Limiters and Discriminators

I—Wide or Narrow Bandwidth : Basic Equivalent Circuits : The Round-Travis Discriminator

By ***G. G. JOHNSTONE, BSc.****

WE shall begin by considering the bandwidth required in a discriminator, more especially in view of the arguments put forward by Prof. Arguimbau and his colleagues in favour of wide-band discriminators.

If two f.m. signals interfere, they may be on the same carrier frequency (i.e., co-channel) or on nearby carrier frequencies (i.e., adjacent channel). Briefly Arguimbau maintains that in the presence of co-channel or adjacent-channel interference, the interfering signal will produce unwanted audio output even when the difference between the frequencies of wanted and unwanted signals is above the audible limit. The two signals add together to produce a composite signal which varies in amplitude and frequency according to the relative magnitudes and frequencies of the two signals. This is

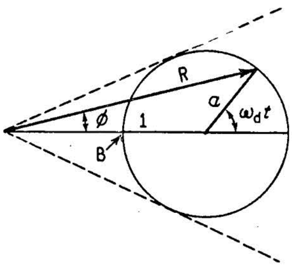

Fig. 1. Vector diagram of wanted signal (1) and interfering signal (a) with a small frequency difference.

shown in the vector diagram of Fig. 1, where the ratio of the amplitude of the unwanted signal to that of the wanted signal is $a:1$. The wanted-signal vector is assumed stationary and the unwanted-signal vector rotates at a relative angular frequency ω_d, which corresponds to the difference f_d between the frequencies of the two signals.

The magnitude of the resultant, R, varies between $(1 + a)$ and $(1 - a)$; this amplitude modulation can be removed by limiting and need not concern us further. The phase displacement (ϕ) of the resultant swings between the limits indicated by the dotted lines in Fig. 1. The phase displacement can be calculated and, since the instantaneous frequency shift is equal to $1/2\pi$ times the rate of change of phase angle, i.e., $f = \omega/2\pi$ and $\omega = d\phi/dt$ we can determine the frequency shift at any instant. The calculation is straightforward, but rather long, and it is sufficient for our purpose to note that the interfering frequency-modulated signal has the form shown in Fig. 2. The peaks at α occur when the vectors are in line as at A in Fig. 1; the peaks at β occur when the vectors have the position B in Fig. 1. The peak value of the frequency shift at α is given by $af_d/(1 + a)$ and at β by $af_d/(1 - a)$. Thus as the ratio of unwanted signal amplitude approaches that of the wanted signal a approaches 1 and the peak shift β becomes very large. The positive-going and negative-going excursions of the waveform of Fig. 2 have a different form but the areas included between the curves and the horizontal axis are equal, the mean resultant frequency shift is zero and there is no d.c. component in the output of the discriminator, provided that the discriminator is linear over the range of frequency excursion.

The period of the wave of Fig. 2 is $1/f_d$ and, if the frequencies of the two signals differ by more than the highest audio frequency, there will be no audible output under the steady state conditions postulated. If the frequency difference is less than the highest audible frequency, there is a heterodyne whistle; this is true whether the discriminator is " wide-band " or not.

To carry the argument further, suppose the difference between the frequencies of the two signals exceeds the highest audible frequency and the discriminator is narrow band. As the unwanted-signal amplitude approaches that of the wanted signal the ratio a tends to unity and the peak frequency shifts become progressively larger, tending to a limit of $f_d/2$ in one direction and infinity in the other. We shall concentrate on the peak at β because this is always greater than that at α. For an idealized narrow-band discriminator characteristic of the type shown in Fig. 3, the peaks at β are " clipped " as shown in Fig. 4. The areas under the positive-

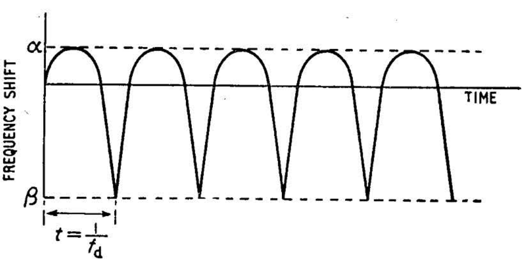

Fig. 2. Form of frequency modulation produced as a result of the conditions shown in Fig. 1.

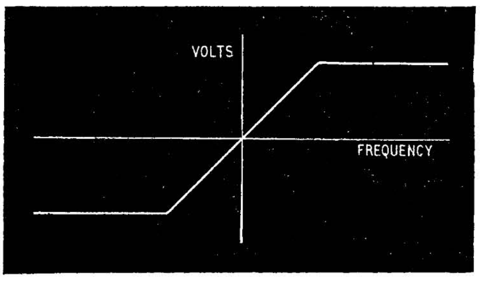

Fig. 3. Idealized narrow-band discriminator characteristic.

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for F.M. Receivers

A GREAT deal has been written about the performance of various types of limiter and discriminator for use in f.m. receivers, but much of this information is contradictory and the choice of a limiter and discriminator for a particular application may appear difficult, particularly to the newcomer to the subject. It is the purpose of this and subsequent articles to discuss the performance of the basic types of limiter and discriminator in an attempt to clarify the subject.

The discriminators to be discussed fall into six classes: (1) the Round-Travis (2) the Foster-Seeley discriminator (3) the ratio detector (4) the locked oscillator (5) the phase-difference comparator and (6) the counter. Types (3) (4) and (5) have a degree of inherent limiting action.

The types of limiter to be discussed comprise (1) the grid limiter (2) the anode limiter (3) the dynamic limiter and (4) the "clipper."

