WS No. 19 Mark III

FREQUENCY MODULATION IN V.H.F. FIELD RADIO SETS

Note: This Issue 3, Pages 1 - 43 and 1001 - 1003 supersedes Pages 1 - 6, Issue 1; 7 - 8, Issue 2; 9 - 14, Issue 1; 15 - 16(a), Issue 2; 17 - 18, Issue 1; 19 - 20, Issue 2 and 21 - 27, Issue 1, of various dates. The title has been changed and the revised regulation is based on the Australian EMEI.

SUBJECT INDEX

Distribution - Class 1175. Code No 3

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INTRODUCTION

1. Modulation is the process by which a radio frequency carrier wave is made to carry intelligence. The intelligence usually consists of an audio frequency signal (as will be assumed throughout this regulation).

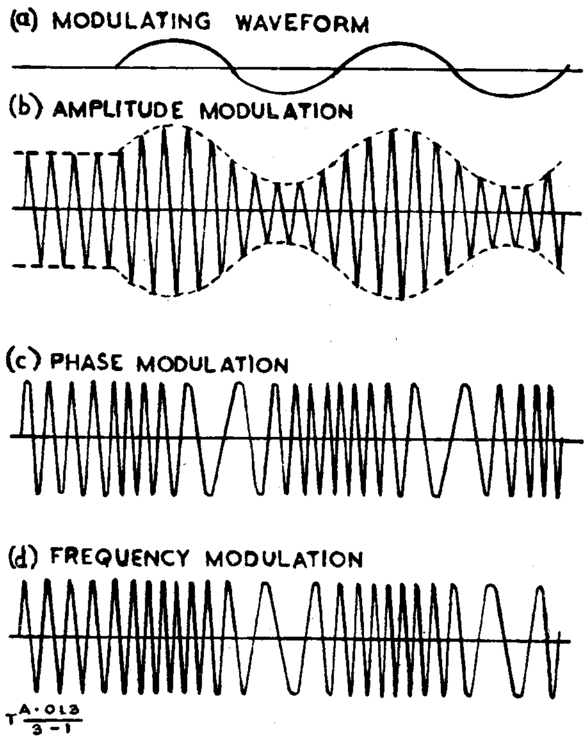

Fig 1 - Modulation waveforms - sinewave

2. A carrier wave has three basic characteristics ie amplitude, frequency and phase, any one of which may be varied by the intelligence; resulting in aplitude, frequency or phase modulation respectively. Fig 1 shows this in graphical form for the case where the modulating signal is a sinewave. Note the similarity between frequency modulation (f.m.) and phase modulation (p.m.). At the receiver it would be impossible to distinguish between the two in this particular instance, and in all cases as will be shown below (para 5 to 10) there is a simple relationship between the two. That there is a difference between f.m. and p.m. when a complex modulating waveform is used is evident from Fig 2.

3. In field radio sets a.m. is generally used for high frequency (h.f.)

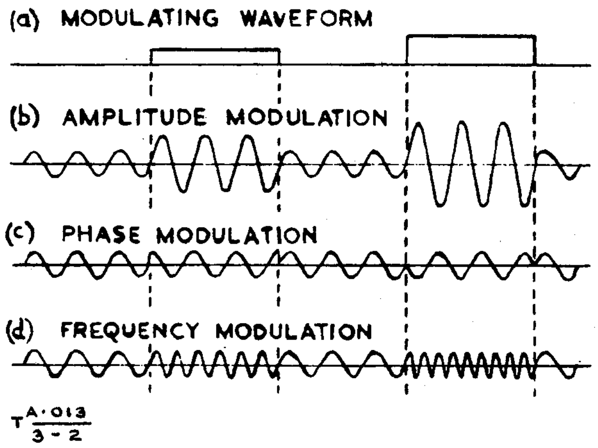

Fig 2 – Modulation waveforms – squarewave

and f.m. for very high frequency (v.h.f.) communications, although there are some exceptions to this rule. P.M. is not employed because of circuit complication.

4. The fundamental differences between a.m. and f.m. are shown in Table 1. A more detailed comparison will be given in para 11 to 40.

A.M.	F.M.
Frequency is constant	Amplitude is constant
Extent of carrier amplitude change is proportional to the amplitude of the audio intelligence	Extent of carrier frequency change is proportional to the amplitude of the audio intelligence
The rate at which the carrier amplitude varies above and below its mean value is the audio intelligence frequency	The rate at which the carrier frequency varies above and below its mean value is the audio intelligence frequency

Table 1 – Fundamental differences between a.m. and f.m.

Difference between p.m. and f.m.

5. The difference between p.m. and f.m. is best explained by a brief mathematical analysis. If $e = A \sin \omega t$ is the equation for an unmodulated carrier wave, then clearly

$$e = A \sin (\omega t + \delta\phi \sin pt) \qquad (1)$$

(where $\sin pt$ is an audio frequency wave and $\delta\phi$ is a constant) represents the same carrier wave after being phase modulated at the audio rate between the limits of $\pm\delta\phi$ radians.

6. If a carrier wave is being frequency modulated between the limits $\pm M_f$ (ie M_f = maximum frequency deviation) then

the instantaneous frequency $= f + M_f \sin pt$

the instantaneous angular frequency $= \omega + 2\pi M_f \sin pt$

and by integration the instantaneous angle $= \omega t - \dfrac{2\pi M_f}{p} \cos pt$

7. Hence the equation for an f.m. wave is:-

$$e = A \sin (\omega t - \frac{2\pi M_f}{p} \cos pt) \qquad (2)$$

8. By comparison of equations (1) and (2) it will be seen that in p.m. the extent of the phase shift of the carrier is independent of the frequency of the modulating voltage whereas in f.m. the extent of the phase shift is inversely proportional to the frequency of the modulating voltage.

9. Expressing the same comparison in a different way we see that if the maximum phase shift of a transmission is fixed then the extent of the equivalent f.m. produced is directly proportional to the modulating frequency, ie if a p.m. transmission were received on an f.m. receiver the higher audio notes would be heard at a stronger level than the lower ones. This fact is very important and reference will be made to it at a later stage in this regulation.

10. Clearly a system which would normally produce p.m. will produce f.m. if a corrective network is introduced. The converse is also true. Fig 3 shows an f.m. output being produced by a phase modulator. The network shown provides an output potential, the amplitude of which varies inversely with the frequency of the input potential for all frequencies above 30c/s.

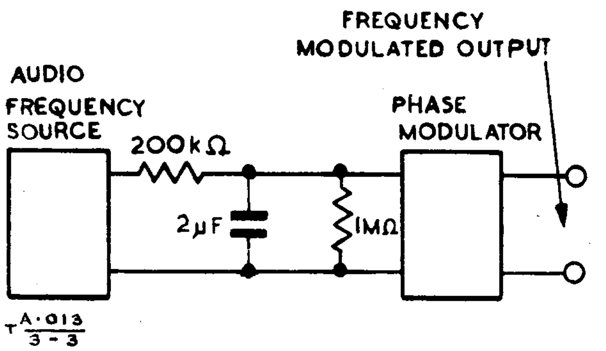

Fig 3 - Frequency modulated output from a phase modulator

Detailed comparison between a.m. and f.m. - modulation percentage

11. In a.m. the amplitude of a carrier modulated 100% varies from zero to twice the value of the unmodulated carrier. A strict analogy would require that in a 100% modulated f.m. transmission the frequency be varied from zero to twice the nominal carrier frequency. This is obviously impracticable and an arbitrary limit (±15kc/s in field radio equipments; ±75kc/s in commercial broadcasting practice) must be set to the permissible frequency variation. Maximum deviation from the centre frequency will occur whenever the audio modulating signal is at the maximum amplitude that the transmitter is expected to handle. Thus whenever maximum deviation occurs we can loosely say that the transmission is 100% modulated. On this basis percentage modulation is sometimes stated as being

$$\frac{\text{actual deviation}}{\text{maximum deviation}} \times 100$$

SIDEBANDS AND BANDWIDTH CONSIDERATIONS

Amplitude modulation

12. A sinusoidally modulated a.m. signal may be shown to consist of three components; the original carrier frequency, and an upper and a lower sideband

component. The effects of changes of modulation frequency and modulation percentage are shown in Fig 4. Note that the amplitude of the carrier component is constant but sideband amplitude is proportional to the modulation percentage. The frequency separation between the carrier and each sideband is equal to the frequency of the modulation source. The bandwidth requirement is thus twice the frequency of the highest audio note it is desired to transmit. A range of frequencies from 50c/s to 3,000c/s is sufficient to reproduce speech intelligibly and hence the necessary bandwidth for a.m. field radio equipment is 6kc/s.

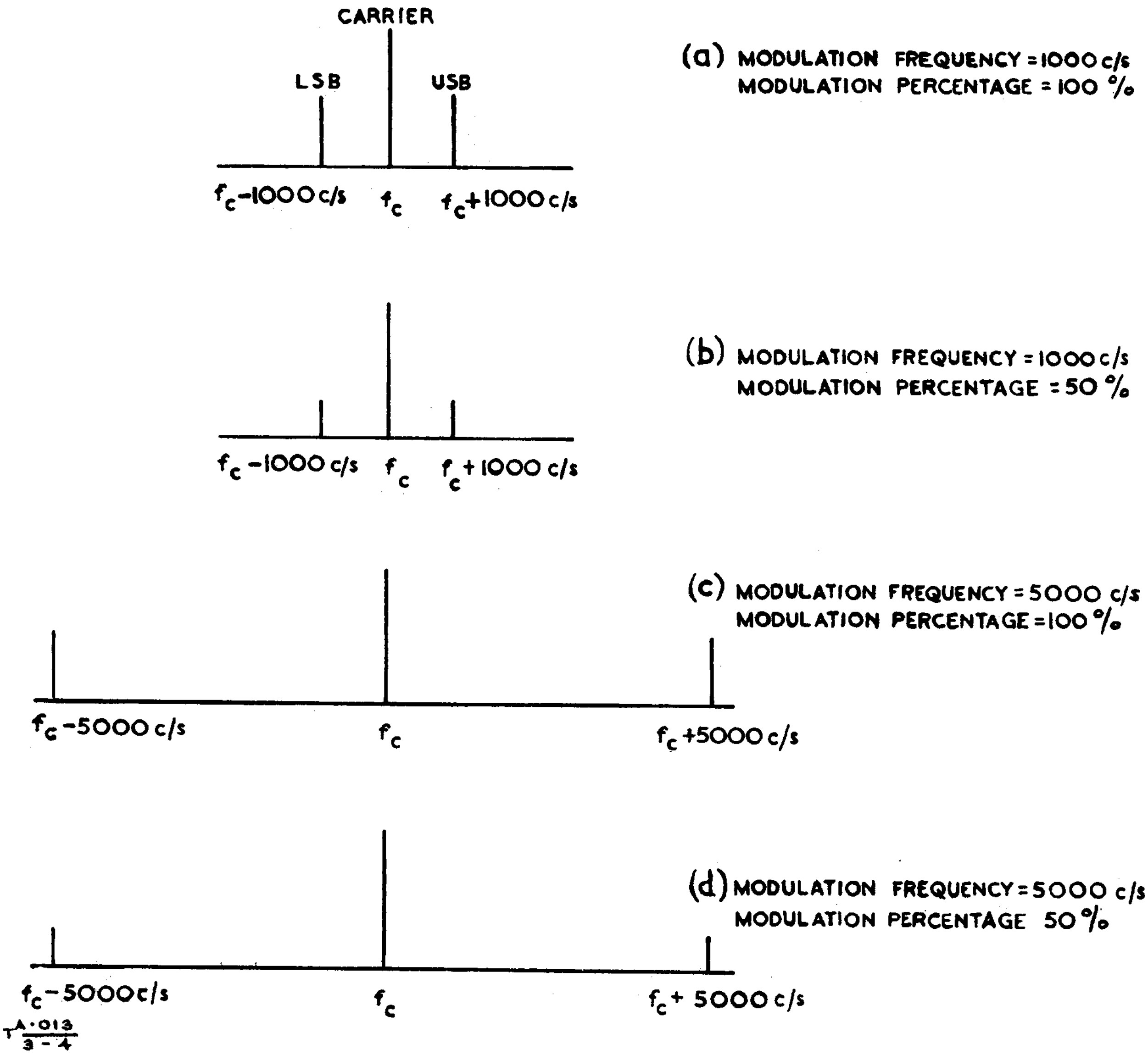

Fig 4 - Sideband distribution for sinusoidally modulated a.m. wave

Frequency modulation

13. For f.m. the situation is more complex and it should be clearly understood at the outset that just because the maximum deviation of a transmission is say 15kc/s this does not mean that the bandwidth is 30kc/s. Critical examination of an f.m.

waveform ((d) in Fig 1) shows that no complete r.f. cycle is a pure sinewave (the time for any positive half cycle is always slightly different from that for the following half cycle) so that it is only to be expected that a Fourier analysis of such a waveform would lead to a very complex result.

14. Mathematical expansion of the equation for a sinusoidally modulated f.m. wave (equation (2) para 7) involves Bessel functions and is beyond the scope of this regulation. The results however are important and show that an f.m. wave has theoretically an infinite number of sidebands symmetrically placed about the centre frequency, the separation between each sideband being equal to the modulation frequency. Typical sideband distributions are shown in Fig 5. Fortunately the sidebands remote from the carrier have negligible amplitude. For bandwidth considerations, only significant sidebands (ie those of greater amplitude than 1% of the unmodulated carrier) need be considered since loss of the others does not cause any reduction in the quality of transmission.

15. If the amplitude of the unmodulated carrier is taken as unity then the amplitude of the centre frequency component of the f.m. wave is given by the Bessel function $J_0(\beta)$ a graph of which is shown in Fig 6. The amplitude of each of the first pair of sidebands is given by $J_1(\beta)$, that of the second pair of sidebands by $J_2(\beta)$ etc where β is the $\frac{2\pi M_f}{p}$ term of equation (2) para 7. That is

$$\beta = \frac{2\pi M_f}{p} = \frac{M_f}{f_a} = \frac{\text{frequency deviation}}{\text{modulation frequency}}$$

and is known as the modulation index.

16. As an example, assume that a modulation signal has a frequency of 2kc/s and sufficient strength to cause a frequency deviation of 8kc/s. Then $\beta = 4$ and we see from Fig 6 that

(a) the amplitude of the centre frequency component = 0.397

(b) the amplitude of each of the first pair of sidebands = 0.066

(c) the amplitude of each of the second pair of sidebands = 0.364

(d) the amplitude of each of the third pair of sidebands = 0.430

17. Tables of Bessel functions must be used if the other sideband amplitudes are desired. Negative signs are of no consequence and have been ignored. Note that for certain values of β (eg $\beta = 2.4$) the centre frequency component has zero amplitude. The particular value of β which applies when the signal is modulated to the maximum permissible extent by the highest audio frequency which it is desired to transmit is known as the deviation ratio, ie

$$\text{deviation ration} = \frac{\text{maximum frequency deviation used}}{\text{highest audio frequency it is desired to transmit.}}$$

Thus in field radio equipments the deviation ratio is $\frac{15\text{kc/s}}{3\text{kc/s}} = 5$

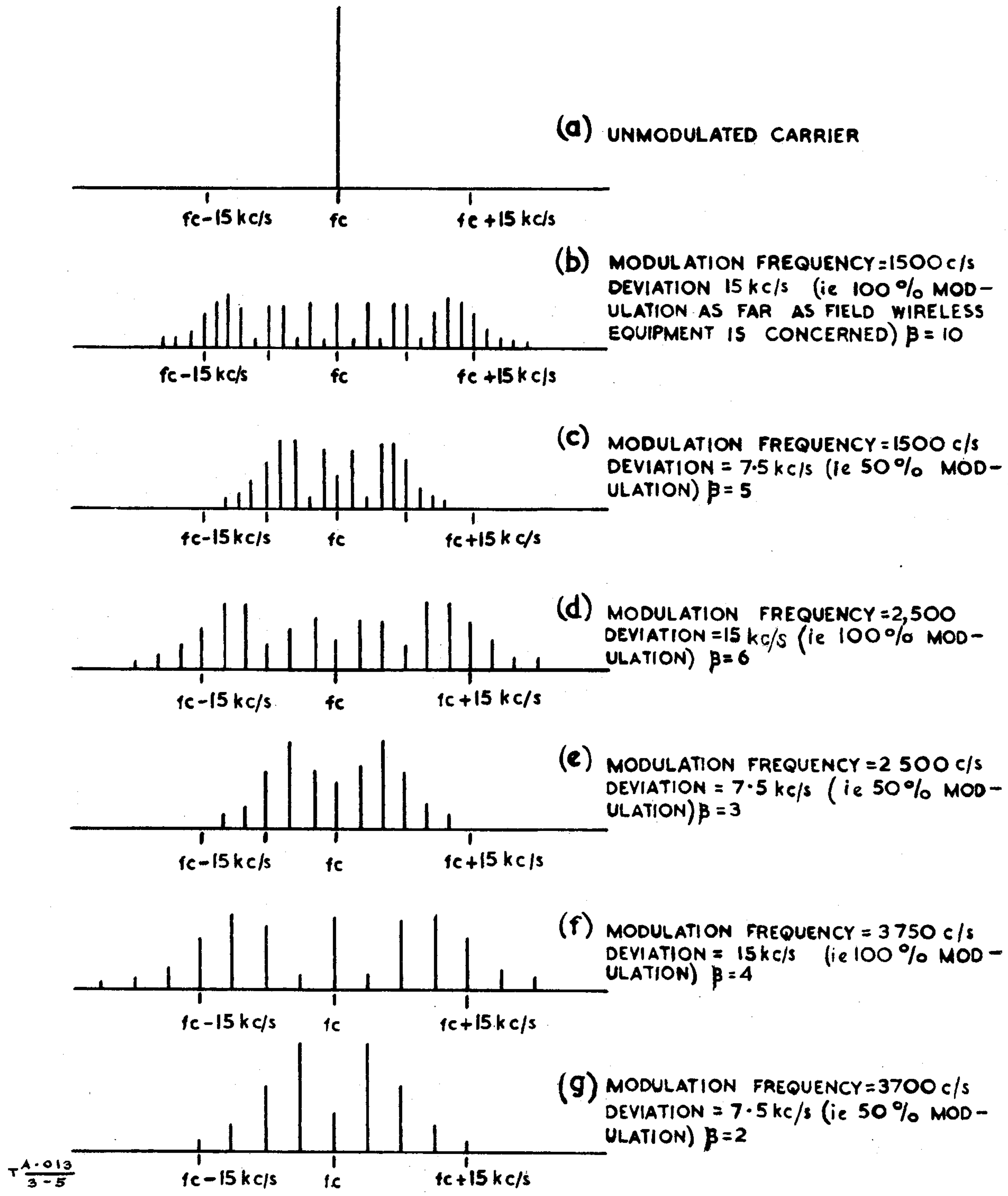

Fig 5 - Sideband distribution for sinusoidally modulated f.m. wave

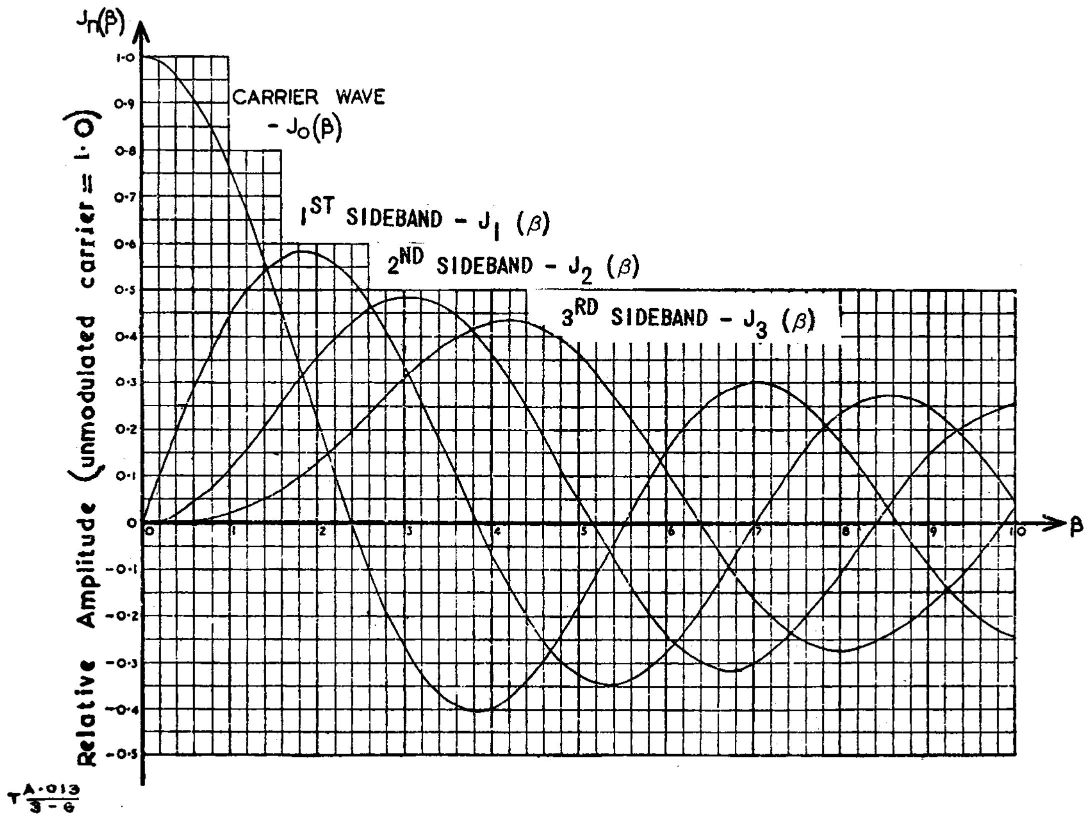

Fig 6 – Relative amplitudes of components of an f.m. wave

18. Table 2 provides the data for determining the bandwidth required by an f.m. transmission. As can be seen from Table 3, a field radio transmission with M_f = 5kc/s and deviation ratio = 5 requires a theoretical bandwidth of 48kc/s (assuming sinusoidal modulation).

