Limiters and Discriminators for

5.—MEASUREMENT OF A.M. SUPPRESSION RATIO : TYPES OF LIMITER

THE purpose of a limiter stage in a f.m. receiver is to reduce the magnitude of the amplitude-modulation component of an applied signal. The performance of a limiter may be judged by the degree of reduction of the modulation depth which it achieves, and this degree of reduction may be termed the a.m. reduction factor. However, a figure for the reduction of the modulation depth would not be directly applicable to self-limiting detectors such as the ratio detector. Furthermore, many discriminators have some degree of inherent a.m. rejection. For example, with a perfectly balanced Foster-Seeley discriminator there is zero output when the carrier is at the centre frequency and hence no output if the carrier is amplitude modulated. Thus a criterion is required which provides a figure of merit for a combination of limiter and discriminator, and which is applicable to self-

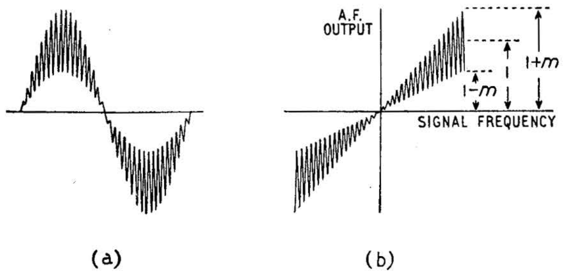

Fig. 1. Oscillograms of a.f. output from perfectly balanced Foster-Seeley discriminator with input signal simultaneously modulated in amplitude and frequency (a) with sawtooth time-base (b) f.m. component applied to X plates as time base.

limiting discriminators. Such a criterion is obtained by employing a test signal modulated simultaneously in frequency and amplitude, and measuring the ratio of the components of the output signal due to the two components of the signal modulation respectively. This criterion is termed the *a.m. suppression ratio.* If a limiter has a given a.m. reduction factor, it would appear that its effect, in combination with a given discriminator, would be to increase the a.m. suppression ratio by an amount equal to the a.m. reduction factor. However, this is not always true as the limiter may introduce spurious f.m. in the course of its limiting action.

However, there is general agreement as yet as to the test conditions to be employed in making a measurement of a.m. suppression ratio; the result is that any one device may have a number of values of a.m. suppression ratio, depending on the conditions of measurement. There is also the undesirable result that the minimum figure for the a.m. suppression ratio of a receiver, under given conditions of operation, is subject to a margin of uncertainty, unless the conditions adopted for the measurement of the a.m. suppression ratio are defined. In this article, the definition and measurement of the a.m. suppression ratio is that recommended by the B.B.C. Research Dept. The modulation depth of both a.m. and f.m. components of the test signal is 40 per cent ($\pm$30 kc/s frequency swing for the f.m. component) and the frequencies of the f.m. and a.m. components are 100 c/s and 2 kc/s respectively. The measurement is carried out as follows.

The apparatus under test is fed from a signal generator which can be modulated simultaneously in amplitude and frequency. The a.f. output is fed to a power-measuring instrument, preceded by a standard aural weighting network. (This network has a frequency response substantially flat in the region of 2 kc/s, where measurements are made, and therefore does not appreciably affect the measurements, save in exceptional circumstances, as, for example, when the a.m. component produces a substantial output of high-order harmonic components.) The input signal is set to the required amplitude for the test, and is first modulated by f.m. only to a swing of 35 kc/s by a signal at 2 kc/s. The a.f. power output P_1 is then measured. (If the apparatus under test is a complete receiver, the gain control is adjusted to give the standard power output of 50 mW). The value of 35 kc/s frequency swing is greater than the modulation depth of 40 per cent specified, and is used to allow for the effect of a 50 microsecond de-emphasis network, so that the a.f. power output is the same as that which would be obtained with a signal of $\pm$30 kc/s frequency swing at 100 c/s.

The frequency-modulating signal is then set to 100 c/s, and the frequency swing to $\pm$30 kc/s. The amplitude modulation signal at 2 kc/s is then applied, with a modulation depth of 40 per cent. The fundamental-frequency component in the output due to the frequency modulation is then filtered out by means of a high-pass filter having a cut-off frequency of 250 c/s. The a.f. power output P_2 due to the remainder of the signal is then measured. The a.m. suppression ratio is then the ratio of P_1 to P_2 expressed in decibels.