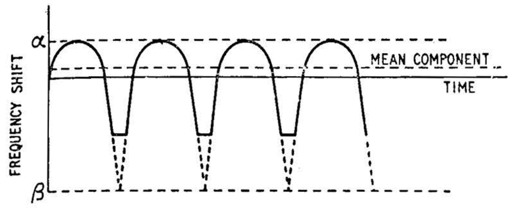


Fig. 4. Clipping of the frequency peaks by the characteristic of Fig. 3 results in a displacement of the average frequency.

going and negative-going portions of the waves are now no longer equal. There is a shift in the average frequency and a d.c. component in the output from the discriminator; it is shown in dotted lines in Fig. 4. If now the unwanted signal is modulated, its frequency swings about the mean value and the difference between the frequencies of the two signals alters correspondingly. The waveform of Fig. 2 now becomes that shown in Fig. 5(a); for comparison the graph of the frequency difference is shown at Fig. 5(b). If now the waveform of Fig. 5(a) is "clipped" as indicated by the dotted line, the d.c. component has superimposed "blips" as shown in Fig. 5(c). These have the same repetition rate as the modulating signal, and have components at multiples of the modulating-signal frequency. This represents cross-modulation between the two received signals accompanied by distortion. With a "wide-band" discriminator there is no clipping and this effect does not occur.

We may summarize the conclusions of the previous paragraphs as follows. In receiving co-channel signals some interference is inevitable because the difference between the frequencies of the two signals must, for part of the time, lie within the audio frequency range; this effect occurs whatever type of discriminator is employed. The peak heterodyne output is given by $af_d/(1 - a)$ which decreases rapidly as a falls below unity and the stronger signal rapidly "swamps" the weaker as the ratio of the amplitudes of the signals a departs from unity; this is, of course, the well-known "capture" effect.

For signals on adjacent channels the difference between the frequencies of two carriers is always greater than the highest audible frequency; there is no cross-modulation if a wide-band discriminator is employed. This is illustrated in Fig. 6 which

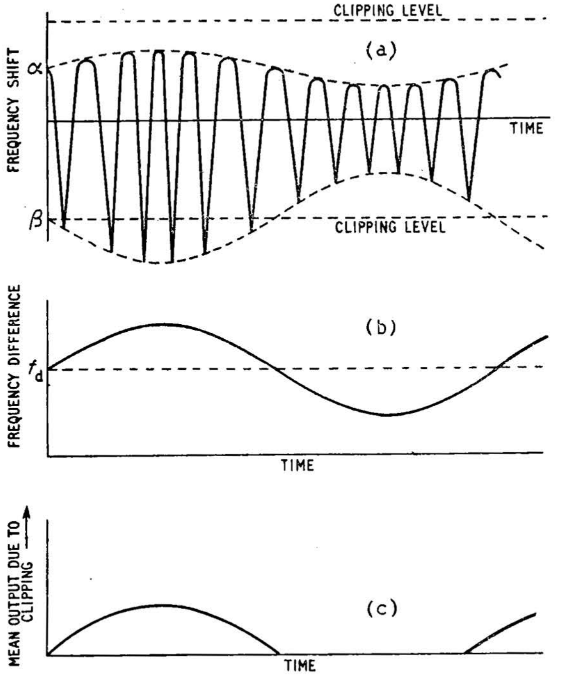


Fig. 5. (a) Waveform when the unwanted signal is modulated by a sine wave. The frequency difference is shown in (b), and the distortion product due to clipping is shown at (c).

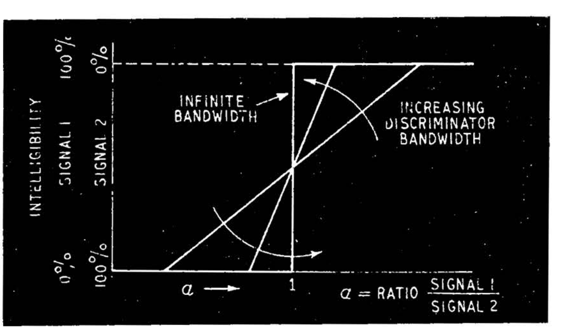


Fig. 6. Interference between signals on adjacent channels as a function of discriminator bandwidth.

shows that for an infinite bandwidth only the stronger signal is received even if its amplitude is only slightly greater than that of the other; but as the bandwidth is made smaller, the ratio of signal strengths for which interference is experienced becomes greater.

The behaviour of a discriminator toward ignition interference can be deduced by similar arguments.

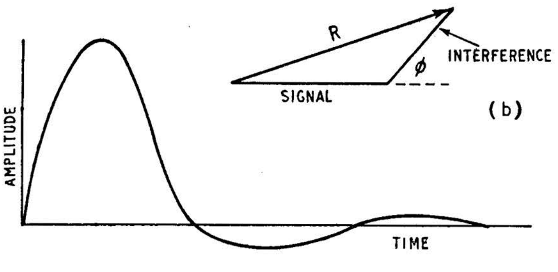


Fig. 7. (a) Waveform of i.f. output due to a pulse of interference and (b) vector relationship of signal and interference.

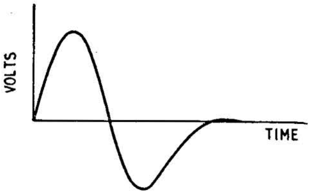


Fig. 8. Frequency shift (and hence discriminator output) due to interference of the form of Fig. 7(a).