19. When the instantaneous frequency is varied in a more complex manner than that corresponding to sinusoidal modulation, the frequency spectrum becomes very complicated. Thus where there are two modulating frequencies the side frequencies present include not only those that would be obtained with each modulation acting separately, but also combination frequencies. Although complex modulation greatly increases the number of frequency components present in an f.m. wave, it does not widen the frequency band occupied by the energy of the wave. This is because the modulation index of each modulating component will in general be reduced when the total modulation is divided between two or more components. In fact, when the maximum frequency deviation produced by a complex modulation is fixed, the energy of the wave tends, if anything, to be concentrated in a narrower band than with a simple sinusoidal modulation having the same frequency deviation. The bandwidth figure given in para 18 may therefore be reduced in practice. A suitable rule of thumb for calculating the bandwidth of an f.m. system is: Bandwidth = $2(M_f + f_a)$, (ie add the maximum deviation to the highest modulation frequency and double the result). For field radio equipments the necessary bandwidth is thus approximately 36kc/s. Note that to provide an adequate safety margin to cover deterioration of performance while the set is in use, the alignment instructions for any given radio set will usually specify a much wider bandwidth than is here indicated.

Theoretical figures applying to any f.m. system		
Modulation index	No of pairs of significant sidebands	Bandwidth as a multiple of M_f
β	N	$\frac{2N}{\beta}$
24	29	2.42
21	26	2.48
18	23	2.56
15	19	2.54
12	16	2.67
10	14	2.80
9	13	2.89
8	12	3.00
7	11	3.14
6	9	3.00
5	8	3.20
4	7	3.50
3	6	4.00
2	4	4.00
1	3	6.00

Table 2 - Data for f.m. bandwidth calculations

Application of the theory to field radio equipments with $M_f = 15$kc/s	
Modulation frequency	Bandwidth necessary to include all significant sidebands
$\frac{M_f}{\beta}$ c/s	$\frac{2NM_f}{\beta}$ kc/s
625	36.3
714	37.2
833	38.3
1000	38.0
1250	40.0
1500	42.0
1667	43.4
1875	45.0
2143	47.1
2500	45.0
3000	48.0
3750	52.5
5000	60.0
7500	60.0
15000	90.0

Table 3 - Bandwidth calculations for deviation of 15kc/s

Note the dependence of bandwith on modulation depth

Application of the theory to cases where deviation is limited to 7.5kc/s (ie representing 50% modulation in a field radio transmission)	
Modulation frequency	Bandwidth necessary to include all significant sidebands
$\frac{M_f}{\beta}$ c/s	$\frac{2NM_f}{\beta}$ kc/s
312	18.1
357	18.6
416	19.2
500	19.0
625	20.0
750	21.0
833	21.7
938	22.5
1071	23.6
1250	22.5
1500	24.0
1875	26.2
2500	30.0
3750	30.0
7500	45.0

Table 4 - Bandwidth calculations for deviation of 7.5kc/s

Response to noise

20. The f.m. receiver is less responsive to noise than the a.m. receiver; in fact, the signal on f.m. appears to suppress the noise, whereas in a.m. the signal is heard above a background of noise. Thus, speech which is just readable with a.m. becomes readily so with f.m., although the signal-to-noise ratio at the aerial is the same on each system. This is specially noticeable when the source of the noise is static or ignition interference. The background noise heard on f.m. receivers is of a higher pitch than on a.m. This is a further advantage since for the same intensity of noise less loss of intelligibility is caused.

21. For any given set of conditions it is possible to calculate a 'noise improvement factor' which denotes the extent of signal-to-noise ratio improvement obtainable by the use of f.m. in place of a.m. It can be shown that an f.m. system with a large deviation (eg 75kc/s as used in commercial broadcasting practice) has a higher noise improvement factor than a system with a smaller deviation. On the other hand if maximum range is desired, and a relatively low signal-to-noise ratio (ie low compared with a high fidelity f.m. transmission but still high compared with a.m.) is tolerable, then a small frequency deviation is best. Hence the choice of 15kc/s deviation for communication purposes where readability is not seriously impaired by a moderate amount of noise, and range is very important.

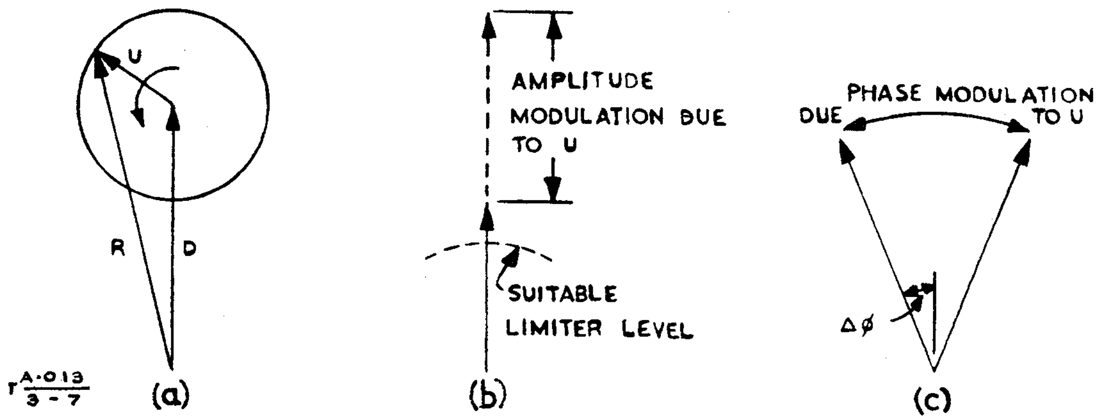

Fig 7 - Production of extraneous a.m. and p.m. by a noise frequency component

22. Calculations of noise improvement factor are beyond the scope of this regulation but it is important that the reasons for the improvement are understood. Section (a) of Fig 7 shows the vector diagram of a desired signal D, an undesired noise frequency component U, and the resultant composite signal R. The page is assumed to be rotating at the frequency of the desired signal so that D appears stationary and U rotates at the difference frequency. Sections (b) and (c) of Fig 7, which are self-explanatory show the extent of the extraneous a.m. and p.m. produced by the presence of U. The modulation frequency is obviously the frequency difference between D and U. In an f.m. receiver the extraneous a.m. can be removed by a limiter stage (see para 66). The p.m. however will produce an output, the amplitude and frequency of which will be directly proportional to the frequency difference between D and U (see para 9).

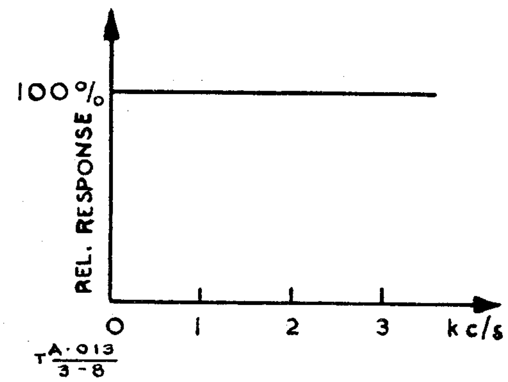

Fig 8 – Noise response of an a.m. receiver

23. Now comparing a.m. and f.m. systems for the practical case where noise frequency components can be assumed to be evenly distributed over the bandwidth of the channel used, we see that in the a.m. receiver the noise output is evenly distributed over the audio range (Fig 8) whereas in the f.m. receiver the noise output has a rising response (Fig 9). In the a.m. case the i.f. bandwidth has been assumed to be 6kc/s so that no noise outputs of higher frequency than 3kc/s are produced. In the f.m. case the i.f. bandwidth has been assumed to be 50kc/s so that noise outputs up to 25kc/s are produced.

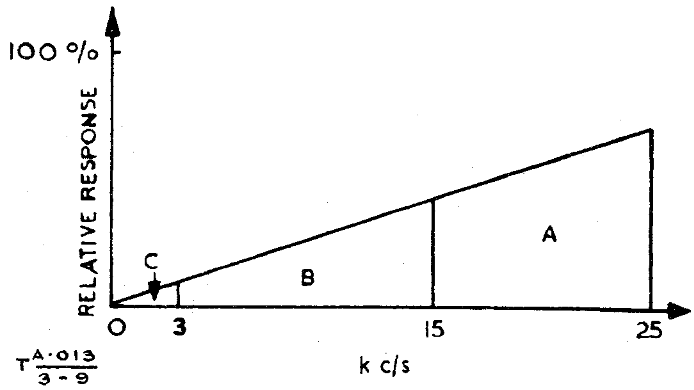

Fig 9 – Noise response of an f.m. receiver

24. For the region marked A in Fig 9 the noise output is greatest but note that the frequencies concerned are inaudible. The region marked B is audible but is outside the range of audio frequencies required in a field radio transmission, hence, if desired, this range of noise frequencies can be eliminated without loss of intelligibility, by filters. The small amount of noise in region C covers the same range of frequencies as the intelligence and thus, at first sight, appears to represent an irreducible minimum. However a further improvement is made possible by the use of pre-emphasis at the transmitter and de-emphasis at the receiver as explained in the next paragraph.

25. The higher frequency components of the intelligence are usually very weak compared with the lower frequency components. The amplitudes of the high frequency components can thus be increased at the transmitter prior to modulation without there being any danger of exceeding the permissible deviation limits. This is called pre-emphasis. De-emphasis of the high frequencies will be required at the receiver in order to restore tonal balance. During de-emphasis the higher intelligence frequencies are reduced to their normal relative amplitude and at the same time the higher noise frequencies suffer a reduction in amplitude. Since the noise was not included in the pre-emphasis process the overall result is an improvement in signal-to-noise ratio. Typical pre-emphasis and de-emphasis curves for a field radio set are shown in Fig 10.

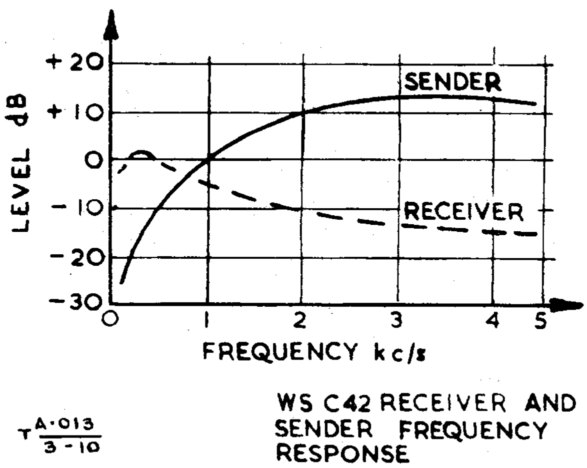

Fig 10 – Typical curves of pre-emphasis and de-emphasis

26. It is possible to improve the signal-to-noise ratio of an a.m. system by the use of pre-emphasis and de-emphasis. The improvement however is not so great as that obtained in an f.m. system, due to the fact that noise in an a.m. receiver is not confined to the high frequency end of the audio spectrum.

Response to interfering transmissions

27. If U in Fig 7 is now considered as representing an undesired carrier it will be appreciated that an f.m. system is also less responsive to interfering transmissions that an a.m. system.

28. In f.m. if two signals are modulated and occupy the same channel then providing one has more than three times (approximately) the signal strength of the other, the stronger signal is heard to the total exclusion of the weaker. This is known as the 'capture effect' of f.m., the stronger signal being said to have 'captured' the receiver.

29. Two f.m. transmitters working on the same frequency will cause mutual interference only in a small transition region midway between the two stations. Outside this transition region reception of the nearer station will be interference free. In any large area it is therefore possible to operate several f.m. stations on the same channel thereby offsetting the disadvantage of the large bandwidth required by an f.m. transmission.

Multipath reception

30. Because v.h.f. waves are reflected from buildings and similar obstructions, the radiated signal can arrive at the receiving antenna over multiple paths from the same transmitter. The different paths are usually not the same length, with the result that the multipath signals take different amounts of time to travel the separate transmission paths to the receiver. With multipath reception of signals slightly displaced in time with respect to one another, the distortion that can result is more objectionable in f.m. than in a.m. because the frequency of the f.m. signal is continuously changing. Since the instantaneous frequency of an f.m. signal varies with time, the multipath signals at the receiver generally will have different frequencies at any instant.

Frequency of working

31. The use of f.m. is in general restircted to frequencies above 30Mc/s because of the large bandwidth required (see also para 33). There is no such restrictuion upon the use of a.m.

Range of working

32. The transmission range of an f.m. system is small (30 miles being a typical figure) since v.h.f. waves are not reflected by the ionized layers of the earth's upper atmosphere. A.M. transmission range is similarly limited if v.h.f. waves are used but, by the use of h.f. waves, world-wide communication by a.m. is possible.

33. It is important to note that even if the bandwidth objections stated in para 31 are ignored, world-wide communication by use of frequency modulated h.f. waves is not practicable because selective fading due to multipath reception results in an almost unintelligible signal at the receiver, ie even in the rare instances where frequency modulation of h.f. waves is used, reception must still be confined to the direct ray from the transmitter.

Transmitter efficiency

34. F.M. enables an increased sender efficiency to be obtained, so that for a given radiated power, the d.c. input power required is considerably reduced. This is a decided advantage when the input power is to be derived from portable batteries. In the case of a.m., a sender power amplifier must be capable of dealing with a peak signal of amplitude twice that of the unmodulated carrier, but, as the average depth of modulation is approximately 25%, the stage is normally under-run. An f.m. signal has constant amplitude and hence the power amplifier can be fully loaded and will, therefore, be more efficient. Note also a further gain in efficiency due to the fact that the power amplifier need no longer operate under linear conditions.

35. High level modulation is often used in the a.m. transmitter to modulate the r.f. power amplifier and this requires a great deal of audio power. Low level modulation can be used to modulate a low power r.f. stage, but this requires that all succeeding stages be linear amplifiers to avoid distortion, thus resulting in a loss in efficiency as compared to class C amplification. Regardless of whether high or low level modulation is used, the h.t. supply works under varying load conditions and must therefore be specially designed to have good regulation.

36. In the f.m. transmitter, efficient modulation is carried out at the lowest possible level, ie at the oscillator itself. Negligible audio power is required and the h.t. supply works under constant load conditions.

Receiver complexity

37. The f.m. receiver is more complex than the a.m. equivalent. The a.m. receiver, for example needs no limiter and the a.m. detector is far more simple than its counterpart the f.m. discriminator.

38. The gain per i.f. stage in an f.m. receiver is comparatively low because of the large bandwidth employed. The total r.f. and i.f. gain must be high since, for successful limiting, the limiter requires a large input signal. This means that the total number of r.f. and i.f. stages in an f.m. receiver will exceed that of an a.m. receiver.

39. F.M. receivers usually have a.f.c. and a tuning indicator to offset the fact that tuning-in an f.m. signal is more difficult and the distortion due to mistuning is more severe than for a.m. In transceivers this problem is usually solved differently. A.F.C. is applied to the transmitter to minimize frequency drift and the receiver is either directly crystal controlled or manually tuned to a channel by reference to crystal check points.

Fidelity

40. The superiority of f.m. for high fidelity transmission is not made use of in field radio equipments. In both a.m. and f.m. field radio sets the highest audio frequency reaching the receiver headset without severe attenuation is approx 3kc/s. Full intelligibility is retained except that, as in normal telephone usage, there may be occasional confusion between the sibilant sounds such as 's' and 'f' which include basic frequencies of the order of 4kc/s. In these rare instances reference to the context of the message or use of the phonetic alphabet clarifies the situation.

41. A feature which is essential for high fidelity is wide dynamic range. A.M. is limited to a volume range corresponding to approximately 5 to 90% modulation, representing an audio power variation of about 320 to 1. The upper modulation limit being set by distortion considerations and the lower limit by the signal-to-noise ratio problem. F.M. is not so restricted and a dynamic range 100 times that of an a.m. transmission is easily obtainable. When it is considered that there is an audio power variation of about 1,000,000 to 1 between very loud speech and a whisper the superiority of f.m. for high fidelity transmission is apparent. However this potentially superior dynamic range is not made use of in field radio equipments. In fact correct voice procedure dictates that only a small volume range be used and this range is still further restricted in some f.m. transmitters by the use of a.m.c. and automatic deviation control. The desired result in the field being that all messages whether spoken softly or loudly should be received loud and clear.

METHODS OF PRODUCING FREQUENCY MODULATED WAVES

General

42. There are many ways of producing frequency modulated waves. Both of the methods used in current field radio sets rely on varying the resonant frequency of the master oscillator circuit in sympathy with the modulating voltage. In the older method a valve is made to act as a variable reactance and forms part of the resonant circuit whereas in the new range of v.h.f. radio equipments ferrite reactors are used for the same purpose. Centre-frequency stability is a problem in both cases and a highly efficient form of a.f.c. is essential.

43. It should not be thought that frequency instability is necessarily inherent in an f.m. system. One of the methods used in commercial practice, ie the Armstrong system, has the distinct advantage that the master oscillator can be crystal-controlled. This system makes use of the fact that if the carrier is removed from an a.m. wave, phase-shifted through 90° and then reinserted, the resulting wave is phase modulated. Fig 11 and 12 illustrate the theory (in both figures the page is assumed to be rotating at carrier frequency so that the carrier appears stationary while the lower and upper sidebands rotate clockwise and anticlockwise respectively at the modulation frequency). Unfortunately only very small phase deviations are permissible in this system if serious distortion is to be avoided. Hence the modulation process must be carried out at low frequency (order of 200kc/s) and a large number of stages of frequency multiplication used to obtained a practical frequency deviation figure. See para 10 for conversion of the resulting p.m. to f.m. (AN/TRC-3 and 4 use this method).