It was mentioned earlier that the Foster-Seeley discriminator has some degree of inherent a.m. rejection, and its a.m. suppression ratio can be calculated as follows. In an ideal discriminator of this type, the a.f. output is very nearly equal to I.f, over a region near the centre frequency, where I is the magnitude of the input signal current, and f is the difference between the signal frequency and the centre frequency. If the input current I is amplitude modulated, its value is given by $(1+m \cos\omega_2 t)$ I, where m is the amplitude modulation depth expressed as a fraction, $\omega_2 = 2\pi f_2$, and f_2 is the frequency of the a.m. component. If the f.m. component has a frequency of f_1, then the a.f. output is proportional to $(1+m \cos\omega_2 t) \cos\omega_1 t$. Such an output gives rise to oscillograms of the type shown in Fig. 1. The expression for the a.f. output

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waveform is precisely similar to the expression for an amplitude-modulated carrier. It can be resolved into three components, corresponding to the carrier and sidebands of an a.m. signal. However, the " carrier " component in this case is the fundamental frequency component of output due to the frequency modulation, and the power output due to this component alone may be taken as unity. When this component due to the frequency modulation is filtered out, there remains two components due to the amplitude modulation. These are of equal amplitude, $m/2$, and have frequencies $(f_1 - f_2)$ and $(f_1 + f_2)$. With the test frequencies postulated, i.e., 100 c/s for the f.m. component and 2 kc/s for the a.m. component, these two residual components of the a.f. output have frequencies of 1900 c/s and 2100 c/s. The total power output is then proportional to $2(m/2)^2 = m^2/2$. With the value of m postulated, 0.4, this power output is proportional to 0.08. Thus the ratio of the power outputs due to the f.m. component to that due to the a.m. component is 1 : 0.08, i.e., some 11 dB. However, this figure includes no allowance for the effect of the de-emphasis network, which introduces a loss of some 1.5 dB in the region of 2 kc/s; thus the value of the a.m. suppression ratio is some 12.5 dB.

The effect of a change of test conditions upon the value of the a.m. suppression ratio can be judged from the foregoing; if the amplitude modulation depth used had been 30 per cent., the a.m. suppression ratio would be 15 dB.

If the input signal carrier frequency is not precisely at the centre frequency of the discriminator characteristic, the a.m. suppression ratio alters. The output signal, with the input signal simultaneously modulated by a.m. and f.m., is then proportional to $(1 + m \cos \omega_2 t)(F \cos \omega_1 t - f_0)$, where f_0 is the displacement of the carrier frequency from the centre frequency of the discriminator, and F is the frequency swing of the f.m. component. The types of oscillogram obtained under these conditions are shown in Fig. 2. In addition to the two components, each of amplitude $m/2$ relative to the fundamental frequency f.m. component, discussed previously,

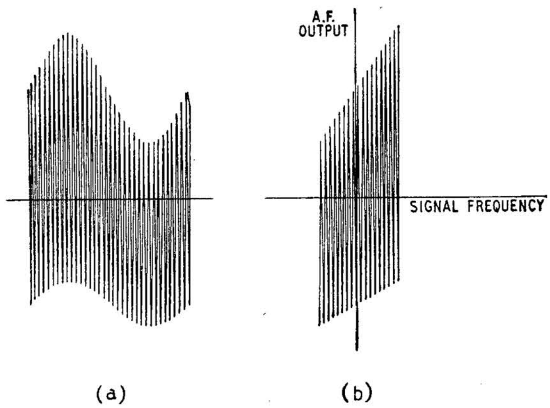


Fig. 3. Oscillograms of a.f. output from a counter-type discriminator with input signal modulated simultaneously in amplitude and frequency (a) with sawtooth time base (b) f.m. component applied to X plates as time base.

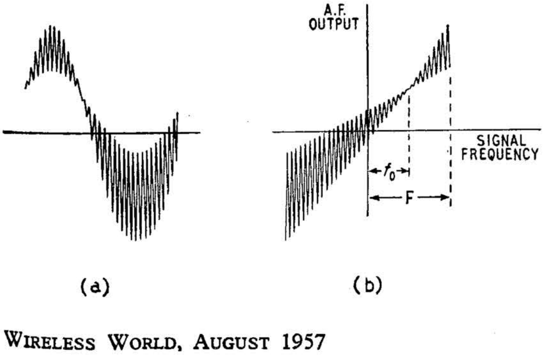
there is now a further a.f. component of amplitude mf_0/F at the a.m. modulating frequency f_2. The power output due to the a.m. component is thus increased to $(m^2/2) + (m^2 f_0^2/F^2)$. If, for example, the carrier frequency is displaced 30 kc/s from the centre frequency, the power output due to the a.m. component rises from 0.08 to 0.24, i.e., the a.m. suppression ratio falls by some 4.8 dB. to 7.7 dB.