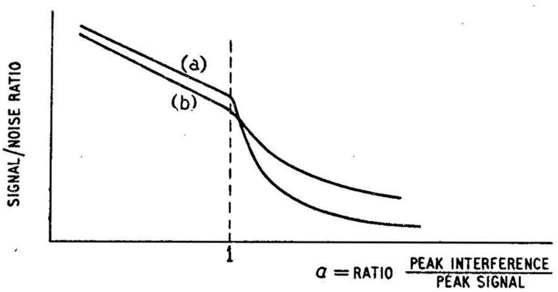
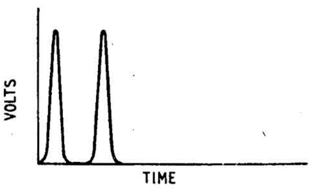


Fig. 9. Signal/noise ratio with impulsive interference for (a) wide-band and (b) narrow-band discriminators.

Fig. 10. Unidirectional double pulse produced in the discriminator output in some cases when the interfering pulse amplitude exceeds that of the carrier.



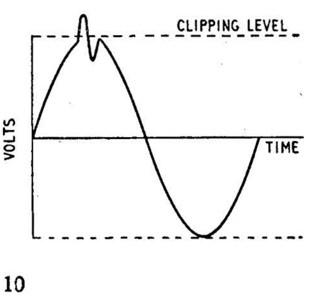


Fig. 11. When the interference pulse coincides with a peak of modulation, clipping may remove part of the pulse and increase noise.

The waveform of a pulse of interference is modified in the i.f. amplifier to a shape similar to that shown in Fig. 7(a). The shape of this pulse varies little with the source of the interference, but depends almost entirely upon the characteristics of the i.f. amplifier itself. The duration of the pulse is very short, approximately 7 microseconds for an i.f. bandwidth of 200 kc/s. The figure shows the envelope, which modulates a " carrier " at the centre frequency of the i.f. amplifier, which we shall assume equal to the signal carrier frequency. The signal carrier and interference carriers may have any phase angle with respect to each other, because the exact moment of incidence of the interference pulse is fortuitous. Thus the signal and interference carriers add as shown in Fig. 7(b); the phase angle between the carrier vectors ϕ has been arbitrarily selected. The interference signal vector grows and decays as shown in Fig. 7(a). Consequently, if the ratio of peak interference envelope to the peak carrier is small, the phase angle of the resultant follows a wave similar in shape to that shown in Fig. 7(a). The rate of change of phase angle, and hence the frequency shift then follow a curve of the type shown in Fig. 8. When the number of tuned circuits preceding the limiter exceeds four, the maximum value of the frequency shift due to the interfering signal is given approximately by $4aF$ where a is the ratio of the peak value of the interfering signal to that of the carrier, and F is the bandwidth measured between 3-dB points. The audible noise output is proportional to the net area under the frequency-shift envelope, and hence to a. However, the area under the positive-going excursion is nearly equal to that under the negative-going excursion and the net area is small. The spectrum of this output rises linearly with frequency, and the output has the sound of a " click." If a succession of constant-amplitude impulses are presented to the receiver, the magnitude of the output pulses varies with the relative phase angle between the carrier and the interfering signal at the occurrence of each impulse. If the amplitude of the interference signal is less than that of the carrier, a is less than unity and the signal-to-noise ratio falls linearly with a, as shown in curve (a) of Fig. 9. If the amplitude of the interfering signal exceeds that of the carrier, a is greater than unity and a new mechanism comes into play by which some of the output-signals have the form of sharp spikes, each pulse of ignition interference producing two spikes of the same polarity, as shown in Fig. 10. The net area beneath the curve is now no longer small. The spectrum of this waveform is level over the a.f. band and the output has the sound of a " pop." For such high-amplitude interfering signals, as the amplitude of the interfering signal is increased the signal-to-noise ratio falls abruptly and then levels off as shown in curve (a) of Fig. 9.

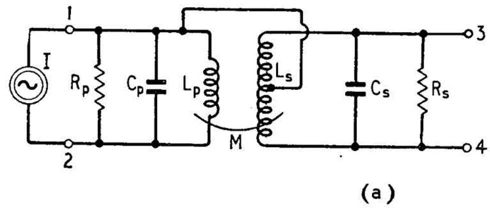
Consider the effect of reducing the discriminator bandwidth on impulsive interference of this type. If a is less than unity, and the signal is at one extreme of its frequency swing, one of the peaks due to impulsive interference in the discriminator output signal may be clipped as shown in Fig. 11. The net area increases and the signal-to-noise ratio is poorer than for a wide-band discriminator as shown in Fig. 9 curve (b). On the other hand, in the region where a is greater than unity, a reduction in discriminator bandwidth causes the amplitude of the spikes (Fig. 10) to be reduced and the signal-to-noise ratio is better than for a wide-band discriminator. Thus for impulsive interference a narrow-band discriminator is an advantage because the worsening of the signal-to-noise ratio in the region where this ratio is good in exchange for an improvement in the area where it is bad is in general desirable. Thus we may summarize the conclusions so far reached as follows:

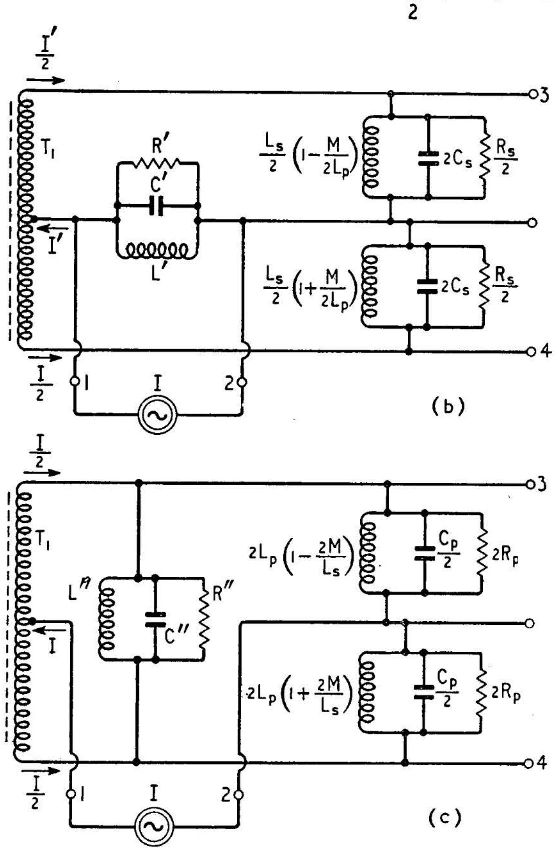
(1) If the principal source of interference is co-channel and adjacent channel, a wide-band discriminator is best.

(2) If the principal source of interference is of the impulsive type, a narrow-band discriminator is best.

In this country co-channel and adjacent channel interference would not appear to present a serious problem, and hence ignition interference may be the principal type of interference to be overcome. In U.S.A., however, ignition interference is probably less troublesome than co-channel and adjacent-channel interference. Thus it would appear that British and American practice in discriminator bandwidth may tend to diverge, with a tendency in Britain to use narrow bandwidth discriminators.

Fig. 12. The basic phase-difference discriminator transformer (a) may be represented by either of the equivalent circuits (b) or (c).





$$L' = \frac{L_p L_s}{L_s - 4L_p}$$
$$C' = C_p - 4C_s$$
$$R' = \frac{R_p R_s}{R_s - 4R_p}$$

T_1 IS CENTRE-TAPPED "IDEAL" TRANSFORMER

$$L'' = \frac{L_p L_s}{4L_p - L_s}$$
$$C'' = C_s - \frac{C_p}{4}$$
$$R'' = \frac{4R_p R_s}{4R_p - R_s}$$

We shall now consider the performance of discriminators in detail and in the subsequent discussion we shall make extensive use of one or other of two equivalent circuit diagrams for the phase-difference discriminator transformer, Fig. 12(a), used in the Foster-Seeley discriminator and the ratio detector. These equivalent diagrams are shown in Figs. 12 (b) and (c), and their derivation is given in the Appendix. Depending on the constants of the original circuit only one of the two equivalents is physically realizable. The equivalent circuit for one special case reduces to two tuned circuits fed with equal currents, and resonant at frequencies equally displaced from the centre frequency of the transformer from which they are derived. Thus the Foster-Seeley and Round-Travis circuits shown in Fig. 13 have identical performances. This circuit transformation simplifies the analysis of the Foster-Seeley discriminator and the ratio detector and we shall analyse the performance of the Round-Travis circuit in detail in order to extend the results to the Foster-Seeley and ratio detector circuits.

Round-Travis Circuit. In this discriminator two

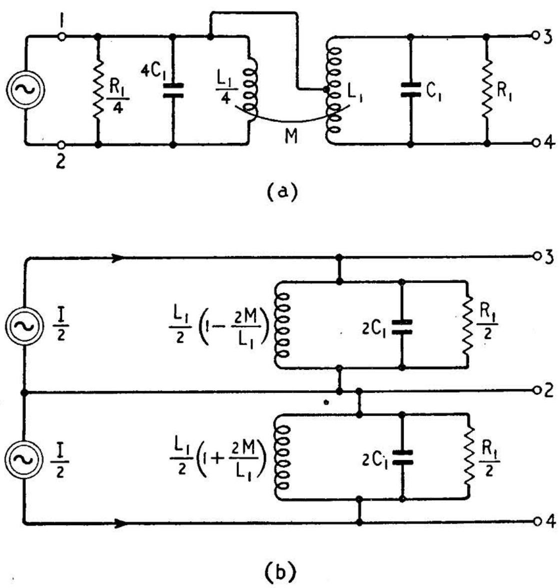


Fig. 13. Identical performances are given by these forms of (a) the Foster-Seeley and (b) the Round-Travis circuits

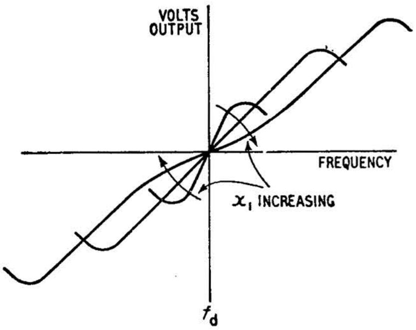


Fig. 14. Dependance of shape of discriminator characteristic on the value of x_1.

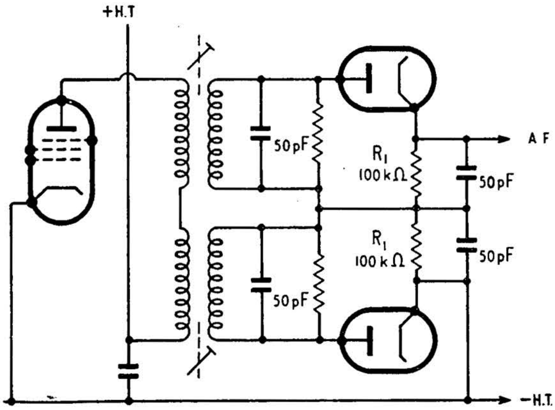


Fig. 15. Basic practical Round-Travis circuit.

tuned circuits are employed, one tuned to a frequency slightly above the centre frequency and the other tuned to a frequency an equal interval below the centre frequency. This circuit has appeared in a wide variety of forms, the variations being principally in the method of applying the signals to the two circuits. We shall consider the circuit in its general form, where the driving stage is represented by a constant current generator. as shown in Fig. 13.