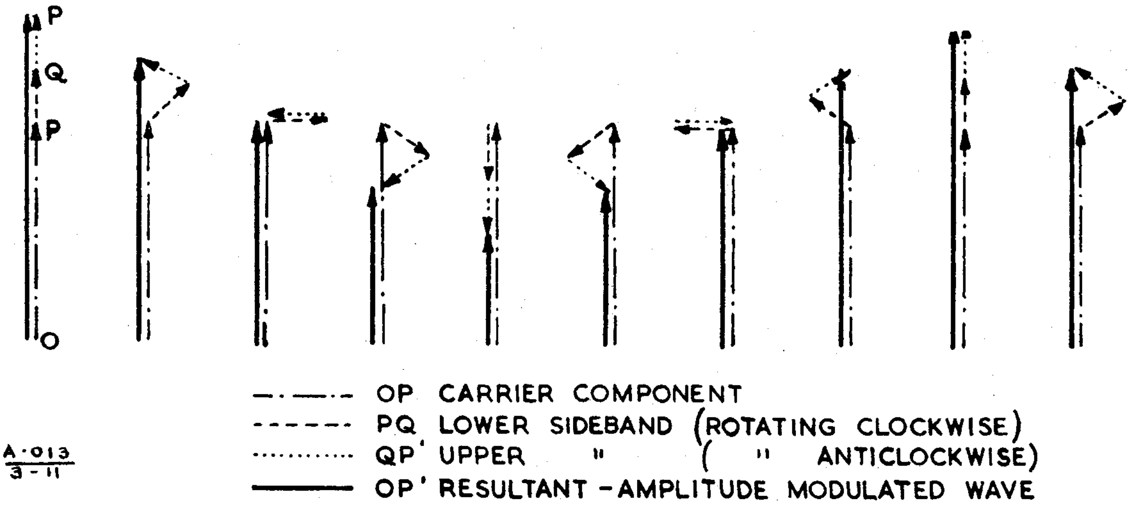

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Fig 11 - Vector relationship between carrier and sidebands in an a.m. wave

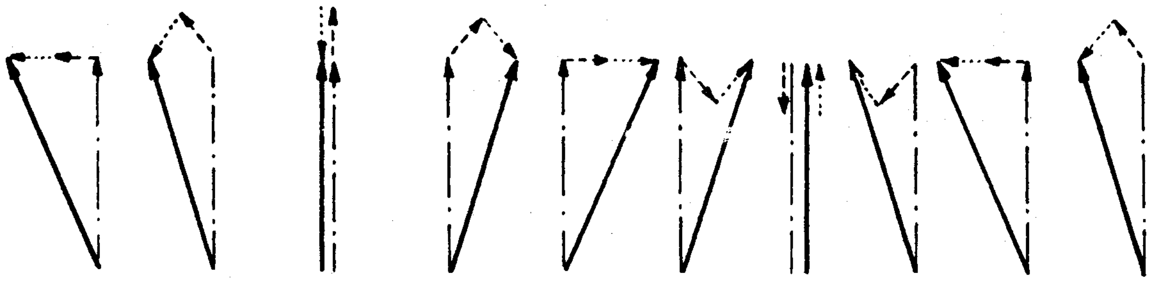

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Fig 12 - Vector relationship with carrier component rotated by 90°

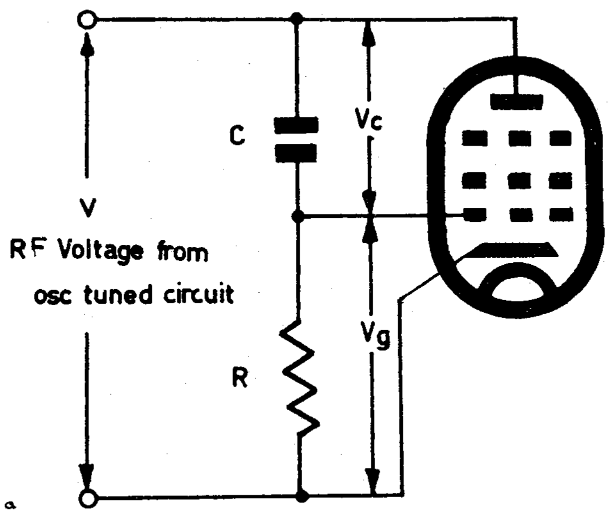

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Fig 13a - Simplified reactance valve circuit

Reactance valve modulator

44. The reactance valve modulator consists of an oscillator with a second valve (the variable reactance valve, which is usually a pentode) connected across the tuned circuit and operated in such a manner as to look like a reactance of magnitude dependent on the modulation voltage applied to its grid. With no modulation voltage applied, the circuit is tuned to the nominal carrier frequency. The variation of the apparent reactance of the valve due to the modulation voltage will thus alter the frequency of the oscillator above and below the carrier frequency, at a rate dependent on the frequency of the audio modulating voltage.

45. The pentode is made to act as a reactance by simply arranging that its control grid is fed with r.f. energy from the same source as the anode but via a 90° phase-shift circuit. Since the anode current of a pentode is almost entirely dependent upon the grid voltage and virtually independent of the anode voltage it follows that the anode current remains in phase with the grid voltage. Anode current and voltage are thus 90° out of phase and the valve is therefore behaving as a reactance. In practice the phase shift employed may not be exactly 90° and the apparent impedance of the valve will then have a resistive component. (In some field radio sets the reactance valve grid is fed via phase-shift circuits from the oscillator grid but the basic principle remains the same in that the voltage fed to the reactance valve grid is still arranged to differ by approximately 90° from that which appears at the anode).

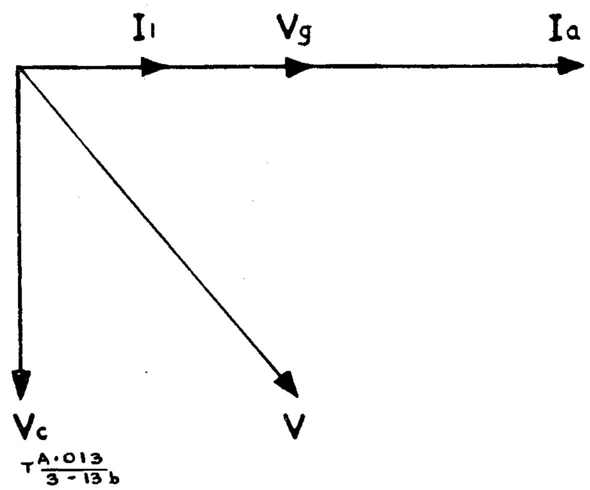

Fig 13b - Simplified reactance valve vector diagram

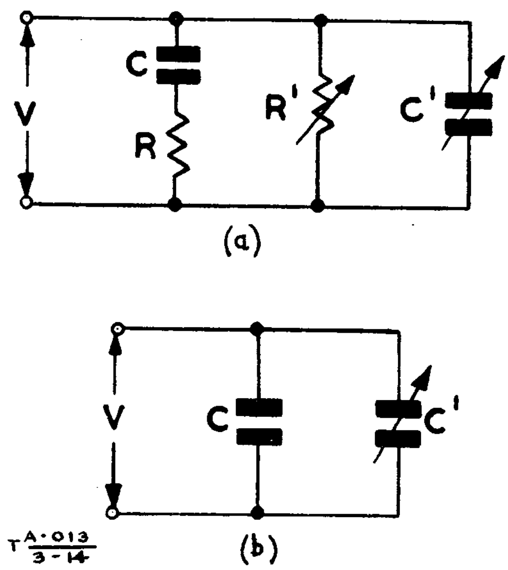

Fig 14 - Reactance valve equivalent circuit

46. One of many possible reactance valve circuits is shown in Fig 13a. For simplicity the d.c. supply and bias arrangements have not been shown and all currents and voltages referred to are r.f. only.

47. In the associated vector diagram (Fig 13b) I_1 is the reference vector. V_g and V_c are the voltage drops across R and C respectively, V_g being in phase with I_1 while V_c lags by 90°. The vector sum of V_g and V_c must obviously equal V. V_g controls the anode current, the actual value of which depends upon the g_m of the

valve. It is clear from the vector diagram that the r.f. voltage V across the valve is accompanied by a leading r.f. current I_a flowing through it. The valve can therefore be regarded as a capacitive impedance and is for convenience represented on the equivalent circuit ((a) in Fig 14) as a variable capacitor C^1 and a variable resistor R^1 in parallel. Note that if the reactance of C is very large compared with R, two simplifications can be made to the equivalent circuit. The first simplification is the obvious deletion of R. Secondly R^1 can be deleted since the angle between V and I_a on the vector diagram is now very close to 90°. The resulting approximate equivalent circuit is shown at (b) in Fig 14.

48. (a) Referring to the circuit in Fig 13a, let Y^1 be the admittance which the valve presents to the r.f. source voltage V.

Then $Y^1 = \frac{I_a}{V}$ and since $I_a = g_m V_g = g_m R I_1$ and $V = \left(\frac{1}{j\omega C} + R\right) I_1$

$$= \frac{g_m R}{\frac{1}{j\omega C} + R} \qquad (1)$$

(b) Rationalizing equation (1) gives

$$Y^1 = \frac{g_m \omega^2 C^2 R^2 + j g_m \omega CR}{1 + \omega^2 C^2 R^2} \qquad (2)$$

from which $$R^1 = \frac{1 + \omega^2 C^2 R^2}{g_m \omega^2 C^2 R^2} \qquad (3)$$

and $$C^1 = \frac{g_m CR}{1 + \omega^2 C^2 R^2} \qquad (4)$$

(c) Equation (4) may be re-written in the form

$$C^1 = \frac{\frac{g_m CR}{\omega^2 C^2}}{\frac{1}{\omega^2 C^2} + R^2}$$

(d) It can then be seen that, if the reactance of C is very large compared with R, then

$$C^1 = g_m CR \qquad (5)$$

49. Observation of equation (4) or (5) shows that the value of C^1 varies directly with g_m. Now g_m for a variable-mu pentode can be varied by changing the value of grid bias. If the bias is varied at an audio rate then the value of C^1 and hence the oscillator frequency also vary about their mean values at an audio rate. Section (a) of Fig 15 shows the circuit of Fig 13a after the addition of the necessary d.c. supply voltages and also indicates one method of injecting the a.f. modulating voltage into the grid of the reactance valve. The capacitors C2 and C3 have low reactances at r.f. and so do not in any way upset the phase-shift components C and R.

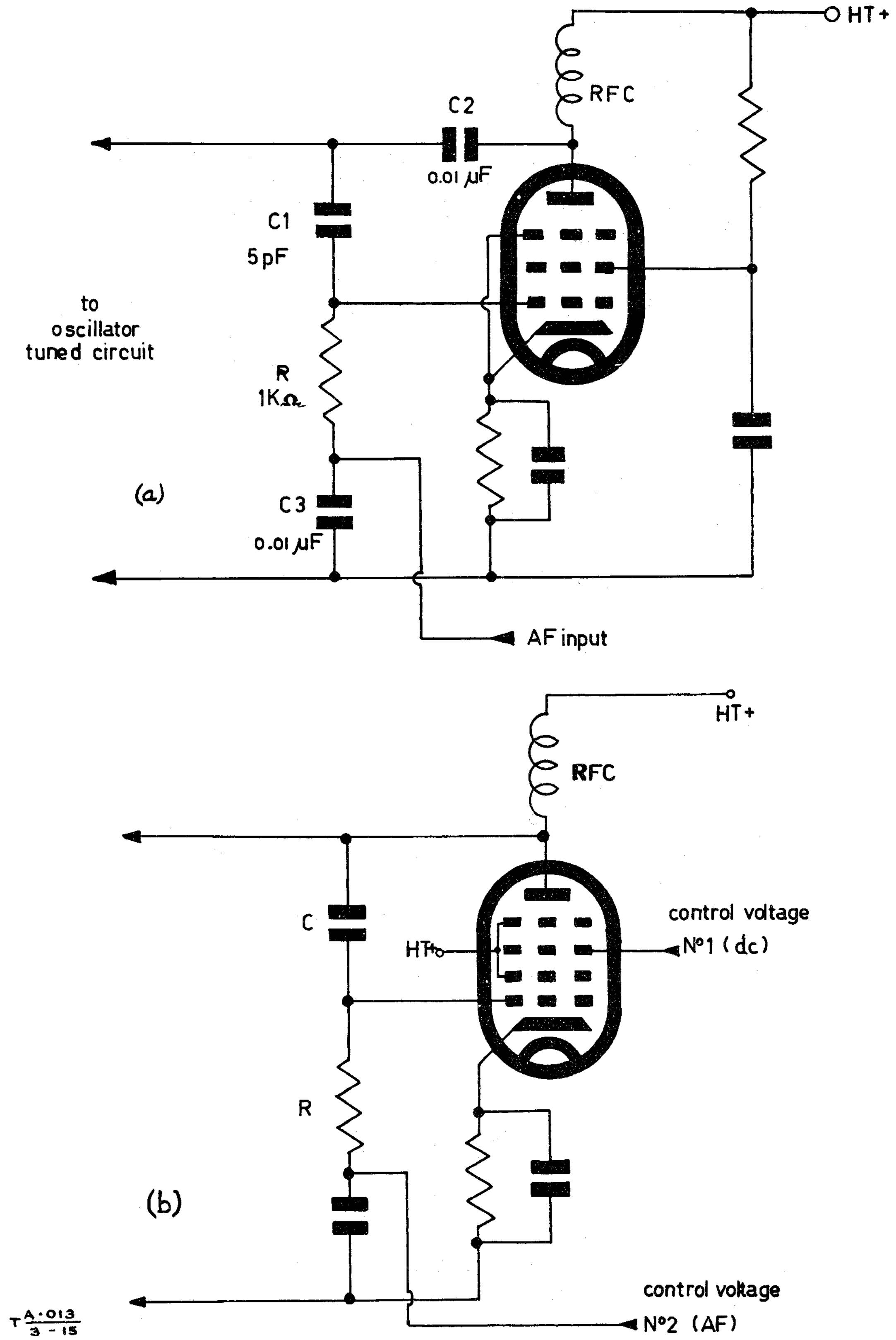

Fig 15 - Reactance valve circuits (capacitive)

50. Section (b) of Fig 15 shows the use of a hexode to simplify the injection of a.f.c. voltages. A.F.C. will be explained more fully in para 101 to 122 but note for the present that a d.c. voltage applied to grid 3 will alter the mean oscillator frequency and so enable frequency drift to be corrected. In sets not making use of a hexode the a.f.c. and the modulating voltage are applied in series to the control grid.

51. In some circuits the reactance valve operates as a variable inductance. In Fig 16 for example it can be shown that the valve behaves as an inductance L^1 with a value approximately $= \frac{CR}{g_m}$. Typical values of C and R for oscillator frequency of 5Mc/s are 25pF and 20kΩ respectively.

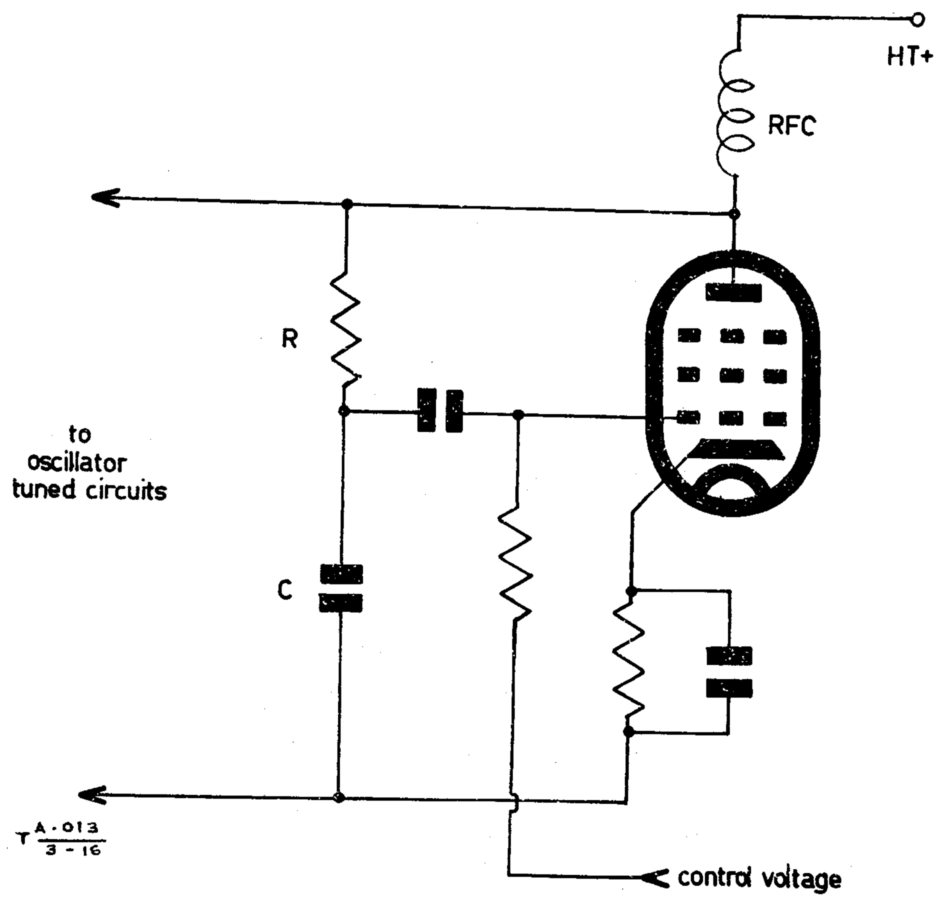

Fig 16 - Reactance valve circuit (inductive)

Ferrites

52. The behaviour of ferrites generally is beyond the scope of this regulation which will be limited to a brief introduction to ferrites and a description of the application of ferrite reactors in field radio sets.

53. Until recent years, frequency modulation has always been effected by means of valve circuits similar to those already described. Due to the devlopment of new materials the reactance valve has been replaced by the ferrite reactor in the new

v.h.f. field radio sets. Ferrites, because of their very high resistivity, have made it practicable to make use, at radio frequencies, of the change in inductance which occurs when a magnetic flux is applied to the ferro-magnetic core of a coil.

54. Ferrites are mixtures of oxides sintered together and in some cases forming solid solutions with one another. Their composition is chosen to give minimum electrical conductivity (high resistivity) in combination with the maximum magnetic effect desired for any given application. Their magnetic properties are such that initial permeabilities between 10 and 3,000 can be achieved, with resistivities at least 10 million times greater than Permalloy.

55. The combination of elements is chosen to give the desired characteristic for any given application. To produce a ferrite with a 'square' hysteresis loop suitable for switching applications in computers and for waveguide elements, magnesium and manganese oxides are usually added to ferric oxide. Manganese, zinc and nickel oxides are used to produce the core materials used in telecommunications.

Ferrite reactors

56. The inductance of a coil is dependent on the permeability of the core material. If the core is ferromagnetic (as are the ferrites) its permeability can be changed by placing it in a magnetic field. Thus if the coil is part of a resonant circuit, it is possible to change the frequency of oscillation by applying a magnetic field to the core. In v.h.f. field radio sets the magnetic field is applied by means of an electromagnet. The ferrite core (on which part of the r.f. tuned circuit inductance is wound) bridges the gap between the two poles of a horseshoe electromagnet. The winding of the electromagnet carries d.c. and voice frequency currents only and the core is of the conventional laminated type.

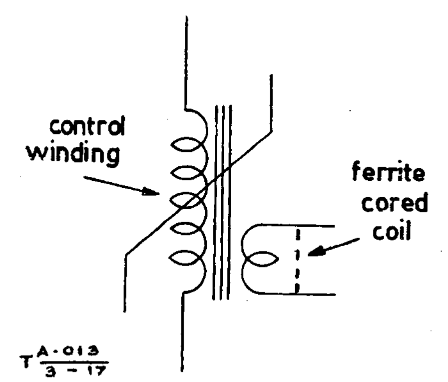

Fig 17 - Circuit symbol for a ferrite reactor

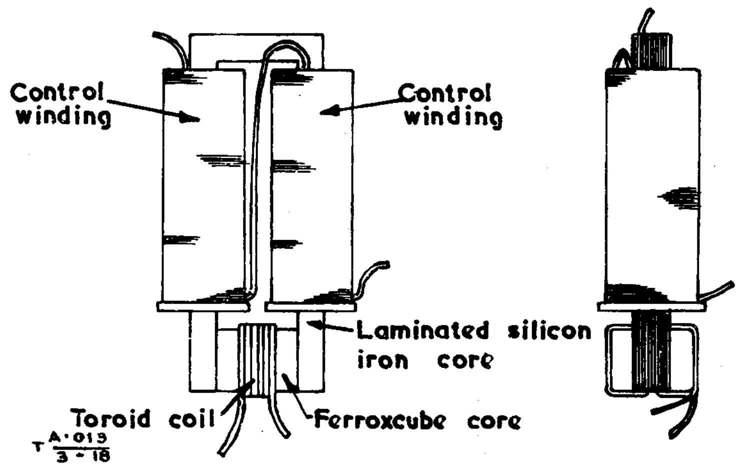

Fig 18 - Construction of ferrite reactor

57. The circuit symbol for a ferrite reactor is shown in Fig 17, while Fig 18 shows the physical construction. The control winding (ie the electromagnet winding) is in some texts called the primary but the ferrite reactor should not be thought of as a transformer. There is no intentional transfer of energy from one winding to the other. The sole purpose of passing current through the control winding is to

change the inductance of the ferrite cored coil. In fact in some ferrite reactor applications the control winding handles d.c. only.

