unlikely to use one of the three channels so far allocated to the I.T.A. owing to its geographical position in relation to the stations already operating in them.