This special case was considered because it can be extended to other discriminators. For example, the counter-type discriminator gives zero output at zero frequency, and its output is proportional to the input signal amplitude and the signal frequency. Thus if a counter-type discriminator is operated at a centre frequency of 200 kc/s, a typical figure, the value of f_0 is 200 kc/s. The type of oscillogram obtained with the input signal modulated simultaneously in amplitude and frequency is shown in Fig. 3. Applying the formulae given previously, the a.m. suppression ratio is $-$ 6.1 dB. Thus a counter-type discriminator requires that very careful attention be given to the performance of the preceding limiter.

The minimum desirable value for the a.m. suppression ratio measured in the manner described above, depends upon the type of a.m. interference encountered, and upon the class of service desired. For a broadcasting service, giving a signal output of good quality, a minimum value of 30 dB would seem to be necessary, and it would appear preferable that the ratio should be greater than 35 dB. It is doubtful if any aural change is perceptible if the value is increased beyond 40 dB.

Although the a.m. suppression ratio provides an excellent criterion of performance, another important factor must not be overlooked. This is the range of input signal amplitude over which a.m. rejection is maintained. The test for a.m. suppression ratio

Fig. 2. Carrier frequency displaced by frequency f_0 from centre frequency.



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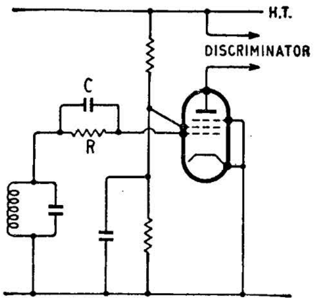

Fig. 4. Basic circuit of grid limiter.

explores limiting performance over a range of 40 per cent. modulation depth, but it is important to know the maximum modulation depth which the limiter can handle. Dependent upon the type of limiter, there may be a minimum signal amplitude below which limiting action fails, or alternatively, there may be a maximum modulation depth which the limiter will handle. In general, limiting action fails when the signal amplitude is decreasing. If the limiter has a minimum value of input signal below which limiting action fails, the limiter is said to have a threshold value. The maximum modulation depth in the "downward" direction then varies with the amount by which the mean signal amplitude exceeds the threshold. With limiters of the type which have a fixed "downward" modulation handling capacity, there is no fixed threshold. However, with such limiters, the a.m. suppression ratio generally falls with mean carrier amplitude, and there is thus a quasi-threshold fixed by the input signal amplitude above which the a.m. suppression ratio is satisfactory.

For most purposes, it would seem desirable that satisfactory limiting action should be maintained for "downward" modulation of 50–70 per cent., although in some locations the depth of amplitude modulation due to reflections may exceed this value. Thus if a limiter has a threshold input of 1 volt for satisfactory limiting, the mean carrier level should normally be greater than 2–3 volts. With a limiter or self-limiting discriminator (e.g., ratio detector), having maximum "downward" modulation handling capacity, this modulation handling capacity is fixed in the course of design and should be in the region 50–70 per cent.

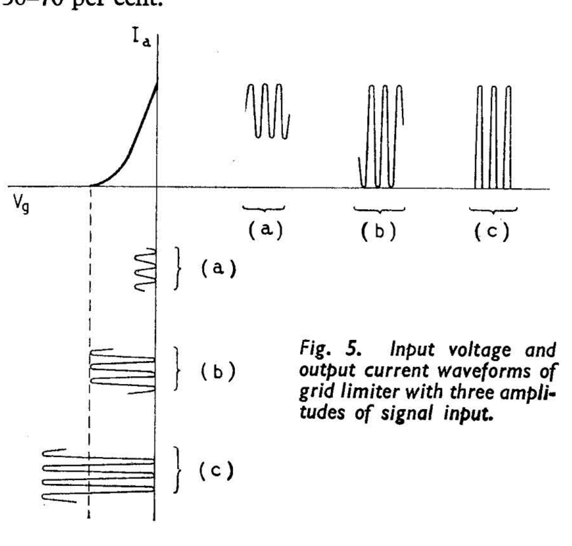

Fig. 5. Input voltage and output current waveforms of grid limiter with three amplitudes of signal input.

Grid Limiter. The grid limiter has been widely used in f.m. receivers, particularly in conjunction with Foster-Seeley discriminators. It comprises a pentode operated with a low screen voltage, fed at the control grid through a self-biasing network, as shown in Fig. 4. No cathode bias is usually employed, and one of the reasons for employing a low screen voltage is to prevent the valve drawing excessive current in the absence of an input signal. The low screen voltage also results in the valve having a short working grid-base, so that the limiting action commences with a relatively small input. The action of the circuit is as follows.

The grid and cathode of the valve function as a diode detector. When the input signal is positive-going, grid current flows at the tip of the cycle, and the capacitor C is charged, biasing the valve. With a reasonably large value grid resistor R (i.e., large compared with the resistance of the grid-cathode path when conducting), the bias so adjusts itself that grid current flows only on the tips of the positive peaks of the input signal. As the input signal amplitude increase from zero, the valve behaves first as a class A amplifier. When, however, the input signal amplitude exceeds half the cut-off bias, the valve is driven beyond cut-off on the negative-going peaks of the signal. As the input signal amplitude increases still further, the periods when the valve conducts become progressively shorter, as shown in Fig. 5.