58. The ferrite cored coil since it forms part of an r.f. tuning inductance has very few turns (eg 10 in one field radio set). The control winding in field radio applications usually forms the anode load of an a.f. amplifier valve and hence has a large number of turns and a high impedance to audio frequencies.

59. Fig 19 shows the control current versus inductance characteristic of a typical ferrite reactor (from WS C42). The inductance referred to being that of the ferrite cored coil. A d.c. current of 2.5mA suffices to bias this reactor to its normal operating point. A.F. currents superimposed upon the d.c. current can then be used to frequency modulate the output of any oscillator of which this reactor forms a part. Note that a change in the value of the d.c. 'biasing' current can be used to compensate for any drift of oscillator centre frequency.

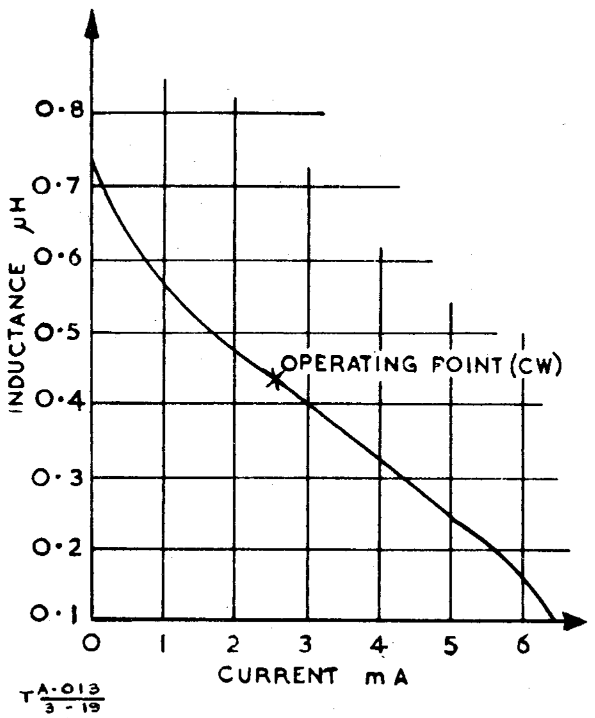

Fig 19 - Control current versus inductance characteristic of a ferrite reactor

60. Fig 20 shows, in a much simplified form, the practical application of a ferrite reactor in a v.h.f. transmitter. The quiescent anode current of valve V1 (the modulator) is used to bias the ferrite reactor to its correct operating point. Speech currents after amplification by V1 are used to frequency modulate the output of the transmitter oscillator V2. A sample of the oscillator output is examined by the a.f.c. circuits which generate, if necessary, a correction voltage. Application of the a.f.c. correction voltage to the grid of V1 shifts the operating point of the reactor, thus compensating for frequency drift. Generation of the a.f.c. correction voltage is treated in a later section of this regulation.

F.M. RECEIVER

General (see also para 37 to 39)

61. The basic circuit of an f.m. receiver is somewhat similar to that of an a.m. receiver of the superheterodyne type. The block diagram, Fig 21, shows the essential stages only. Desirable extras such as a.f.c. squelch circuits, and a tuning indicator have been omitted.

62. The heart of the receiver is the discriminator which is used to reverse the modulation process, ie the discriminator converts the frequency variations of the signal into a replica of the transmitter microphone output voltage. The limiter, which removes any a.m. that may exist in the signal, is essential if the full benefit of the many advantages of f.m. transmission is to be obtained.

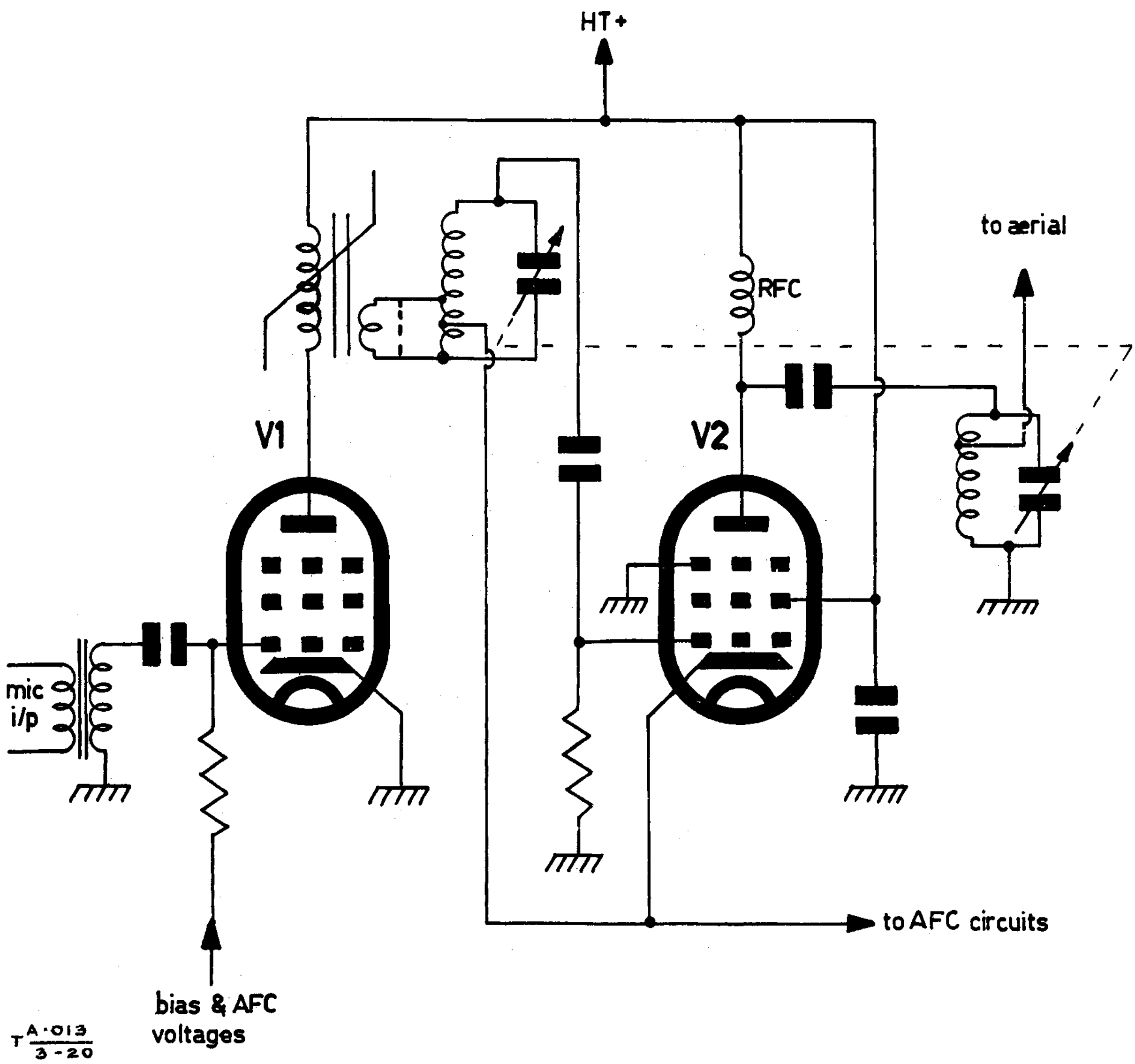

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Fig 20 - Simplified circuit of an F.M. transmitter using a ferrite reactor

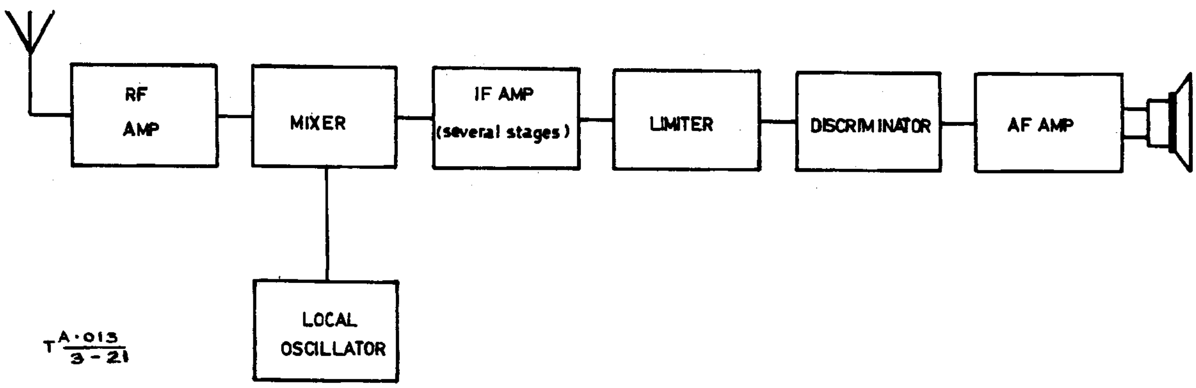

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Fig 21 - Basic f.m. receiver - block diagram

63. The i.f. amplifiers should be capable of passing all the significant sidebands (see para 18 and 19). The intermediate frequency is usually greater than 2Mc/s thus enabling the necessary bandwidth to be obtained while still retaining reasonable Q values for the i.f. tuned circuits. Techniques such as damping resistors, overcoupling, and staggered tuning are used to obtain the desired i.f. response.

64. Double conversion is frequently employed to reduce the possibility of instability in an i.f. channel which must of necessity consist of many stages.

65. Due to the combination of high overall receiver gain and wide bandwidth, the noise output in the absence of a signal is rather severe. Squelch circuits have been devised which mute the a.f. amplifier stage until the receiver is tuned to a carrier. For circuit details refer to the relevant equipment regulations.

Limiters

66. The limiter ensures that a constant amplitude waveform reaches the discriminator. Most equipments use only one limiter stage but for completely satisfactory amplitude limitation two limiter stages are necessary. Although f.m. detector circuits have been devised which are inherently insensitive to a.m. (eg the ratio-detector, the gated-beam detector, and the Bradley detector) the types of discriminator used in current field radio sets do respond to amplitude variations. Note that the limiter not only removes peaks due to interference but must also remove the amplitude modulation introduced onto the f.m. signal by the receiver due to the i.f. response curve not being perfectly flat over the whole channel width.

67. The type of limiter in field radio sets uses a pentode valve operating as a saturated i.f. amplifier. The circuit, Fig 22, is essentially that of a leaky-grid detector with the anode circuit tuned to the carrier instead of being designed to select the audio frequency. A non-variable-mu, short grid-base valve is used, operated with low anode and screen voltages so that anode current saturation occurs for all positive values of grid potential. (When a pentode is operated with low anode and screen potentials the operation region falls below the knee of the anode characteristic and the anode current becomes independent of the amplitude of the grid potential over wide ranges).

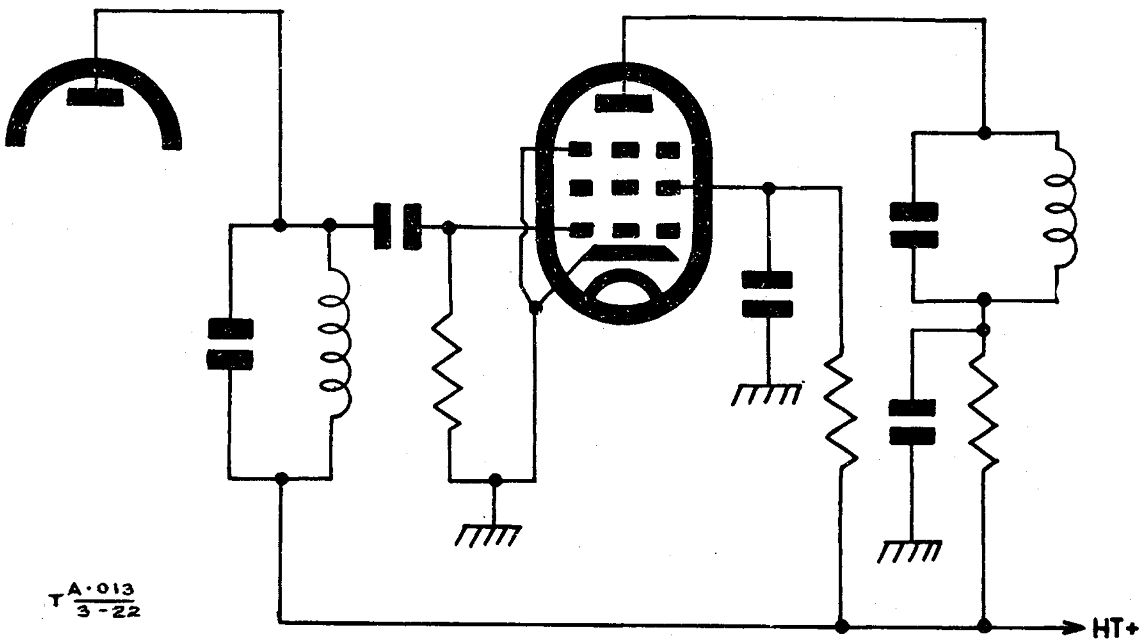
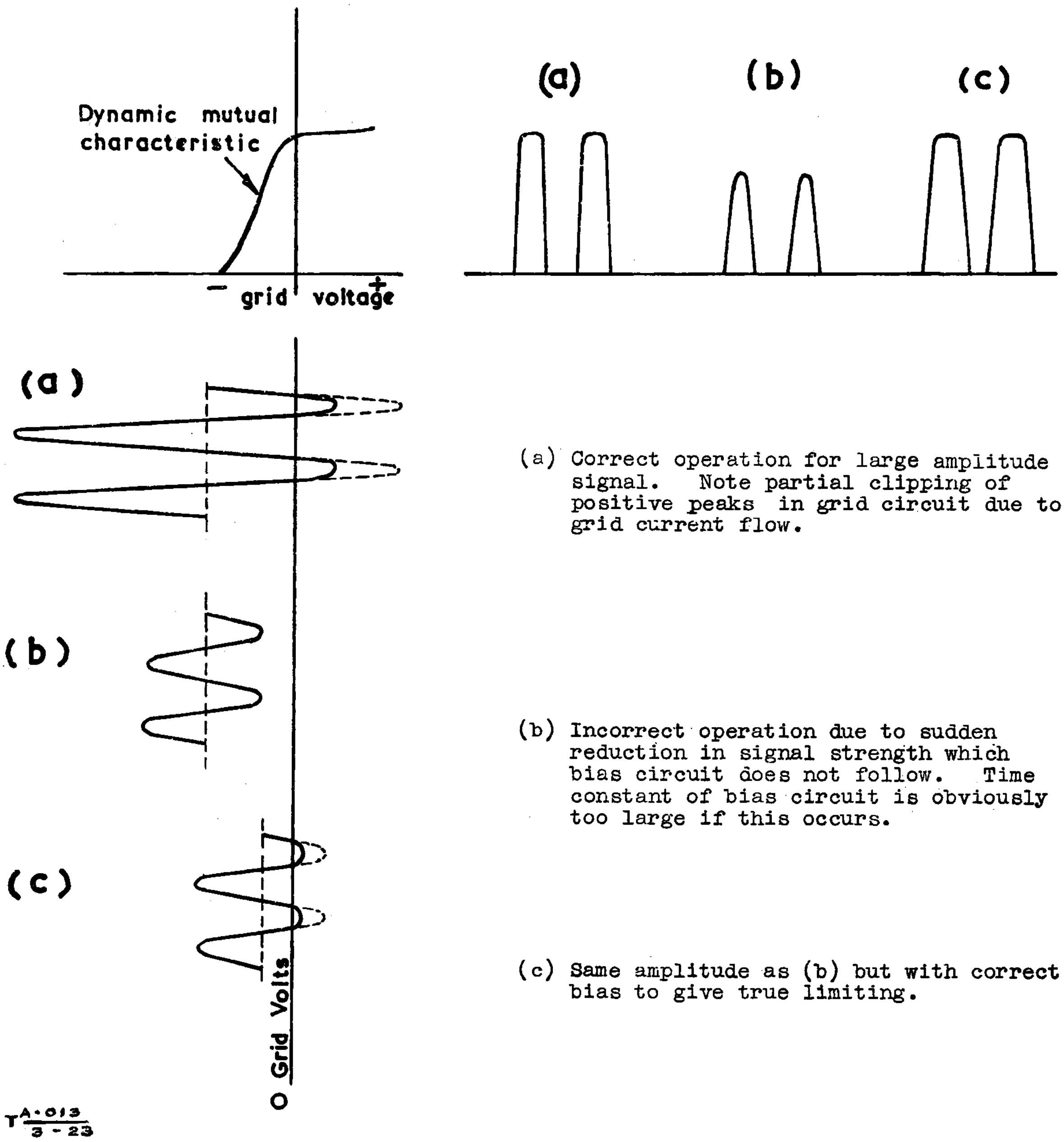


Fig 22 - Typical limiter circuit



(a) Correct operation for large amplitude signal. Note partial clipping of positive peaks in grid circuit due to grid current flow.

(b) Incorrect operation due to sudden reduction in signal strength which bias circuit does not follow. Time constant of bias circuit is obviously too large if this occurs.

(c) Same amplitude as (b) but with correct bias to give true limiting.

Fig 23 - Limiter performance curves

68. The self-bias arrangement must be able to follow any amplitude variations otherwise reductions in signal strength will cause the positive signal peaks to fall within the grid-base, thereby impairing the limiting process. Waveforms (b) and (c) of Fig 23 clarify this point.

69. Before f.m. detection takes place, the rectangular anode current waveforms of Fig 23 are rounded-off by the tuned circuit.

70. Since, at the grid of the limiter, there is a negative voltage proportional to carrier amplitude it may be used as a source of a.g.c. for the r.f. and i.f. stages. At first sight the use of a.g.c. on an f.m. receiver appears to be superfluous and in most field radio sets it is not used. However, the use of a.g.c. on very strong signals prevents overloading of the mixer stage and so reduces the possibility of the generation of undesired spurious responses. A.G.C. of the mixer stage is not usually employed because of electron-coupling between the signal and oscillator circuits, any change of which causes oscillator frequency drift.

Discriminators

General

71. Many discriminator circuits have been developed, some of which will be considered in detail below. Both of the types used in current field radio sets (ie the Bond and the Foster-Seeley) demodulate f.m. waves by first converting variations of frequency into variations of amplitude and then applying the resultant a.m. waves to conventional detector diodes. A typical characteristic for a discriminator of this type is shown in Fig 24. As can be seen from this figure the unmodulated carrier produces zero output. An increase in frequency produces a voltage with magnitude proportional to the extent of the frequency increase. Similarly for a frequency decrease except that the polarity is reversed. Thus if a modulated signal is applied to the discriminator the audio output is proportional to the maximum frequency deviation and also to the slope of the discriminator characteristic.