We can write the response of a tuned circuit, comprising L, C, and R in parallel as

$$\frac{1}{Z} = \frac{1}{R} + \frac{1}{j\omega L} + j\omega C$$

In the frequency range near resonance (ω_0) we can write ω as ($\omega_0 + d\omega$), where $d\omega$ is small compared with ω_0. Using this substitution we can make the approximation that

$$\frac{1}{Z} = \frac{1}{R} + 2jd\omega C = \frac{1}{R} + 2j\frac{d\omega}{\omega_0} \cdot \frac{1}{\omega_0 L}$$

$$= \left(1 + \frac{2jd\omega}{\omega_0} \cdot \frac{R}{\omega_0 L}\right)/R$$

If we let $x = 2Qd\omega/\omega_0$, where Q is $R/\omega_0 L$, then

$$Z = R/(1 + jx)$$

in which x is a variable directly proportional to the frequency shift.

If one tuned circuit is tuned above the centre frequency and displaced by an amount $x = x_1$ and the other is tuned below and displaced by an amount $x = -x_1$ we can represent the impedance of the two circuits at any frequency in the neighbourhood of the centre frequency by $Z_1 = R/[1 + j(x - x_1)]$ and $Z_2 = R/[1 - j(x + x_1)]$ respectively.

The two diodes rectify the voltages developed across each tuned circuit independently, as shown in Fig. 15, and for 100% rectification efficiency, the output voltage is equal to the difference between the peak values of the voltage developed across each tuned circuit. If I is the peak value of the input current the output voltage is given by

$$E = IR[\{1 + (x - x_1)^2\}^{-\frac{1}{2}} - \{1 + (x + x_1)^2\}^{-\frac{1}{2}}]$$

A family of curves for various values of x_1, is shown in Fig. 14. These are the output-voltage frequency-shift curves for the discriminator, and ideally should have a linear region about the centre frequency. As shown in Fig. 14, for small values of x_1 the characteristic has continuous curvature and small peak separation. Larger values of x_1 give better linearity and greater peak separation, but above a critical value of x_1 the characteristic develops a kink near the centre frequency. Thus there would appear to be an optimum value of x_1.

The expression for E given above can be expanded as a series in ascending powers of x. The expansion is symmetrical about $x = 0$, and contains odd powers of x only. We can thus write

$$E = a_1 x + a_3 x^3 + a_5 x^5 \ldots$$

The first term represents the required output, i.e., an output varying linearly with frequency shift. The other terms indicate non-linearity, producing harmonic distortion and intermodulation in the output. This distortion can be minimized by making a_3/a_1, a_5/a_1, etc., as small as possible.

The evaluation of a_1, a_3, a_5 in terms of x_1 is straightforward but tedious. The values of the first three coefficients are:

$$a_1 = 2x_1(1 + x_1^2)^{-\frac{3}{2}}$$

$$a_3 = x_1(2x_1^2 - 3)(1 + x_1^2)^{-\frac{7}{2}}$$

$$a_5 = \tfrac{1}{4}x_1(8x_1^4 - 40x_1^2 + 15)(1 + x_1^2)^{-\frac{11}{2}}$$

From these expressions we have

$$a_3/a_1 = \tfrac{1}{2}(2x_1^2 - 3)/(1 + x_1^2)$$

and

$$a_5/a_1 = \tfrac{1}{8}(8x_1^4 - 40x_1^2 + 15)/(1 + x_1^2)^2$$

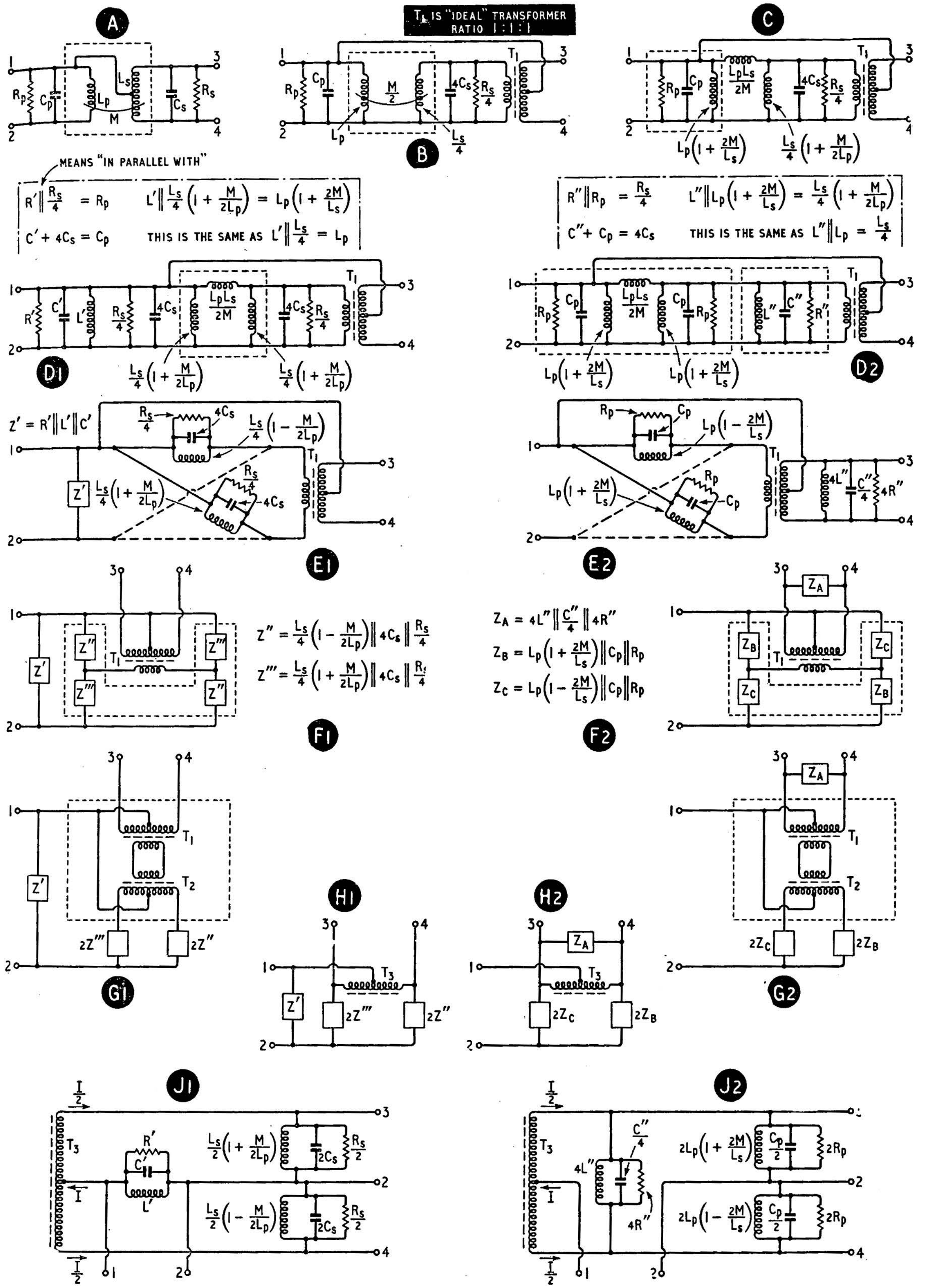
To minimize distortion, we can choose the value of

APPENDIX

THE derivation of the equivalent circuit for a phase difference discriminator transformer, Fig. A on the opposite page, is obtained by the successive manipulation of the portions enclosed by dotted lines. In Fig. B the secondary circuit comprising L_S, C_S, R_S is transferred to the primary side of the "ideal" transformer T1; this transformer has a centre-tapped secondary, providing the two voltages of opposite polarity in series with the primary voltage. In Fig. C, the "T" network is converted to "π" form. In Fig. D1 the terminations of the "π" network are made equal. If the terminations are made equal to the secondary components, the sequence follows through Figs. D1 to H1, and if to the primary components through Figs. D2 to H2.

In Fig. E, the "π" network is translated to a lattice by means of Bartlett's Bisection Theorem. Figs. F show Figs. E re-drawn. Figs. G show Figs. F re-arranged by the introduction of transformer T2, also of 1 : 1 · 1 ratio. Figs. H are Figs. G with transformers T1 and T2 combined into transformer T3. Figs. J show Figs. H re-drawn.

The derivation of these equivalent circuits is an extension of the treatment due to H. Marko (*Frequenz*, January, 1952).



$R' \parallel \frac{R_s}{4} = R_p \qquad L' \parallel \frac{L_s}{4}\left(1 + \frac{M}{2L_p}\right) = L_p\left(1 + \frac{2M}{L_s}\right)$

$C' + 4C_s = C_p \qquad$ THIS IS THE SAME AS $L' \parallel \frac{L_s}{4} = L_p$

$R'' \parallel R_p = \frac{R_s}{4} \qquad L'' \parallel L_p\left(1 + \frac{2M}{L_s}\right) = \frac{L_s}{4}\left(1 + \frac{M}{2L_p}\right)$

$C'' + C_p = 4C_s \qquad$ THIS IS THE SAME AS $L'' \parallel L_p = \frac{L_s}{4}$

$Z' = R' \parallel L' \parallel C'$

$Z'' = \frac{L_s}{4}\left(1 - \frac{M}{2L_p}\right) \parallel 4C_s \parallel \frac{R_s}{4}$

$Z''' = \frac{L_s}{4}\left(1 + \frac{M}{2L_p}\right) \parallel 4C_s \parallel \frac{R_s}{4}$

$Z_A = 4L'' \parallel \frac{C''}{4} \parallel 4R''$

$Z_B = L_p\left(1 + \frac{2M}{L_s}\right) \parallel C_p \parallel R_p$

$Z_C = L_p\left(1 - \frac{2M}{L_s}\right) \parallel C_p \parallel R_p$

x_1 which makes one of these ratios zero; it is usual to make $a_3 = 0$, because this is the dominant distortion-producing term for values of x less than unity. This gives $x_1 = \sqrt{1.5}$, and the expression for output voltage becomes

$$E = IR(0.62x - 0.055x^5 \ldots)$$

For this value of x_1, the useful extent of the characteristic is limited to values of x of less than unity. This is illustrated in the calculation below where it is shown that the distortion rises steeply for greater values of x.