The resultant anode current waveform may be

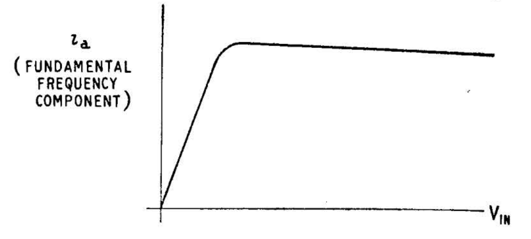

Fig.6. Fundamental frequency component of anode current plotted against input voltage for typical grid limiter.

analysed into components at the input signal frequency, and harmonics of this frequency. We may plot this fundamental-frequency component against input signal amplitude and obtain a curve of the type shown in Fig. 6. It will be seen that this fundamental-frequency component rises linearly at first with increasing input signal amplitude, until a "threshold" is reached, beyond which the output current is substantially independent of input signal amplitude. With a typical circuit of this type, comprising a pentode having a screen voltage of the order of 40–50 volts, the threshold input signal is of the order of 1 volt.

In practice, the portion of the curve above the threshold is not flat, but generally tends to fall slowly. This tendency may be minimized by careful choice of screen feed resistor, anode feed resistor and grid resistor. The selection of these components is usually done on test. A well-designed grid limiter will have an a.m. reduction factor of the order of 20-30 dB.

The grid limiter of the type described above has one major disadvantage. This is bound up with the time constant CR of the grid circuit components.

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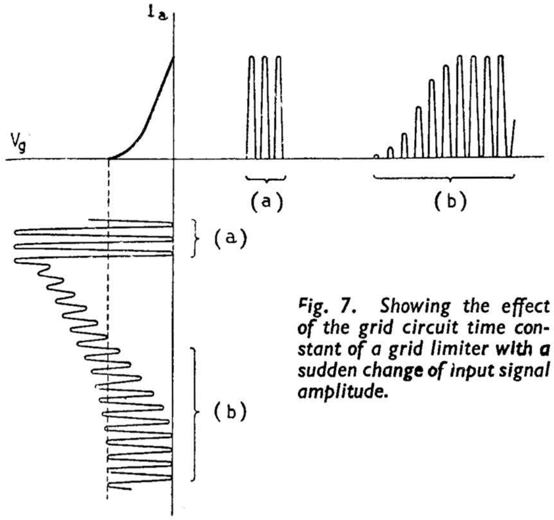


Fig. 7. Showing the effect of the grid circuit time constant of a grid limiter with a sudden change of input signal amplitude.

For satisfactory operation, the bias developed across the capacitor C should change instantaneously with any change of signal amplitude so that the positive-going peaks of the signal are always at zero bias. With an increasing signal amplitude, this condition is very nearly fulfilled, since the capacitor is charged through the relatively low resistance of the grid-cathode path. When, however, the input signal amplitude falls, the capacitor has to discharge through the resistor R, and whilst this is happening, the amplitude of the signal may be insufficient to cause the valve to conduct at all, or only partially. This is shown in Fig. 7. Thus, to ensure rapid discharge of the capacitor when the signal amplitude is decreasing, the time constant CR must be small. The desirable value of the time constant depends upon the maximum rate of change of input signal encountered, and the problem is the same as that of avoiding "tracking" distortion in conventional diode detector. It can be shown that this form of distortion can be avoided if $CR < 1/2\pi f\ m$, where f is the modulation signal frequency, and m is the maximum modulation depth. To consider the simplest case, co-channel interference, if the wanted signal is at one extreme of the frequency swing, and the unwanted signal is at the other, the wanted signal will be amplitude modulated at a rate of 150 kc/s. If the ratio of wanted to unwanted signal is 2:1, then m is 0.5, and in this case CR must be less than 2.1 microseconds approximately. In practice a value in the region of 2.5 microseconds is frequently adopted.

The choice of values for C and R individually is limited in two ways. If C is made small, a capacitance potential divider with the input capacitance of the valve is formed, and the signal appearing at the grid is materially reduced. Thus there is a lower limit to the value of C which can be tolerated; if the valve input capacitance is 10 pF, this minimum value is of the order of 20 pF. If the value of R is reduced, the damping of the tuned circuit feeding the limiter is increased. Since the grid and cathode of the valve behave as a diode detector, the damping resistance is given by $R/2\eta$, where η is the rectification efficiency of the circuit. Thus if the preceding circuit has a dynamic resistance of 20 kΩ, the minimum value of R would appear to be in the region of 40 kΩ, giving a reduction of approximately one-half in the working Q-value of the tuned circuit feeding the limiter. Values of C and R often adopted for a time constant of 2.5 microseconds are 50 pF and 47 kΩ respectively.