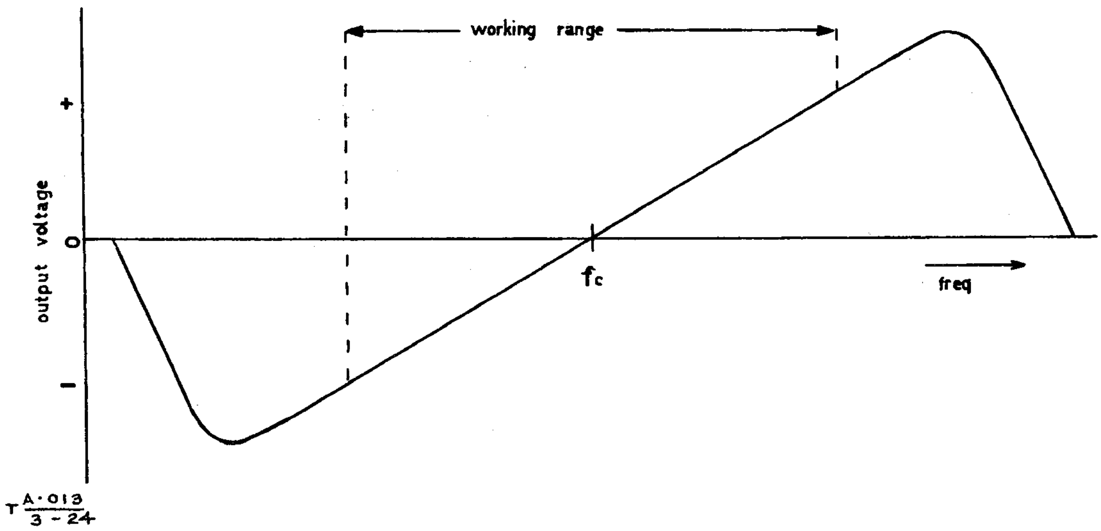


Fig 24 - Discriminator characteristics

72. Note the steep return slope at the extremes of the characteristic shown in Fig 24. Even if this steep slope is not inherent in the design of the discriminator stage itself it will nevertheless form a part of the overall receiver characteristic due to the limitations imposed by i.f. bandwidth. As stated in para 71 the audio output is proportional to discriminator slope so that when tuning a field radio set to an f.m. signal generator (or vice versa) during the alignment procedure, the output meter response will, in general, be three-humped as shown in Fig 25. The centre and smallest of the three peaks is the correct tuning point. The exact shape of Fig 25 will depend on the amount of frequency deviation used. For large values of deviation the centre peak will increase in magnitude and width and may absorb the other two peaks.

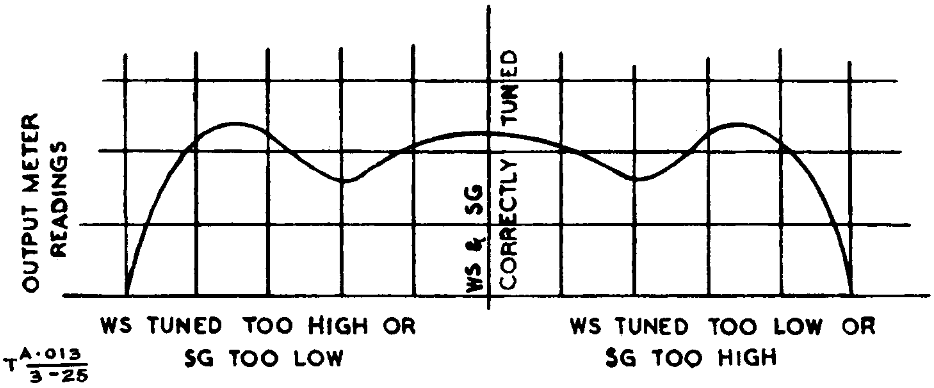


Fig 25 - Tuning curve of wireless set during alignment

De-tuned circuit used as a discriminator

73. The simplest form of discriminator is a parallel resonant circuit detuned slightly from the centre frequency of the signal. Fig 26 shows the circuit and method of operation. Changes of frequency are converted into changes of amplitude and the resulting wave rectified by the diode to extract the intelligence. In actual fact the signal applied to the diode is both frequency and amplitude modulated but the diode responds to the a.m. only.

74. The imperfect reception of f.m. signals by a.m. receivers depends upon the above principle. A conventional a.m. receiver, if tuned to one side of the carrier frequency, will yield an intelligible (but imperfect) output in response to an f.m. signal.

Double-tuned (Round-Travis) discriminator

75. The elementary circuit just described introduces distortion because the slope of the resonance curve is not linear. Only by the use of a very low Q circuit and by accepting the consequent low output can adequate linearity be obtained. The Round-Travis discriminator considerably reduces the distortion and increases the output in a manner analogous to the way in which class AB push-pull operation increases the output and reduces the distortion in audio frequency amplifiers.

76. In the circuit, Fig 27, the primary is tuned to the signal centre-frequency f_c, secondary A is tuned to a frequency somewhat higher, and secondary B to a frequency somewhat lower than f_c. Fig 28, which explains the operation of the circuit, should be compared with Fig 24. Note that although the voltages across R1 and R2 are in opposition the changes in these voltages in response to a change in frequency are additive in the output, ie audio output voltage is double that of the previous circuit and linearitey over the working range of frequencies has been considerably improved.

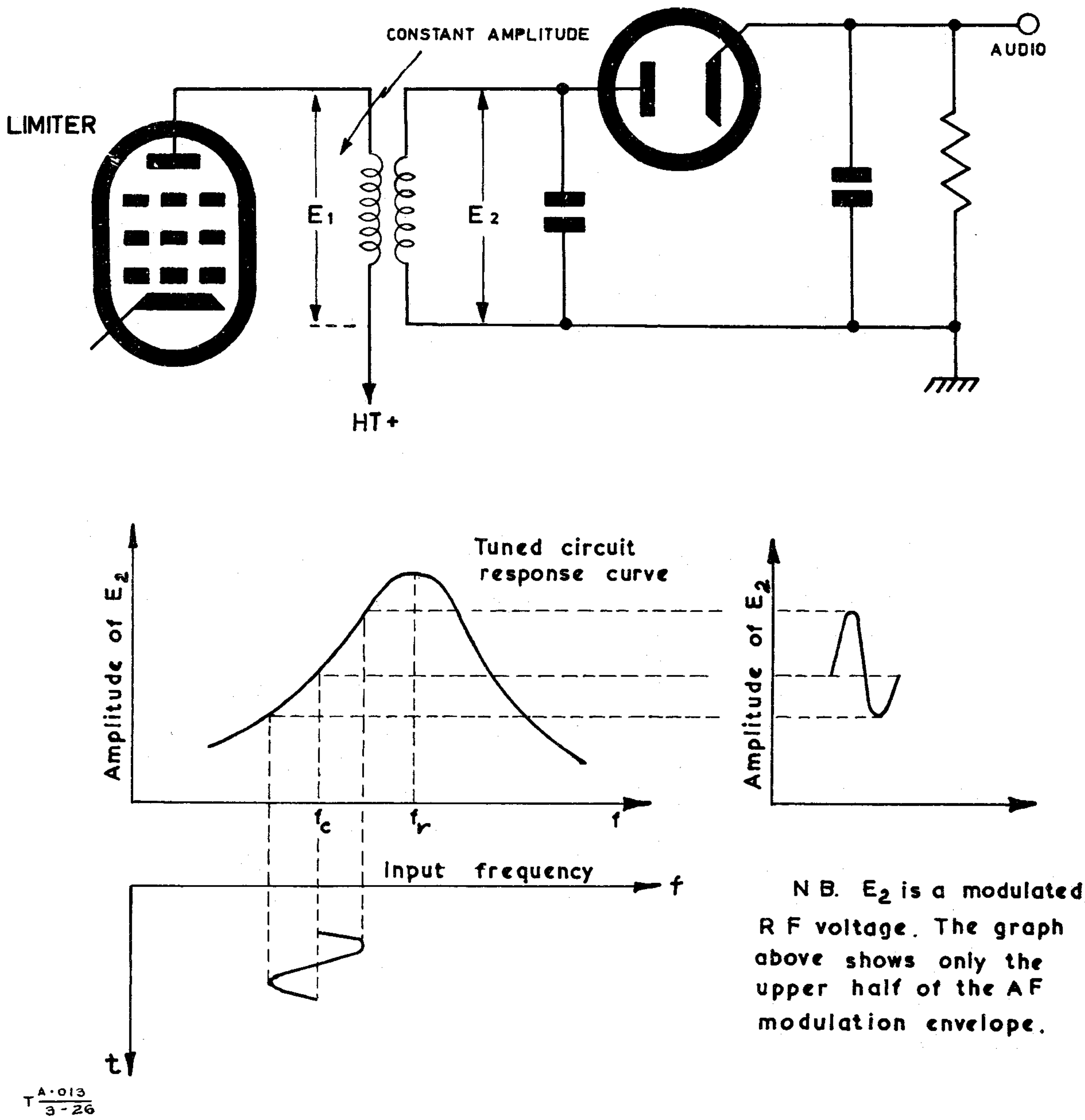


Fig 26 – Discriminator elementary circuit and curves

Foster-Seeley discriminator

77. The two discriminator types previously discussed are so-called 'amplitude' discriminators. The Foster-Seeley discriminator is an example of what is sometimes called a 'phase' discriminator for reasons which will become apparent when its mode of operation is discussed.

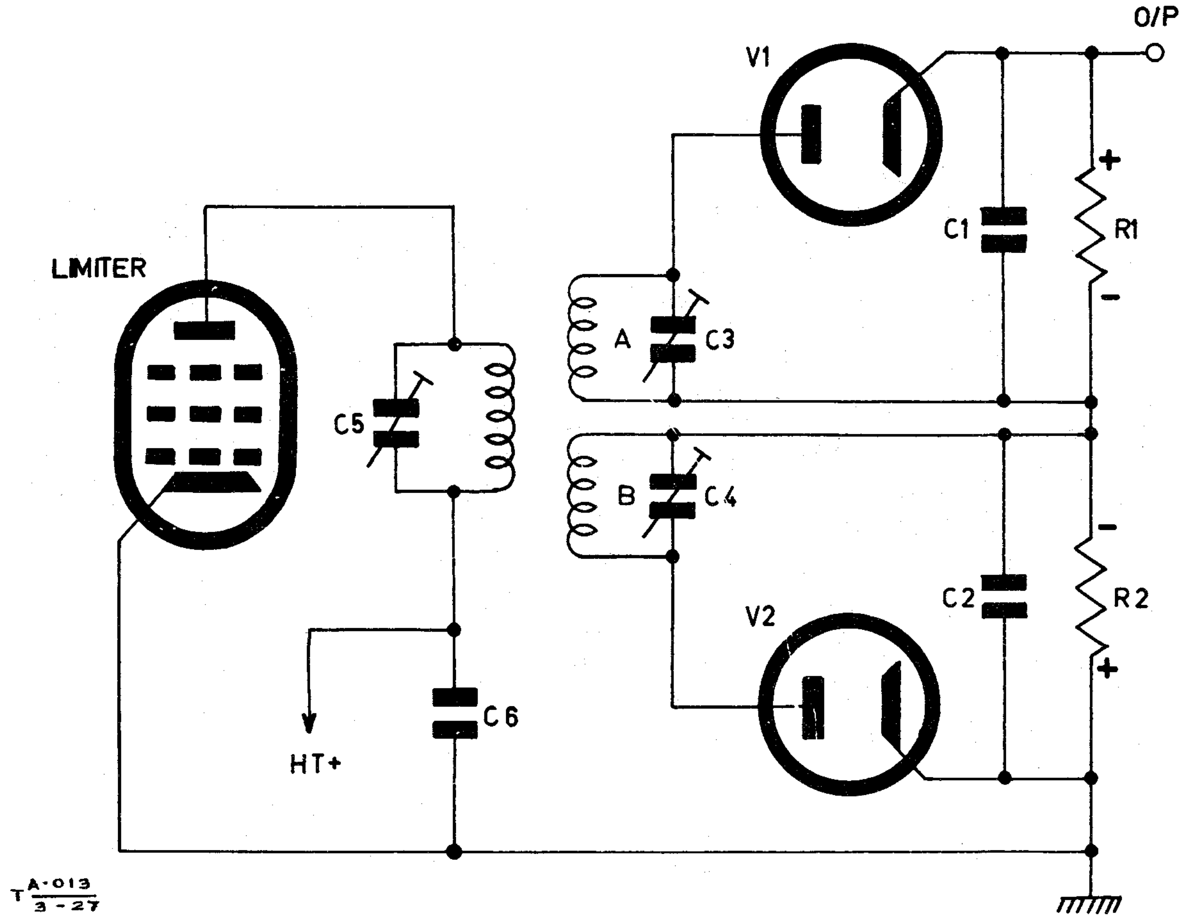


Fig 27 - Round-Travis (amplitude) discriminator

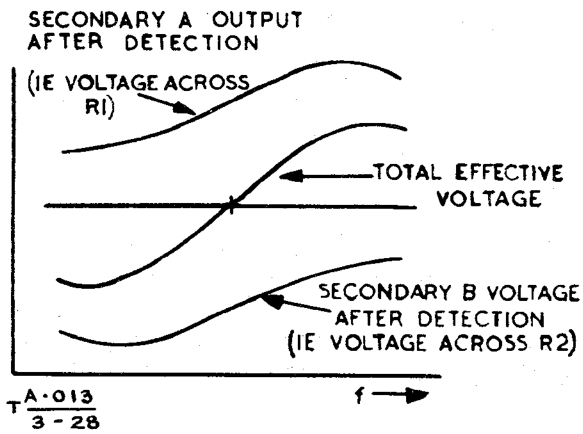


Fig 28 - Round-Travis discriminator frequency response curves

78. Both the tuned circuits of Fig 29 are tuned to the signal centre-frequency. The output circuit of the diode is the same as the Round-Travis. The capacitors C1 and C3 have such low r.f. reactance that the full primary voltage V_p appears across the r.f. choke. Each diode is therefore being fed with the vector sum of V_p plus half the voltage appearing across the secondary winding. When making these vector additions relative phase is extremely important.

79. From the theory of inductively coupled circuits the voltage induced in any secondary circuit by a primary current I_p has a magnitude of ωMI_p and lags the current that produces it by 90°. In complex notation the induced voltage is $-j\omega MI_p$ (where M = mutual inductance). It is important to note that this voltage

is a series voltage. The resulting secondary current and the voltage distribution in the secondary circuit depend upon the secondary circuit series resistance.

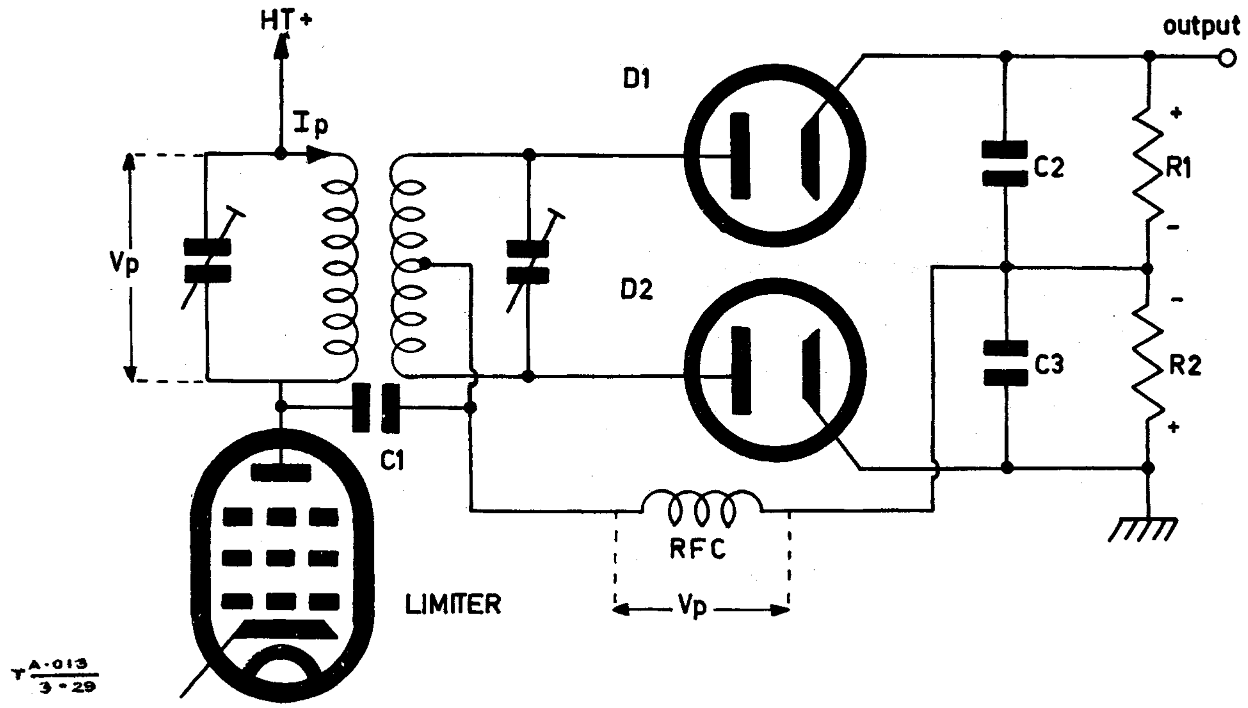


Fig 29 - Foster-Seeley discriminator circuit

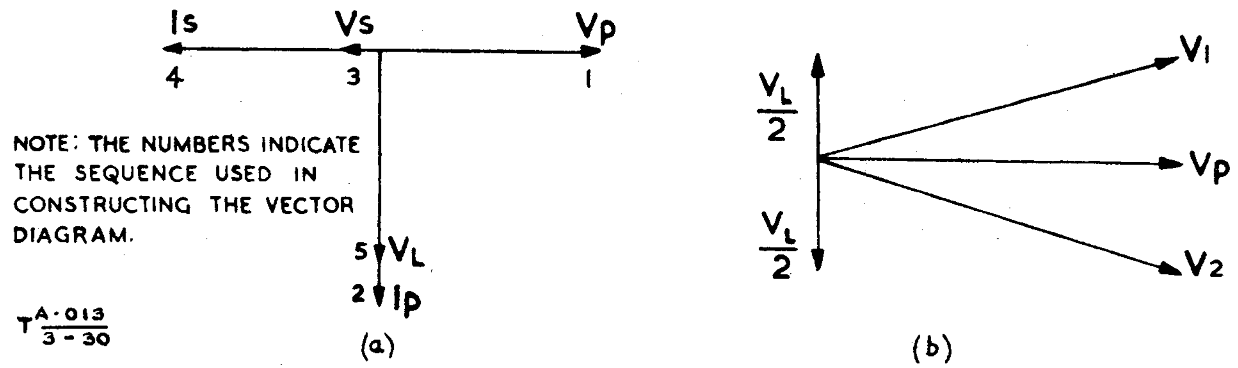


Fig 30 - Vector diagram at centre-frequency

80. Section (a) of Fig 30 shows, in vector form, the situation when the incoming signal is unmodulated and correctly tuned. V_p the reference vector is shown leading the primary current by 90°. I_s is the secondary current which is in phase with V_s since the circuit is resonant. V_L is the voltage developed across the secondary inductance due to the flow of I_s. As shown vectorially at (b) in Fig 30 the voltage V_L is split into equal and opposite halves by the centre tap. The resultant voltages V_1 and V_2 which are applied to the diodes D1 and D2 respectively are obviously equal in magnitude and the discriminator output is therefore zero.