To evaluate the distortion likely to be obtained with this circuit arrangement, we shall consider an input signal frequency-modulated by a signal of frequency f_a to a maximum swing of f_s kc/s on each side of the centre frequency. The value of x corresponding to f_s is represented by x_s and is given by $x_s = 2Qf_s/f_0$. Substituting the value $x_s \cos\omega_a t$ for x in the expression above gives

$$E = IR\,(0.62x_s \cos\omega_a t - 0.055x_s{}^5\omega_a t \ldots)$$

We can expand $\cos^5\omega_a t$ by means of the identity

$$\cos^5\theta = \tfrac{1}{16}(\cos^5\theta + 5\cos^3\theta + 10\cos\theta)$$

which gives

$$E = I\,R[0.62x_s \cos\omega_a t - 0.0034x_s{}^5(\cos^5\omega_a t + 5\cos 3\omega_a t + 10\cos\omega_a t)]$$

$$= IR\,[(0.62x_s - 0.034x_s{}^5)\cos\omega_a t - 0.017x_s{}^5\cos 3\,\omega_a t - 0.0034\,x_s{}^5\cos^5\omega_a t)]$$

The amplitude of the fundamental-frequency component is less for an ideal characteristic but the reduction is small and will be ignored, the values of x_s being limited to approximately unity. The percentage of third-harmonic distortion is then given by $\frac{0.017 \times 100}{0.62}x_s{}^4$, and the percentage of fifth harmonic is one-fifth of this figure. For $x_s = 1$, there is 2.7 per cent third-harmonic and 0.54 per cent fifth-harmonic distortion. The magnitude of the harmonic distortion is proportional to $x_s{}^4$, and thus falls rapidly if a lower value of x_s is considered. Thus reduced distortion can be obtained by using a lower value of x_s, but this in turn means a smaller output at the fundamental frequency.

With a broadcast signal, the frequency deviation is fixed at 75 kc/s. If it is desired to operate the discriminator with 75 kc/s corresponding to $x_s = 1$, the parameters of the circuit are determined by the relationship $x = 2Q\ df/f_0$. With $df = 75$ kc/s at $x = 1$, and a centre frequency (f_0) of 10.7 Mc/s, the value of Q is 71. The resonant frequencies of the two tuned circuits are given by 10.7 $(1 \pm \sqrt{1.5}/142)$ Mc/s, i.e., 10.7 Mc/s $\pm$ 92 kc/s. Such a discriminator would have very little margin to allow for oscillator-frequency drift and mis-tuning and a value $x_s = 1$ corresponding to 100 kc/s would be better. The circuit parameters are then Q = 53 with the resonant peaks at 10.7 Mc/s $\pm$ 123 kc/s. For such a discriminator at 75 kc/s deviation the third harmonic distortion is 0.84 per cent and the fifth harmonic 0.17 per cent. If we assume that the two tuned circuits each employ a capacitor of 50 pF, the dynamic resistance, $R = Q/\omega C$ of the circuits is approximately $53 \times 300 = 16\ k\Omega$. The input current, I, is the peak value of the fundamental-frequency component in the output of the preceding limiter stage; a typical value is 1 mA. The peak audio output is given by $0.62\,IR\,x_s$ and in the example chosen, x_s for 75 kc/s deviation is 0.75, giving a peak audio output of approximately 7.5 volts. The inductance required to resonate at 10.7 Mc/s with 50 pF is 4.45 μH; final adjustment of the resonant frequencies of the two circuits is made by means of dust-iron cores in the inductor formers. To secure the correct value of Q, the usual procedure is to design the coils for a higher value of Q than required and add damping resistors. The diode detectors provide part of this damping equivalent approximately to $R_1/2$, where R_1 is the value of each diode load resistor. The basic practical circuit is shown in Fig. 15; coupling windings to the two inductors are employed to isolate the tuned circuits from the h.t. supply. In practice, additional precautions are necessary to eliminate the effect of the primary circuit capacitance.

The Round-Travis circuit offers no real protection against amplitude modulation. The audio output is proportional to Ix, where I is the input current, and x is a measure of the frequency shift. If there is amplitude modulation the magnitude of I varies, but if $x = 0$, i.e., if the signal is at the centre frequency there is no output due to a.m. For any other value of x, i.e., if the signal is mistuned or frequency-modulated, there is an output due to the amplitude modulation. Because the output is proportional to Ix the a.m. and f.m. signals are multiplied together and there is complete cross modulation. Thus a circuit of this type must be preceded by a limiter stage.

(*To be continued.*)

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Residual Magnetism in Recording Heads

MOST tape recorders incorporate a long-time-constant smoothing circuit in the h.t. supply to the bias oscillator to ensure that the h.f. current in the record/replay head dies away slowly when the instrument is switched off. This is necessary because any "d.c." component of remanent magnetism in the head is known to cause an increase in background noise from the tape.

Unfortunately, the amplitude of h.f. bias for best results from the point of view of either low distortion or high recording level is much less than is necessary to drive the core to a state of magnetic saturation, so that if by any mischance the magnetic state is carried beyond the maximum represented by the bias the head will not be automatically demagnetized.

Fortunately, the head can easily be demagnetized by bringing up a strong (saturating) external alternating field and then removing it slowly. This field can conveniently be provided by a small 50-c/s transformer with an air gap arranged to coincide with the gap in the recording head, but it is not too easy to devise suitable means with conventional components, due to the smallness and inaccessiblity of most recording heads.