There is one variant of the grid limiter which is worthy of special mention. This is the form of grid limiter in which the valve employed has a high input resistance at the input grid, even when driven positive with respect to the cathode. Valves of this type have been discussed before in the section on gated-beam discriminators. Thus the valve type 6BN6 may be employed, and because of its high input resistance, it may be operated without the grid-current biasing network. The valve is biased to a class A operating condition in the absence of a signal by the cathode components, and when the signal amplitude is sufficiently large, square waves of anode current are generated, the magnitudes of which do not vary appreciably with input signal amplitude above the threshold value. This type of limiter thus has the advantage over the conventional type of grid limiter that its operation is not affected by the action of the grid circuit time constant. The damping presented to the input circuit under operating conditions is of the order of 20 kΩ.

A similar type of action to that described above may be secured with a normal pentode by inserting a resistor in series with the grid as shown in Fig. 8. When the signal drives the grid positive with respect to the cathode, the series resistor limits the flow of grid current, and ensures that the grid is not driven appreciably positive with respect to the cathode. The behaviour of the circuit is thus essentially similar to that of the 6BN6 type of limiter described above, the input signal driving the valve between anode current cut-off and its value at zero bias. However, there is one major disadvantage

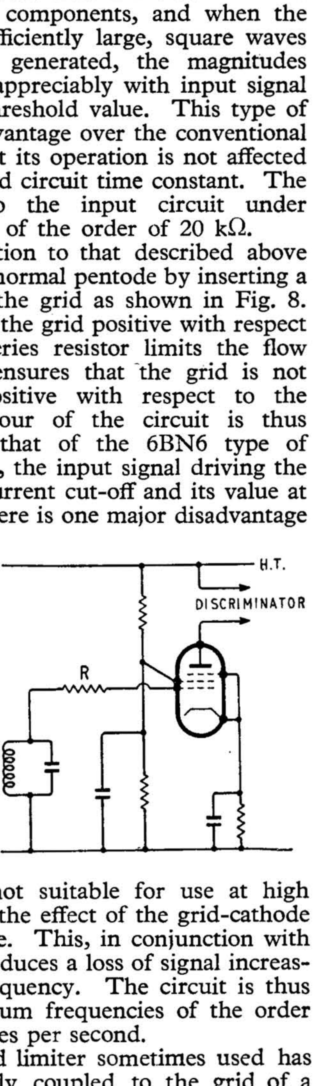


Fig. 8. Series-resistor grid limiter.

of this circuit: it is not suitable for use at high frequencies because of the effect of the grid-cathode capacitance of the valve. This, in conjunction with the series resistor, introduces a loss of signal increasing with increasing frequency. The circuit is thus useful only for maximum frequencies of the order of hundreds of kilocycles per second.

Another form of grid limiter sometimes used has a tuned circuit directly coupled to the grid of a pentode stage operated with a small bias. Grid current then damps the circuit, and gives a limiting action.

Anode Limiter.—The anode limiter relies for its action upon the existence of the "knee" of the I_a-V_a characteristic of a pentode. The valve is operated with a high value of anode load impedance, so that when the grid voltage is a few volts negative, the anode voltage "bottoms" at the value given by the intersection of the load line and the portion of the I_a-V_a curve below the "knee" of the characteristic, as shown in Fig. 9. The anode-current/grid-voltage characteristic is modified by the "bottoming" action to the form shown in Fig. 9(b). If the valve is biased to the working voltage V_g indicated in Fig. 9(b), symmetrical limiting of the input wave form occurs. Such a limiter was described by Scroggie* for use with a counter-type discriminator with the added refinement that a series grid resistor was used to assist the limiting action when the grid-cathode voltage was driven positive.

Dynamic Limiter—The dynamic limiter operates by presenting a varying impedance to a source of signal, the impedance varying in such a way as to tend to maintain the output signal at a constant amplitude. In its simplest form the circuit is shown in Fig. 10. We shall assume that the diode has a very low forward resistance compared with its load resistance. Then under quiescent conditions, the damping of the tuned circuit by the diode circuit is approximately equivalent to a parallel resistance R/2. This resistance is usually comparable with, or less than, the dynamic resistance of the tuned circuit. The time constant of the load circuit of the diode is made large compared with the period of the a.m. signals it is desired to suppress. Under quiescent conditions the diode load capacitor charges to a voltage just less than the peak signal voltage. If now the signal amplitude tends to increase, the diode will take increasing current on the peaks of the signal, in an attempt to increase the voltage across the capacitor. If the capacitor is large, however, the voltage across it cannot increase appreciably and the damping of the tuned circuit will increase, so that the load presented to the driving valve falls, and the gain from grid to anode decreases. The decrease in gain largely offsets the increase of input signal amplitude, so that there is only a small change in output signal amplitude.