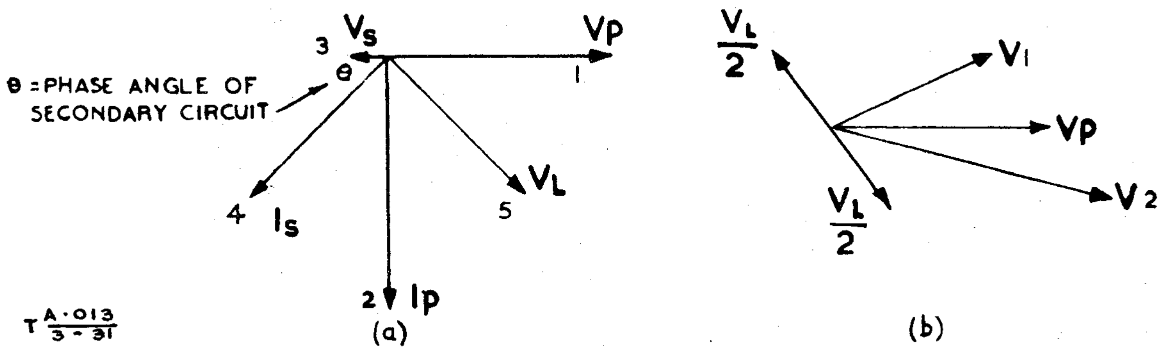


Fig 31 - Vector diagram below resonanace

81. Fig 31 shows the case where the incoming frequency is less than the centre-frequency. I_s now leads V_s since a series tuned circuit behaves capacitively below its resonant frequency. The values of I_s and V_L will be reduced somewhat since the circuit is no longer resonant. Section (b) of Fig 31 shows that the r.f. voltages applied to the two diodes are now no longer equal. Since V_2 is greater than V_1 the discriminator output is negative.

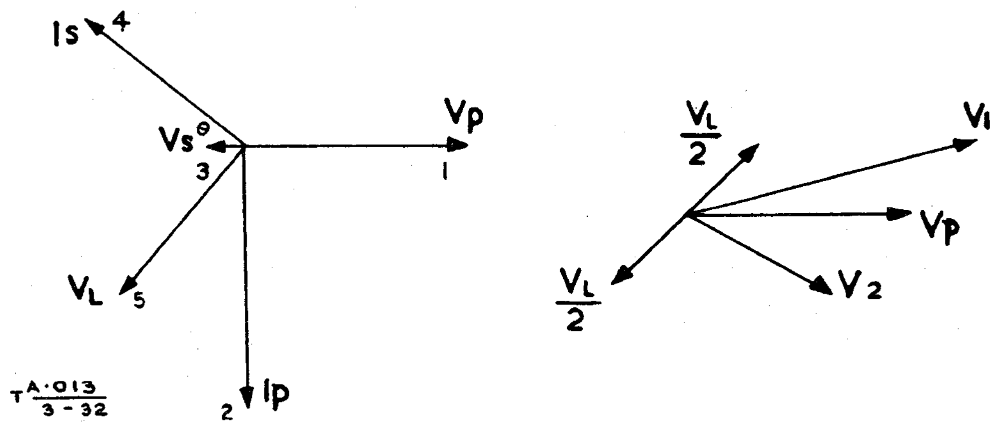


Fig 32 - Vector diagram above resonance

82. The operation at frequencies above the centre-frequency should now be self-evident from the vector diagrams of Fig 32.

83. The slight changes in the amplitude of Vp with frequency have been ignored in the above treatment. The manner in which Vp varies with frequency depends upon the coefficient of coupling employed between primary and secondary. Some control over the linearity of the characteristic is therefore possible. By correct design the working range is linear and the characteristic appears as in Fig 24.

Bond discriminator

84. The Bond discriminator appears in various forms, some of which are shown in Fig 33 and 1002 - 1004. In each case the symbol ● indicates which of the output terminals is the more positive when the input frequency is greater than the centre-frequency. The form shown in Fig 33 will be considered in detail in order to explain the principle of operation.

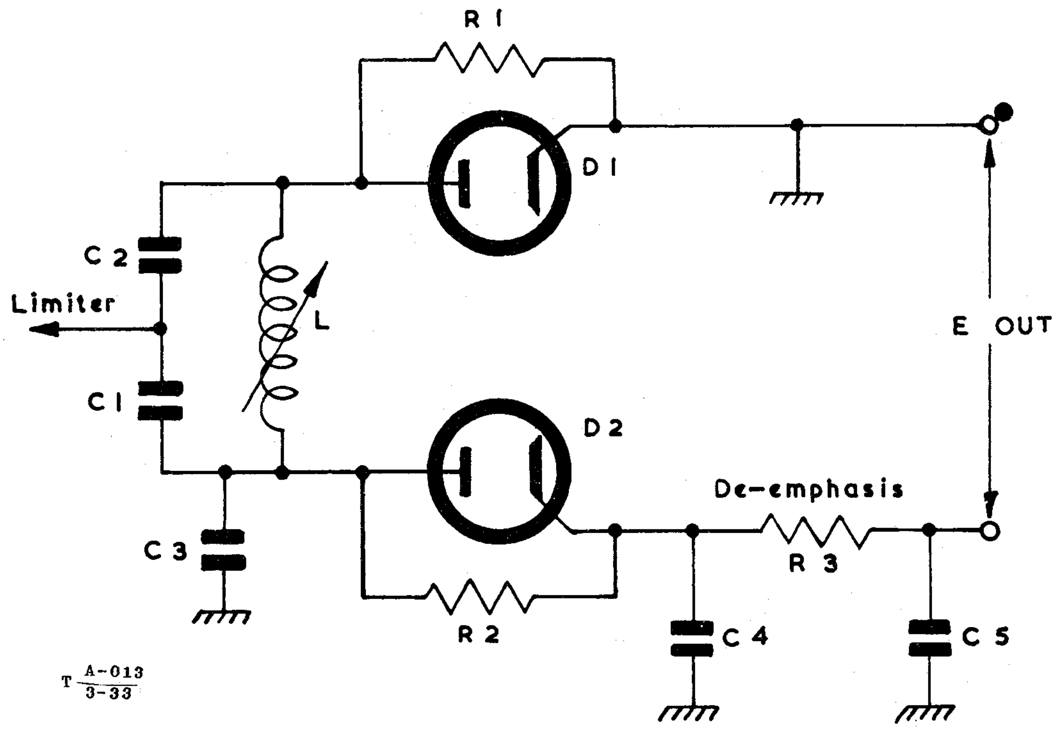


Fig 33 - Bond discriminator - circuit diagram

85. Fig 34 shows a simplified r.f. equivalent circuit of Fig 33. To simplify the analysis the internal impedance of the generator and the loading effect of the diodes have been ignored. The generator, representing the limiter stage, is assumed to have a constant voltage output of variable frequency. A and B are the diode points of attachment. R is the effective resistance of the coil. Component values have been chosen to give a centre operating frequency of approximately 2.4Mc/s.

86. The variation of circuit impedance with frequency is shown in Fig 35. The peak at approximately 2.46Mc/s is due to the parallel resonance of C1, C2, L and R. The minimum at 2.395Mc/s is due to C3 forming a series resonant circuit with the combination C1, C2, L and R which behaves as an inductance at this frequency.

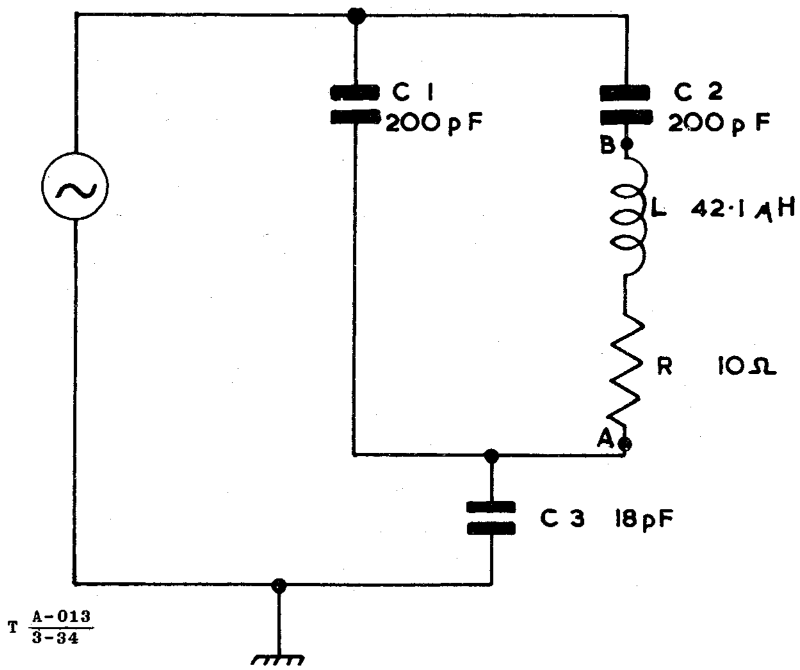


Fig 34 - Simplified r.f. equivalent circuit

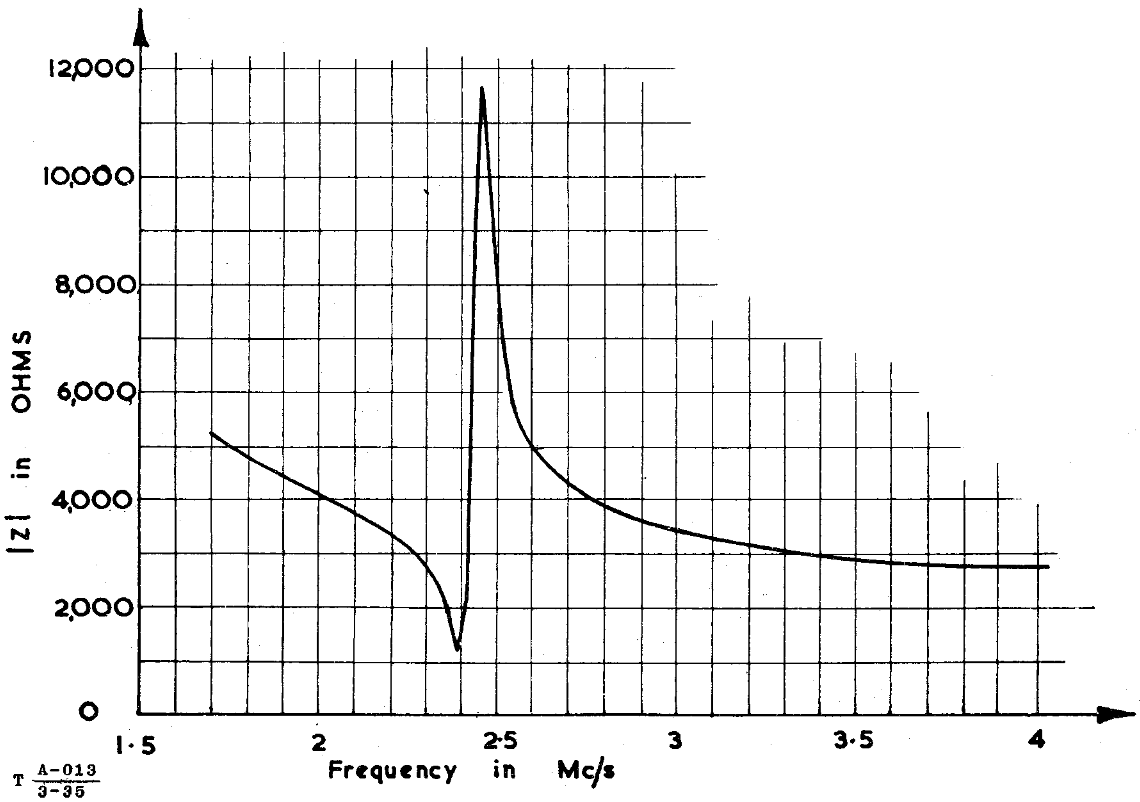


Fig 35 - Impedance variations

87. Assuming an input voltage of one volt, curve A of Fig 36 shows the magnitude of the r.f. voltage from point A to earth (ie across C3). Note the peak at 2.395Mc/s due to the relatively large line current which flows at this frequency. Curve B shows the magnitude of the r.f. voltage from point B to earth. Both of these curves have been obtained by calculations making use of complex algebra. An outline of the procedure is given in para 91. Note that the curves represent magnitudes only. At the crossover point (2.399Mc/s) the magnitudes are equal but this does not mean that there is no r.f. potential across the coil at this frequency. When relative phase is taken into account the voltage across the coil at 2.399Mc/s is found to be a little over 5V.

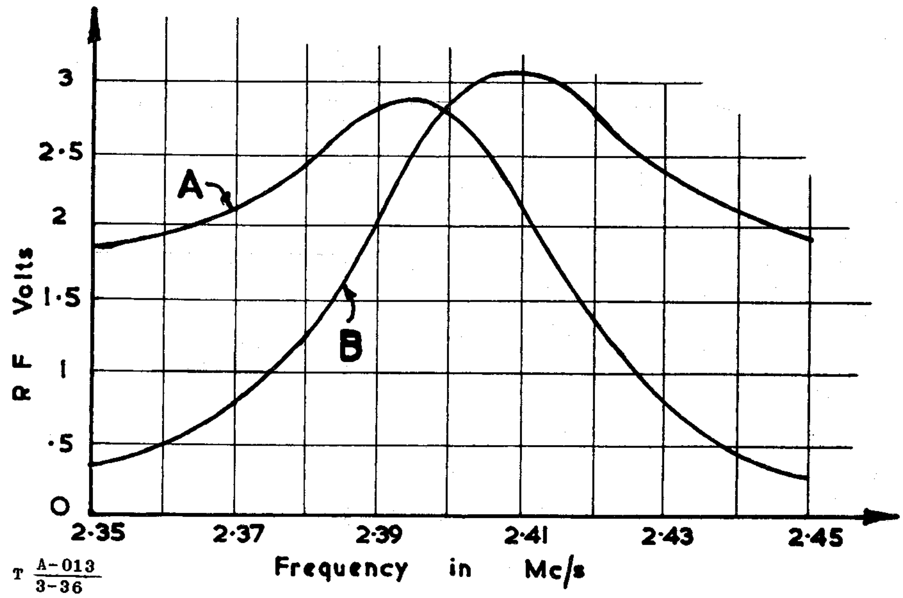


Fig 36 - R.F. potentials existing across L

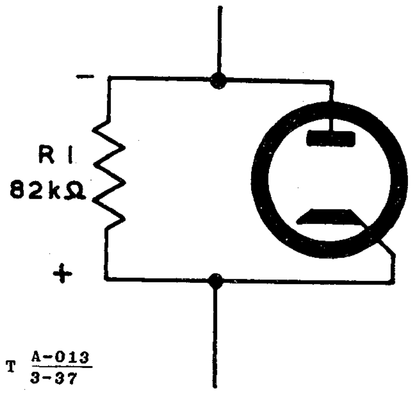


Fig 37 - Circuit used to rectify r.f. voltage at point B on Fig 34

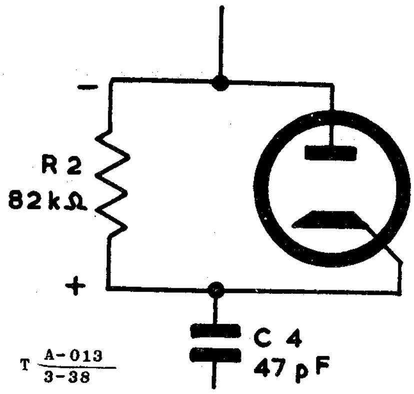


Fig 38 - Circuit used to rectify r.f. voltage at point A on Fig 34

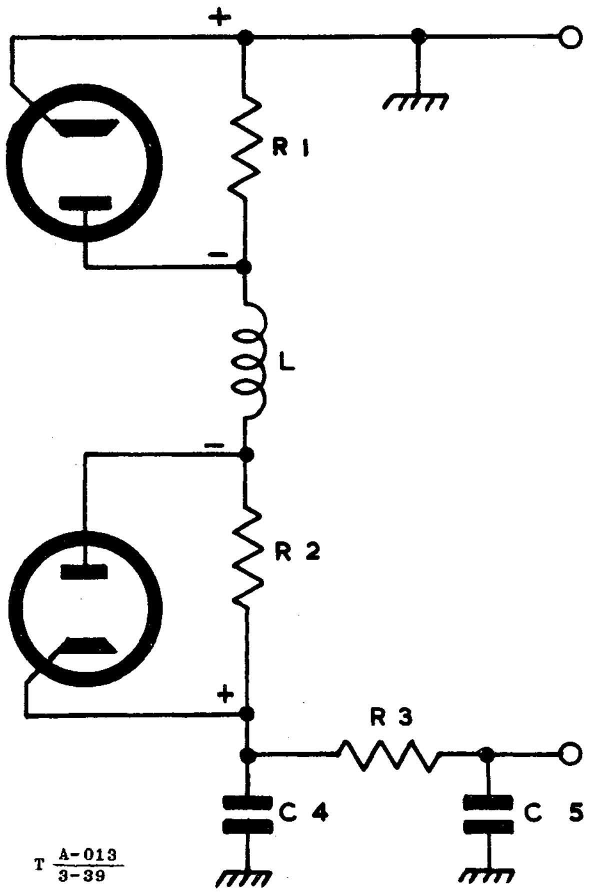


Fig 39 - Simplified output circuit

88. If the combination shown in Fig 37 is placed across a source of r.f. voltage with infinite internal impedance, a d.c. voltage of the polarity indicated will appear across the resistor. The magnitude of the d.c. voltage being directly proportional to that of the r.f. voltage. Hence if the combination is placed between point B and earth, the d.c. response curve would have exactly the same shape as that of curve B of Fig 36.

89. Due to the low impedance of the 47pF capacitor at 2.4Mc/s (ie 1,410Ω which is negligible compared with 82kΩ) the circuit of Fig 38 will behave in essentially the same fashion as that of Fig 37 when placed across a source of r.f. voltage. Hence when this combination is placed across C3 a graph of the d.c. voltage across R2 would have exactly the same shape as that of curve A of Fig 36.

90. Fig 39 shows that the d.c. voltages produced across R1 and R2 are in opposition in the output. Thus at 2.399Mc/s when these two voltages are equal there is zero output from the Bond discriminator. When the frequency increases (to say 2.41Mc/s) the voltage across R1 rises while that across R2 falls, resulting in a negative output voltage (and vice versa for a frequency decrease). Fig 40 shows this in graphical form, the ordinates for this curve being proportional to the difference between curves A and B of Fig 36.

91. The following is a summary of the method used to obtain the ordinates for curves A and B of Fig 36. A frequency of 2.385Mc/s has been used in this example.

Impedance of C1	=	$0 - j333.7$	=	$333.7\underline{/-90^\circ}$
Impedance of L and R in series	=	$10 + j630.9$	=	$630.9\underline{/89.1^\circ}$
Impedance of C2, L and R in series	=	$10 + j297.2$	=	$297.2\underline{/88.1^\circ}$
Impedance of C1 in parallel with C2, L and R	=	$778 + j2500$	=	$2620\underline{/72.7^\circ}$
Impedance of C3	=	$0 - j3707$	=	$3707\underline{/-90^\circ}$
Total circuit impedance	=	$778 - j1207$	=	$1435\underline{/-57.2^\circ}$
Voltage across C3 if supply voltage is 1V	=	$\dfrac{3707\underline{/-90^\circ}}{1435\underline{/-57.2^\circ}}$	=	$2.58\underline{/-32.8^\circ}$
			=	$2.17 - j1.40$

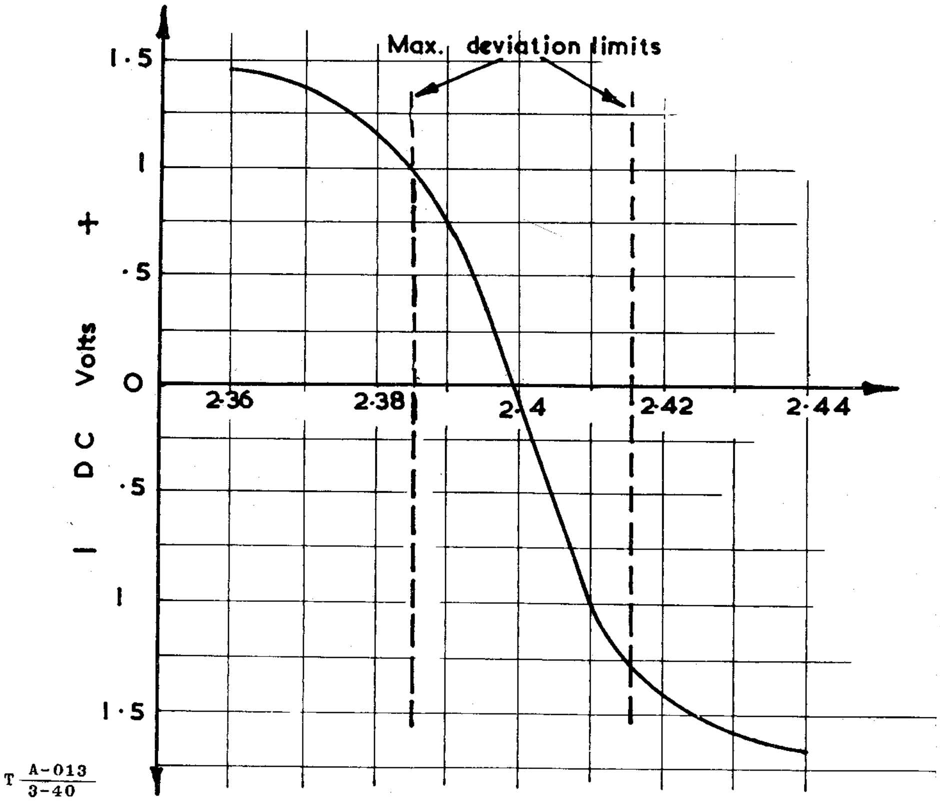


Fig 40 - Theoretical output curve

92. Thus 2.58 is the ordinate of curve A at 2.385Mc/s

$$\text{Voltage across L and R} = \frac{2620\angle 72.7^\circ}{1435\angle -57.2^\circ} \times \frac{630.9\angle 89.1^\circ}{297.2\angle 88.1^\circ}$$

$$= 3.88\angle 130.9^\circ = -2.54 + j2.97$$

$$\text{Voltage from point B to earth} = (-2.54 + j2.97) + (2.17 - j1.40)$$

$$= -0.37 + j1.57 = 1.61\angle 103.3^\circ$$

93. Thus 1.61 is the ordinate of curve B at 2.385Mc/s. Fig 41 shows this situation in vector form. Note that the above treatment is idealized. Apart from the simplifications mentioned in para 85 it should be borne in mind that the response of shunt diode circuits is greatly dependent upon the impedance of the r.f. source involved; thus the practical equivalents of curves A and B Fig 36 may differ somewhat from those shown.