Wright and Weaire have recently developed a "defluxer" for this purpose in which projecting poles are arranged to give easy contact with the face of the record/playback head. The transformer is housed in a cylindrical case which falls conveniently to hand, and a press-button switch is provided for operation.

The "de-fluxer," which can also be used for selective erasure when editing tape records, costs £2 10s.

LETTERS TO THE EDITOR

The Editor does not necessarily endorse the opinions expressed by his correspondents

"*Limiters and Discriminators for F.M. Receivers*"

MR. G. G. Johnstone has raised an interesting issue in his article in the January, 1957, issue. He agrees that the use of wide-band limiter and discriminator circuits is advantageous in suppressing small amounts of f.m. interference, but says of a narrow-band detector: "in the region where *a* is greater than unity, a reduction in discriminator bandwidth causes the amplitude of the spikes to be reduced and the signal-to-noise ratio is better than for a wide-band discriminator."

He may like to know that we made no very subtle choice between the relative nuisance values of co-channel and ignition interference, but rather failed to consider the point he has raised.

It is a very difficult matter to predict the response of a narrow-band detector to wide rapid changes in frequency. Some of us made static analyses of the type indicated in his Fig. 4, but found that although the method gives results that are qualitatively right the dynamics of the tuned circuits are such that experimental results differ considerably from the static analysis. However, I agree that for impulses that are slightly larger than the signal the narrow-band detector should be expected to give better results than the wide-band, particularly when the impulses simultaneously exceed the signal by small amounts and occur at a moment of high deviation. On the other hand, for impulses considerably larger than the signal, high spikes do not ordinarily result, and the two receiver types should give about the same results. Thus I feel that the narrow-band circuit would have advantages over the wide-band for only a small range in which ignition noise exceeds signal strength by a small margin and at high deviation; for all other cases the wide band should be equal or better.

Actually this feeling of mine is an over-simplification, since the maximum spike height depends upon signal/noise ratio, upon instantaneous phase, and upon the instantaneous frequency deviation present at the time the disturbance occurs. It would seem that the results calculated for some statistical distribution of all these factors would be by no means so simple as Mr. Johnstone and I have assumed. My own guess is that under these conditions the wide-band circuits may show up more favourably than would at first seem likely.

Mr. Johnstone's suggestion should also be of interest in the suppression of the effects of random noise. When a wideband detector operates on the combination of an f.m. signal and random noise the observed results are similar to those with ignition noise. For large values of S/N the output is clean, having at worst some "clicks." When the r.m.s. noise approaches within a few decibels of the signal, random "pops" of identical form are observed. The number of these pops per second increases rapidly as the value of S/N approaches 1. This experimental result is readily explained by considering the amplitude-modulated character of restricted-band noise, but I am not aware of any numerical statistical analysis of "pop" probability, an analysis that is difficult but should be worth carrying through.

In U.S. practice most receivers use narrow-band detectors. Many of them are inexpensively built and do not have flat i.f. responses, but may fall off by 5 or even 10 dB at the edges of the band. Further, many fail to limit properly on noise. When a fully modulated signal of small amplitude is impressed on such a receiver the response falls below limiter control and noise on the peaks. As a result these receivers tend to distort badly on peaks not only because of spike clipping but for more earthy reasons as well. Receivers using wideband detectors are usually more carefully made in these routine ways, and it is not easy to say whether their observed superiority over narrow-band receivers in suppressing random noise is due to routine care or to their inherent properties. More experimental work is needed.

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Audio Output Power

IT is to be hoped that P. J. Baxandall's amplifier design, published in your March issue, will herald a return to sanity in the field of domestic sound reproduction. His contention that an output of 5 watts is adequate in the home will be supported by many.

Large numbers of existing 10-15-watt amplifier designs are used with speakers which, even with acoustic loading, can handle little more than 5 watts. For these, the use of a lower-power amplifier would merely mean operation near the limit of the gain control setting, probably for the first time.

Percy Wilson recalls, in the same issue of *Wireless World*, that G. A. Briggs surprised many in the Festival Hall by the low readings given by his output power indicators. If my memory serves, the 5-watt mark was not often passed and during full orchestral reproduction peaks of 40-50 watts were indicated only occasionally. Further, I well remember that a demonstration was at that time given of reproduction from an 8-in acoustically loaded speaker which, without noticeable distortion but admittedly with some loss of realism compared with the multi-speaker system in use immediately beforehand, managed to produce a very fair sound intensity at a point well back in the hall. Had the music contained the largest peaks this speaker might well have overloaded at the existing gain control setting; I merely wish to suggest that power close to 5 watts was giving effects at the ear not vastly different from that produced by live music under identical conditions.

If modest power can give good results in the Festival Hall, surely a comparable figure should provide all the reserve required for peak reproduction in the average living room?

R.A.E., Farnborough. W. E. DEAN.

Stereophonic Broadcasting

I AGREE with G. H. Russell (Jan. issue) that v.h.f. transmission offers an unprecedented opportunity for giving sound broadcasting a new lease of life, but I do not think that extension of h.f. response in a single channel is the best way of using the available bandwidth.

The human auditory mechanism is very intricate and is normally based on the transduction of separate stimuli at each cochlea into electrical waveforms, then into nerve pulses which are projected at both sides of the cortex. It is upon the relationship between these *two* different sources that our mind primarily depends for an appreciation of the spatial qualities of sound, and when this relationship is properly established our hearing is "contented"—much more so than by the "hi-fi" frequencies. As J. Moir has reminded us (Nov., 1956, issue, p. 543) "the frequency range that is produced seems