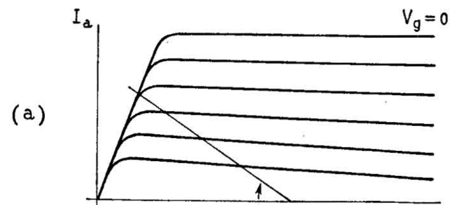
If the input signal amplitude decreases, the opposite effect occurs, the damping decreasing, so that the driving valve gain rises, offsetting the fall in signal amplitude.

However, there is a lower limit below which the limiting action ceases. This occurs when the signal across the tuned circuit falls below the voltage across the diode load capacitor, and the diode fails to conduct on the peaks of the signal. Below this signal amplitude, the driving valve behaves as a linear amplifier. The range over which limiting is maintained can be calculated from the graph of Fig. 11. The curve of $E=R_d i$, where R_d is the dynamic resistance of the tuned circuit, shows the relationship between the output voltage and the current supplied by the driving valve in the absence of the diode circuit. The curve of $E=R'i$ is the curve obtained when diode limiter is connected, R' being equal to R_d in parallel with R/2, the equivalent damping resistance due to the diode circuit (the diode rectification efficiency is assumed 100 per cent.). In plotting this graph, the input current is assumed to change its amplitude very slowly, so that at each stage the diode load capacitor charges to the peak value of the output voltage.

Under dynamic conditions, when the input signal amplitude decreases from the quiescent value i_0 the charge on the diode load capacitor does not have time to change appreciably, and the output voltage amplitude remains constant until the input current falls to the value i_1, when the diode ceases to conduct; the equivalent damping resistance due to the diode circuit is then infinite. If the input current decreases further, the output voltage falls linearly, following the curve $E=R_d i$. Thus the maximum "downward" modulation depth which the limiter will suppress is given by $(i_0-i_1)/i_0$. From the geometry of the figure, this modulation depth is given by $1-(i_1/i_0) = 1-(R'/R_d)$. But $R' = R_d(R/2)/(R_d + R/2)$, so that this maximum value is given by $R_d/(R_d + R/2)$. Thus for good "downward" modulation handling capacity, R_d must be large compared with R/2. This means that the tuned circuit is heavily damped under quiescent conditions, and hence the gain of the driver stage is low.

Because a diode employed in a practical circuit

* Wireless World (April, 1956)



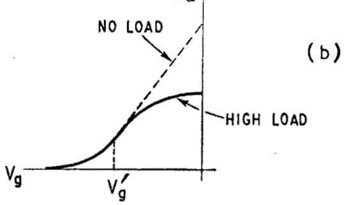


Fig. 9 (a) Showing the "bottoming" effect occurring with an anode limiter with a high-value anode load. (b) Modification of the I_a-V_g curve of the valve by the high-value anode load.

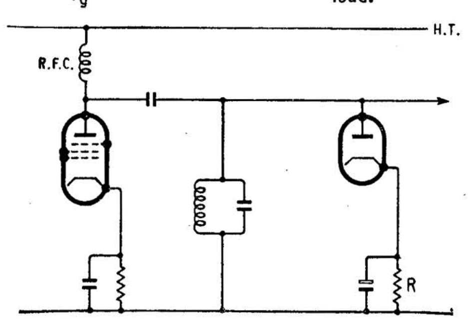


Fig. 10. Basic circuit of dynamic limiter.

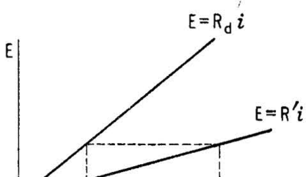


Fig. 11. Showing graphical construction for determining maximum "downward" modulation handling capacity of dynamic limiter.

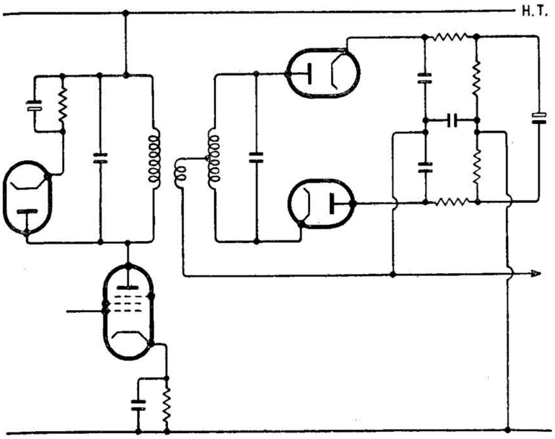
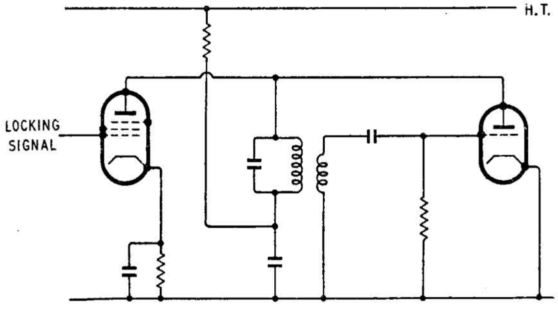