Counter discriminator

94. The counter discriminator relies for its operation on the following property of a frequency modulated wave; that the time for any half cycle is slightly different from that of the half cycles on either side. Fig 42 depicts an f.m. signal sinewave modulated at an audio rate.

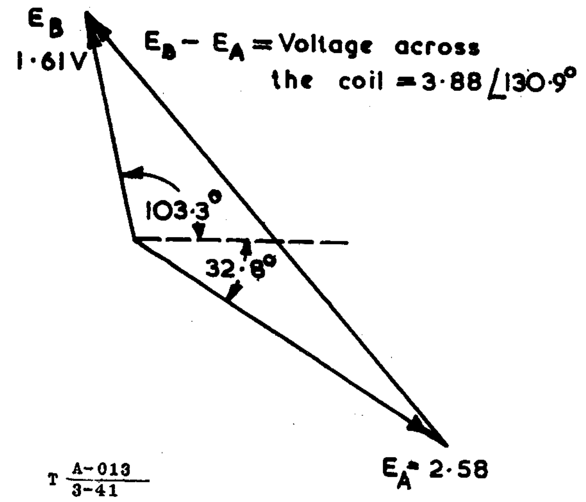
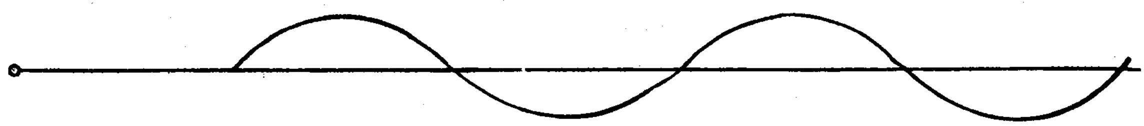


Fig 41 - Vector diagram showing voltage relationships at 2.385Mc/s

95. By amplifying and then limiting this signal the result is a series of squarewaves of varying width (Fig 43). The amplitude of this squarewave signal is maintained at a constant level since this is a requirement for the correct operation of the discriminator.

96. The counter discriminator operates as follows. On the positive half-cycle of one of the waves in Fig 43, diode A conducts and the capacitor C charges rapidly (time-constant: for 5kc/s deviation 0.1μsec, for 25kc/s 0.025μsec, for 75kc/s 0.01μsec) so that the charge across C almost reaches a steady state before the negative-going portion of the waveform occurs. The anode potential of diode A approximates to the cathode potential which is determined by the small positive bias applied from h.t. +ve via R1.



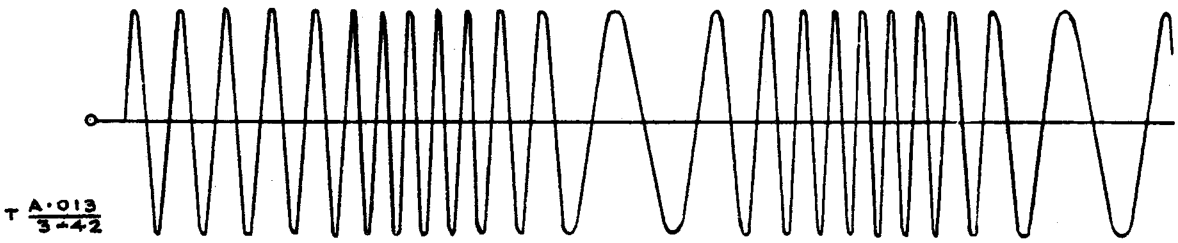


Fig 42 - F.M. signal sinewave modulated

97. The negative portion of the waveform causes A to cut off and B to conduct when the small positive bias is exceeded (the positive bias being the charge on C). C is almost completely discharged before the onset of the next positive half-cycle.

98. The output, developed across R is therefore a series of negative-going pulses as in Fig 45, since diode A effectively clamps the reference level at the cathode potential of A.

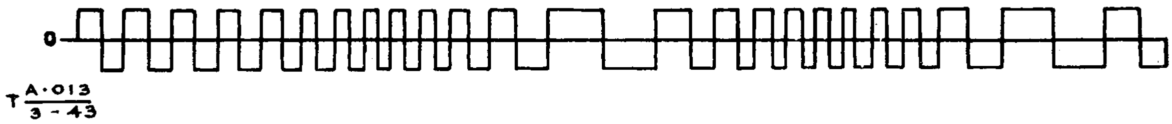


Fig 43 - Squarewave signal after amplifying and limiting

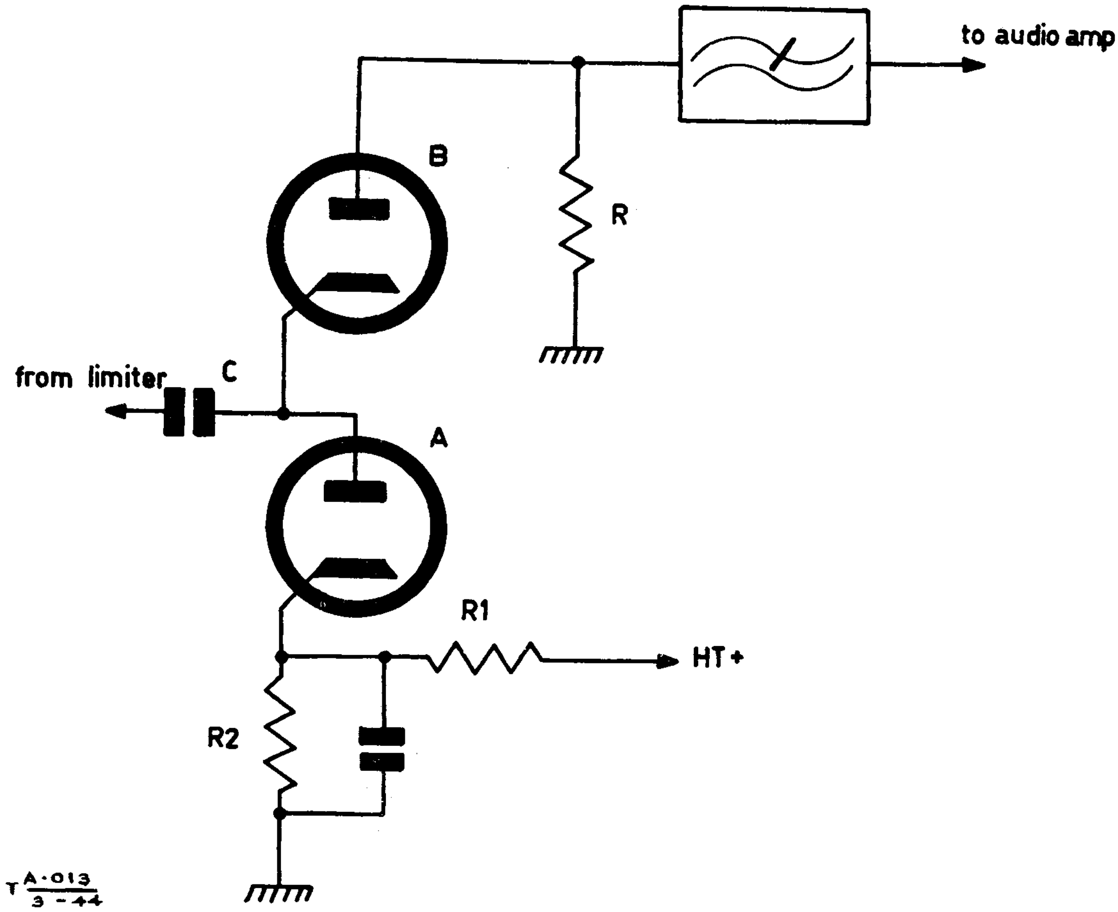


Fig 44 - Counter discriminator - basic diagram

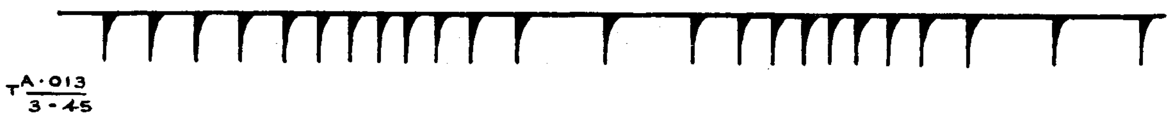


Fig 45 - Signal developed across R

99. Now consider the pulse train as shown in Fig 45. The amplitude of the pulses is constant owing to the constant input. The area under the pulse curve (shown shaded in Fig 45) is determined by the discharge time-constant only and is therefore constant for a given deviation (see para 96). The only variable is the pulse rate which depends on the instantaneous frequency of the signal applied to the discriminator.

100. If the signal applied to the discriminator is 100kc/s unmodulated, the output from the discriminator will be a series of identical pulses at the rate of 100,000 per second. If the signal is modulated the pulse rate will vary with the deviation. Take the case of a 100kc/s signal modulated at 1,000c/s. One cycle of the modulating signal will alter the time of each of 100 cycles of the 100kc/s signal ie each cycle of the modulated signal will be different. Thus the pulses from the discriminator will vary in time exactly as each cycle of the modulated signal varies.

101. By applying the pulse train output of the discriminator to an integrating network which includes a low pass filter, a negative voltage will result. For an unmodulated input signal the output voltage will be constant and its amplitude will depend on the frequency; ie more pulses integrated = greater voltage. This output voltage for an unmodulated signal is the reference level for the reproduction of any intelligence contained by the modulated signal. As the frequency increases the number of pulses increases and the amplitude of the output voltage increases. As the frequency decreases, the number of pulses decreases and the amplitude of the output decreases. Fig 46 shows the output from the low pass filter, ie the sinewave of Fig 42.

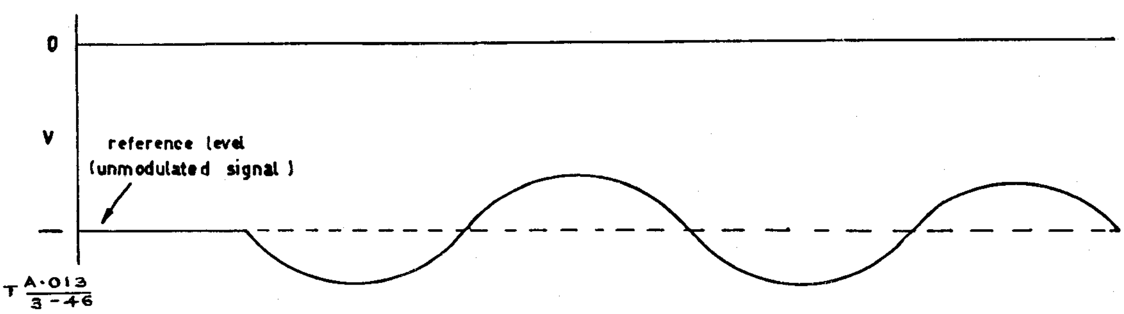


Fig 46 - Output from low pass filter

102. This output (Fig 46) is directly proportional to the deviation of the input signal. For this reason it has an important application in the accurate measurement of frequency deviation. Because of complicated circuitry it is not used in field radio sets but is used in deviation meters.

AUTOMATIC FREQUENCY CONTROL (A.F.C.)

A.F.C. as applied to receivers

103. Inaccurate tuning by the receiver operator, transmitter frequency drift, or frequency drift in the receiver local oscillator may cause the frequency applied to the i.f. amplifier to stray from the centre of the i.f. passband. An automatic frequency control (a.f.c.) system is used to reduce this tuning error.

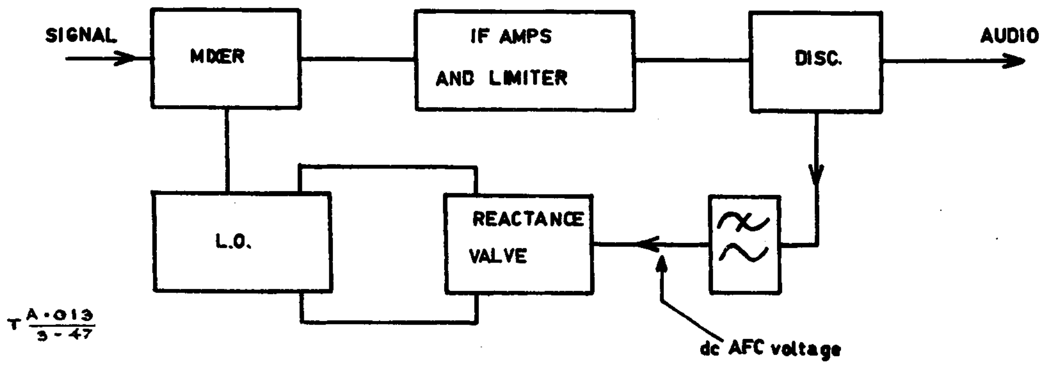


Fig 47 - Basic receiver a.f.c. system

104. The main requirement is a device which will

(a) give an output proportional to the amount of error, and

(b) indicate by the polarity of the output the direction of the error.

105. In f.m. receivers the existing discriminator will do this since if the signal is off-tune the discriminator will supply, in addition to the normal audio output, a d.c. output conforming to the above requirements. A low pass filter enables separation of d.c. from audio output for a.f.c. purposes. Note that a.f.c. can be applied to a.m. receivers if a discriminator fed from the i.f. channel is incorporated in the circuit (see para 50 for method of applying a.f.c. voltages to reactance valves).

106. When tuning a receiver fitted with a.f.c. the frequency at which the a.f.c comes into operation and tends to pull the receiver into tune with the incoming carrier is called the 'pull-in frequency'. When tuning away from a carrier the frequency at which the a.f.c. loses control is called the 'throw-out frequency'. The throw-out frequency is usually much greater than the pull-in frequency. Because of this one or two adjacent stations many be missed when searching with a.f.c. on. Hence a.f.c. should, if possible, be disconnected during tuning. Note that if a station drifts outside the 'pull-in' range and the transmission is interrupted (even if only for an instant) then that station is lost. For this reason mechanical a.f.c. is used in some high grade equipments (particularly on s.s.b. working where tuning requirements are stringent).

107. In the mechanical a.f.c. system the polarity of the a.f.c. voltage is used to control the direction of rotation of a small electric motor, the shaft of which is mechanically coupled to a tuning capacitor in the l.o. circuit. Thus if the station disappears (during a fade or at the end of a transmission) the receiver tuning remains at the last setting, at which time it was in exact tune.

108. Reactance valve type a.f.c. reduces the tuning error to approximately 0.01 of the initial error but note that there must always be some residual error (otherwise no correction voltage is obtainable from the discriminator to hold the set in tune). Note that with mechanical a.f.c. the motor can be set to drive until there is zero output from the discriminator (or if there is no OFF position for the motor it can be arranged to oscillate at a very slow rate about the correct tuning point). Thus the tuning accuracy is greater with this system.

A.F.C. as applied to transceivers

109. The frequency stability of the transmitter section of an f.m. field radio transceiver is not good and the use of a.f.c. somewhere in the radio link is essential. Several possibilities exist, ie either the transmitter or the receiver may use a.f.c. (or conceivably both transmitter and receiver could use a.f.c.). Some of the systems in current (and past) use will be considered in detail.

WS No 88

110. This equipment operates on four spot frequencies in the v.h.f. range. The receiver is crystal-controlled; the transmitter has a.f.c. and 'the a.f.c. loop' includes many of the receiver stages and in particular makes use of the receiver discriminator as an error detecting device.

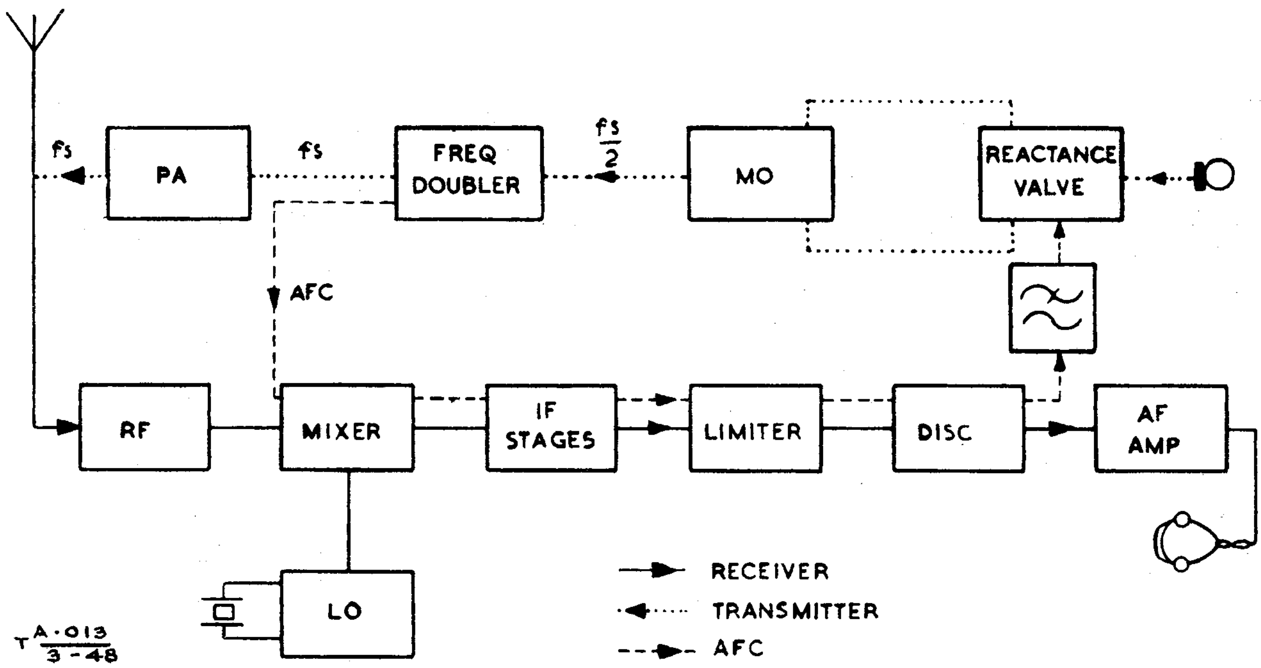


Fig 48 - A.F.C. - WS No 88

WS No 31

111. In this set the receiver frequency stability is controlled by a.f.c. but not the transmitter. However, the a.f.c. loop controls several transmitter stages which operate in the receiver. A.F.C. controls the frequency of the m.o. stage on receive, making use of the receiver discriminator as an error detecting device. As the ratio of receive to send is normally five to one or greater the frequency drift to the m.o. and hence the transmitter is not serious on short transmission. On long transmission the frequency drift is considerable and often contact is lost with the distant station on switching to receive, as the receiver has drifted beyond the capture range of the discriminator, owing to the narrow bandwidth of the i.f. channel (30kc/s at 6dB down).

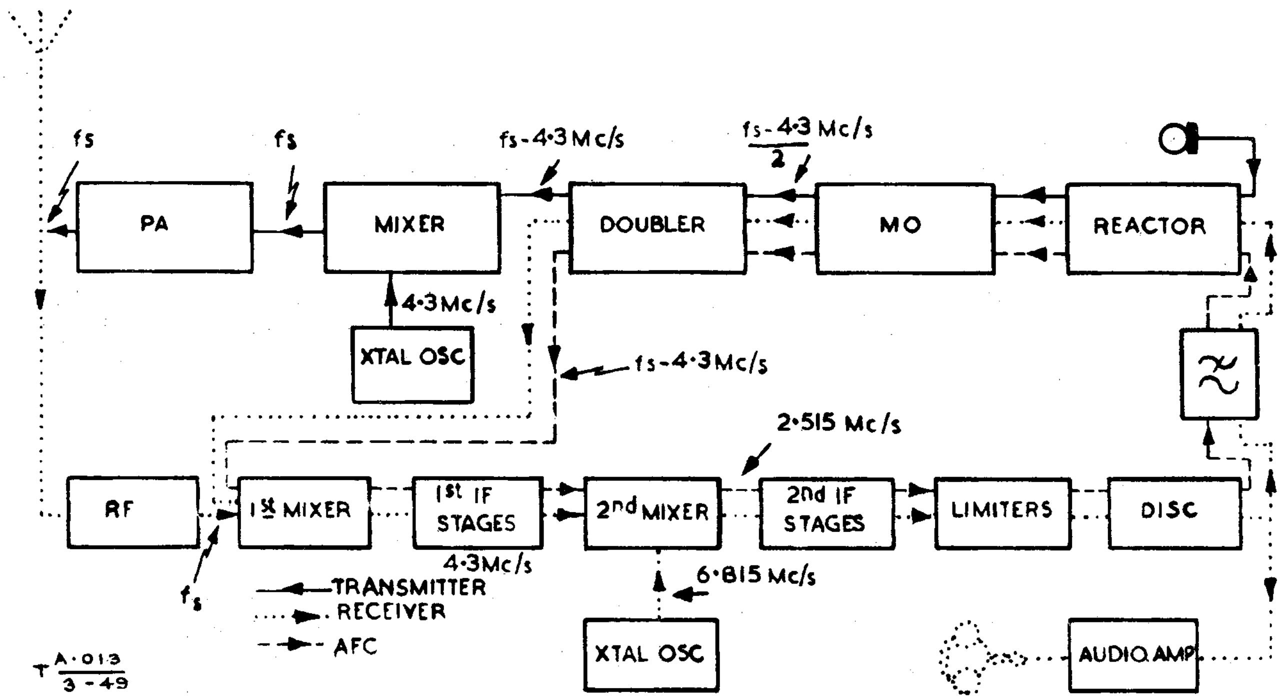


Fig 49 - A.F.C. - WS No 31

Radio sets CPRC-26 and WS A40

112. To counteract the frequency instability of the transmitter and the narrow capture range of the discriminators of such sets as the 31 and 88 special a.f.c. or wideband discriminators are used in the above sets. These extend the capture range to up to ±250kc/s which is sufficient to compensate for frequency drift on long transmissions.

113. These wideband discriminators have an inherent defect when used as frequency error detecting devices. Because of their wide range the slope of the a.f.c. voltage applied to the reactor is gradual so that the a.f.c near the centre frequency is not effective.

WS C42 and C45

114. These sets use two discriminators for the production of a.f.c. voltages, one a wideband and the other a narrow band discriminator. These have a capture range of ±500kc/s and ±50kc/s respectively. Fig 50 is a block diagram of the a.f.c. loop.

115. Portion of the transmitted signal is picked up by the receiver 1st r.f. stage and fed through to the 1st i.f. where it splits, one portion going to the a.f.c. discriminator and the other to the 2nd mixer and on to the a.f. discrminator. The d.c. output from the a.f. discrminator is fed through a delay network to prevent the audio frequency output of the discriminator from affecting the a.f.c. voltages.

116. The a.f. discrminator voltage output is in series with the a.f.c. discrminator and the combined control voltage is applied to the grid of the modulator valve.

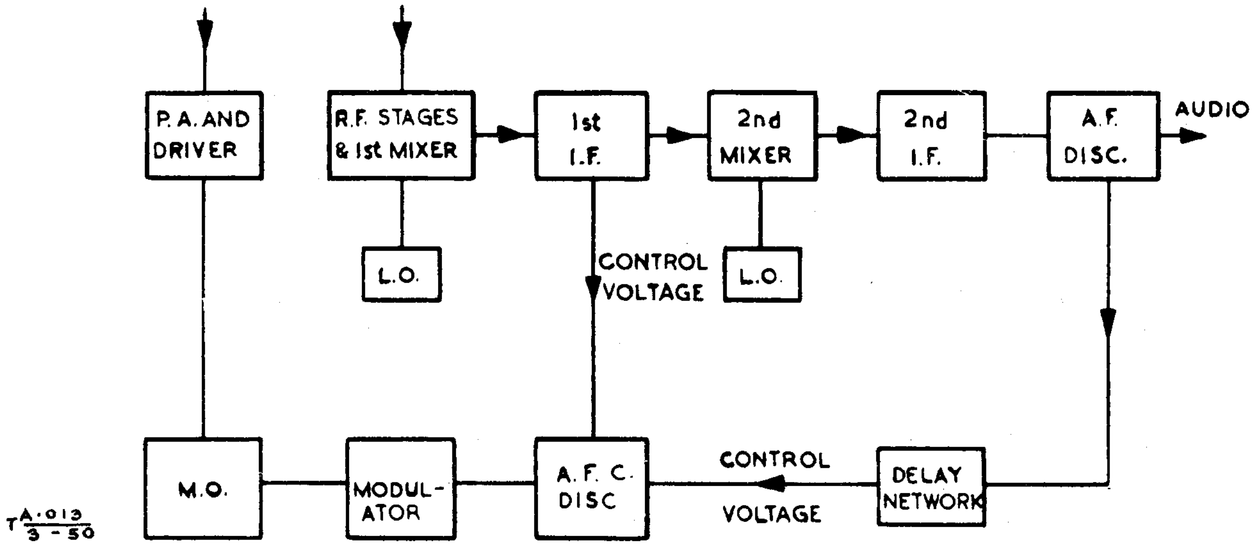


Fig 50 - A.F.C. - WS C42

117. Consider the requirements of the m.o. valve for frequency control voltages. As the transmitted signal increases in frequency a control voltage must decrease the frequency of the m.o. Primarily, the inductance of the coil in the frequency determining circuit of the m.o. must be increased to cause a decrease in the transmitted frequency. Now we have seen that the output of the a.f. discrminator (see Bond discrminator, para 84 and Fig 33) is negative when the transmitted frequency is above the operating frequency. Therefore the modulator plate current will decrease. This causes the inductance of the secondary of the ferrite reactor to increase. The overall inductance of the m.o. grid tuned circuit increases and the frequency decreases; thus the desired correction of frequency drift is achieved.

118. Referring to Fig 1001, simplified circuit of frequency control; assume that the transmitter is operating at 40Mc/s and that it drifts by 100kc/s so that the transmitter frequency is 40.1Mc/s. The 1st i.f. output will be 1st l.o. frequency minus signal frequency, ie 46 - 40.1Mc/s = 5.9Mc/s. This 5.9Mc/s signal passes to the a.f.c. discrminator where the lower frequency diode V12 conducts more heavily and the resultant voltage output applied to the grid of m.o. is negative, thus fulfilling the requirement for a negative control voltage to correct an increase in the transmitted signal frequency.

119. The 5.9Mc/s signal is also passed to the 2nd mixer where it is mixed with the 2nd l.o. output of 8.4Mc/s. The resultant which passes to the 2nd i.f. is 2.5Mc/s. This 2.5Mc/s signal passes to the a.f. discrminator via the limiters and causes the high frequency diode V18 to conduct more heavily than V19. The resultant output of the a.f. discrminator (read between the cathode of V19 and chassis) is negative. As the two discrminators are in series with regard to d.c. output their outputs are additive. If the transmitted signal frequency drifts below the operating frequency the operation of the discrminators reverses and the resultant voltage output applied to the modulator is positive.

120. A careful study of the two discrminators shows that they are reversed. This is illustrated in Fig 1001. This is necessary because of the two mixer stages which, in the 1st mixer changes an increase in transmitted signal frequency into a decreased signal frequency which activates the a.f.c. discrminator while in the 2nd mixer this decreased frequency is converted to a higher frequency signal.

Note: The next page is Page 1001

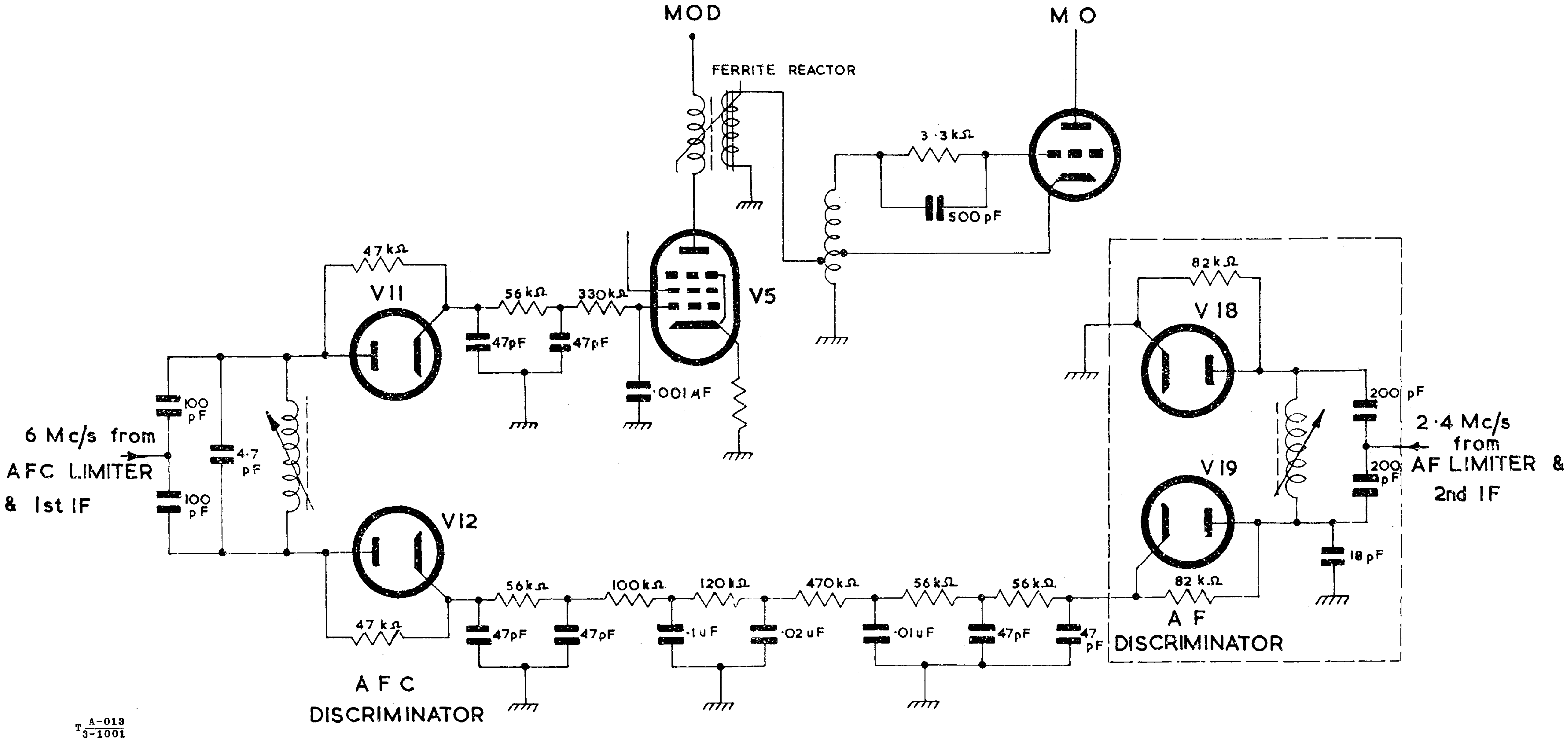


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Fig 1001 – Simplified a.f.c. circuit – WS C42 and C45

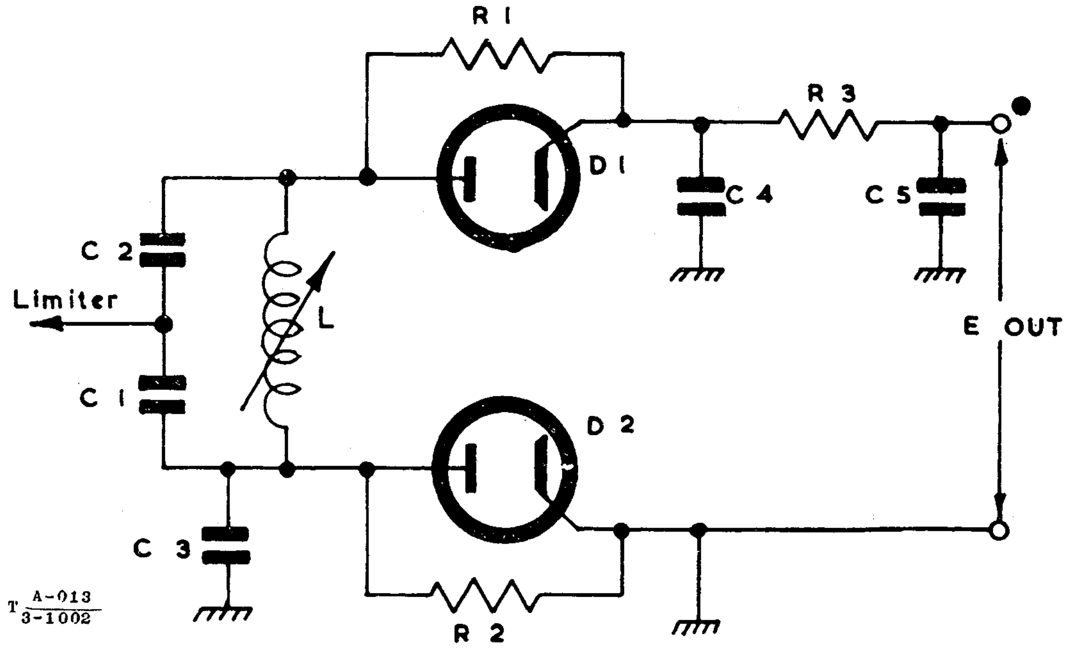

Fig 1002 - Bond discrmiminator circuit where increase of frequency gives positive output voltage

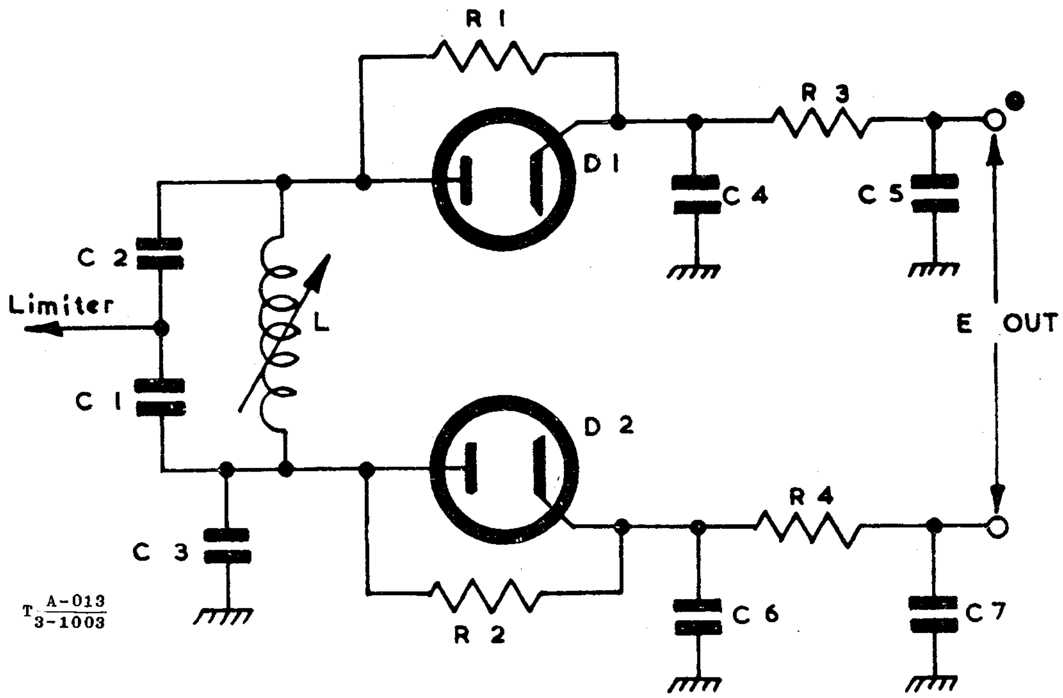

Fig 1003 - Bond discriminator where the d.c. output is isolated from earth

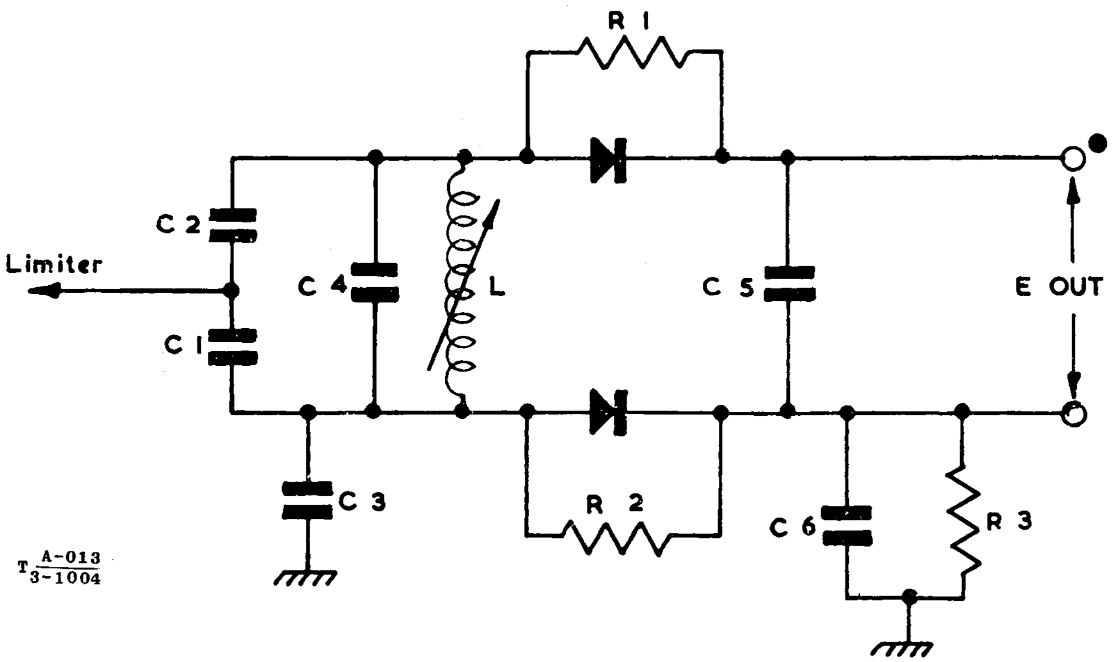

Fig 1004 - Bond discriminator circuit using germanium diodes

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