Fig. 12. Ratio detector with added dynamic limiter.

Fig. 13. Basic locked-oscillator circuit.



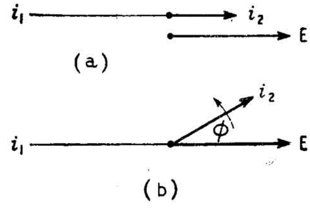
cannot have 100 per cent. rectification efficiency, the limiting action is not perfect, and the a.m. reduction factor is usually of the order of 10-20 dB. This type of limiter offers no protection against long-term changes of input signal amplitude or slow flutter due to reflections from aircraft, as the diode load capacitor will charge to the mean value of the input signal amplitude. It has, however, the advantage that the limiting action is maintained over a constant range of modulation depth at all levels of input signal amplitude, down to the quasi-threshold value at which the diode efficiency falls to the point where the limiting action is seriously impaired.

The damping imposed on the tuned circuit under operating conditions varies with the amplitude modulation of the input signal, and so, therefore, does the pass band of the tuned circuit.

The dynamic limiter may be employed with advantage in combination with those forms of limiter/discriminator where the a.m. suppression ratio falls somewhat short of the desirable value. In particular, it may be employed with a ratio detector, connected in parallel with the primary circuit, as shown in Fig. 12.

The Locked Oscillator. The type of oscillator used widely in receivers employs grid-current biasing, and in this resembles the grid limiter. The limiting action which takes place stabilizes the oscillation amplitude. Thus if an oscillator of this type can be made to change in frequency in step with an applied signal, the oscillator output amplitude will tend to remain constant independently of changes in the input signal amplitude. A suitable arrangement for locking an oscillator to an applied signal is shown in Fig. 13.

Fig. 14. Vector diagrams for circuit of of Fig. 13.



Consider firstly the conditions of operation when the applied signal frequency is equal to the free-running frequency of the oscillator. The equilibrium relationship between the vectors representing the oscillator valve anode current (i_1), the locking valve anode current (i_2) and the voltage across the tuned circuit (E) are as shown in Fig. 14(a). The current vectors are in anti-phase, and the oscillator valve anode current is also in anti-phase with the tuned circuit voltage, indicating that the oscillator valve is equivalent to a negative resistance in parallel with the tuned circuit. The locking valve current is in phase with the oscillator voltage, corresponding to a damping resistance of magnitude E/i_2. If now the locking signal frequency increases, the locking-valve anode current vector will commence to rotate in a clockwise direction, as shown in Fig. 14(b). There will now be a component of the locking signal current in quadrature with the oscillator voltage, i.e. the locking valve now behaves as a resistor and reactor in parallel. The reactance is equivalent to an inductive component under the conditions postulated, and this equivalent inductance re-tunes the oscillator, circuit to a higher frequency. Equilibrium is restored when the oscillator frequency is equal to that of the locking frequency, i.e. synchronization has been achieved. If the phase angle between the locking valve anode current and the oscillator voltage in this equilibrium condition is θ, then the reactance of the equivalent parallel inductance is $E/i_2\sin\theta$. The range over which locking occurs is limited because when the phase angle between the oscillator voltage and the locking valve anode current is equal to 90 degrees, the equivalent parallel inductance is at its minimum value; any increase of phase angle beyond 90 degrees increases the magnitude of the equivalent parallel inductance. At the minimum value of inductance, the reactance is given by E/i_2. For maximum effect, this reactance should be as small as possible, and hence for locking over a wide frequency range, E should be small, and i_2 large. Further, the main tuning inductance of the oscillator circuit should be as large as possible, for the equivalent parallel inductance to have maximum "pulling" effect.

The argument applied above can be applied also when the locking signal frequency is below that of the oscillator; in this case the locking valve behaves as an equivalent capacitor.

The relationship between the oscillator frequency and the locking frequency is as shown in Fig. 15(a). Outside the locking range, the oscillator frequency tends to swing between wide limits, as the heating effect with the locking signal produces alternately the effect of a parallel capacitance and parallel inductance; the mean frequency tends to follow the curves shown dotted in Fig. 15(a). In practice, however, non-linearity in the oscillator valve causes the oscillator to lock over a succession of small ranges at fractional multiples of the locking signal frequency (e.g. 10/9, 9/8), as shown in Fig. 15(b).

The greater the amplitude of the locking signal current, the greater is the frequency range over which locking is maintained. When the input signal is amplitude modulated, therefore, the minimum value to which the input current falls must be sufficient to lock the oscillator over the full frequency range of the f.m. signal. The locked oscillator thus resembles the grid limiter, in that there is a threshold limit of signal which must be exceeded for satisfactory performance. The "downward" a.m. modulation handling capacity is determined by the ratio of the mean input signal amplitude to this threshold amplitude.

The locked oscillator limiter may also suffer from the effects of the time constant of the grid-biasing components in the same way as the grid limiter, and the grid components must thus be chosen with care.

Examples of the locked oscillator limiter were given in Part 4 of this series. This type of limiter is not now widely used. One of its principal disadvantages is the feedback of the oscillator signal to early i.f. stages, which can lead to overloading in the i.f. amplifier. A number of schemes have been proposed for locking an oscillator at a sub-multiple of the intermediate frequency to avoid this effect. Additionally, a single stage locked oscillator limiter-discriminator has been described by Bradley (see references).

The "Clipper." This type of limiter employs two triodes cathode coupled as shown in Fig. 16. It thus comprises a cathode follower driving an earthed-grid triode. Under quiescent conditions, both triodes are conducting, and each is biased approximately to the mid-point of its grid base. If now the input signal at the grid of V1 increases positively, the anode current of V1 increases. The cathode potential therefore rises, and this in turn leads to a decrease of anode current in V2. In fact,

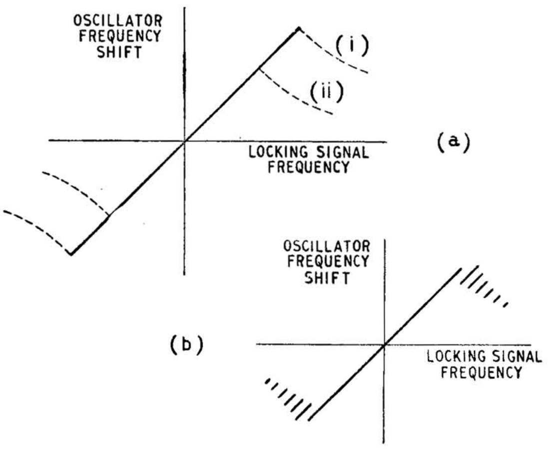


Fig. 15. (a) Locking of oscillator frequency to that of locking signal. The dotted portions of the curves indicate the mean value to which the oscillator frequency tends outside the locking range. Curve (i) is that for a larger locking signal input than curve (ii); (b) Showing locking of the oscillator frequency at fractional multiples of the locking signal frequency over restricted ranges outside the true locking range.

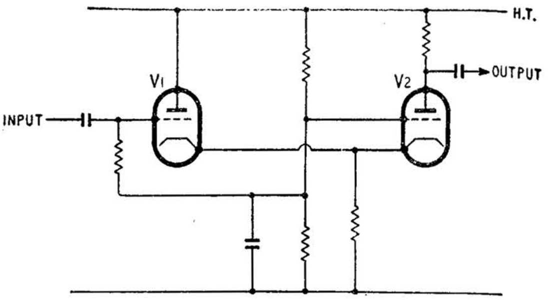


Fig. 16. Basic circuit of the "clipper" type of limiter.

the common cathode potential rises by an amount equal to half the increase of V1 grid potential. If the input to the grid of V1 increases sufficiently, the common cathode potential rise to the point where anode current is cut off in V2. The degree of feedback applied to V1 then increase sharply, and V1 behaves as a true cathode follower. During the rise of common cathode potential the anode of V2 rises to h.t. potential, and when anode current is cut off, it remains at this value, i.e. the output signal is then limited.

When the input signal at the grid of V1 is negative-going, the common cathode potential falls, and the anode current in V2 rises. Whilst both valves are conducting, the change in cathode potential is approximately half that of V1 grid. If the input signal rise to a value sufficiently negative, anode current in V1 is cut off, and there is no further change in common cathode potential. Thus this circuit "clips" the input signal symmetrically, providing a chain of square-wave output pulses.

This form of limiter is very useful at low frequencies, and may be used with input signal frequencies of the order of a few Mc/s. It has a fixed threshold of the order of 5-10 volts above which limiting occurs. As it does not depend for its action upon the charging and discharging of capacitors, it is free from "blocking" effects.

The author wishes to record his thanks to Dr. G. J. Phillips and Mr. J. G. Spencer for helpful discussion.

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