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"T"-STUB CALCULATION FOR VHF TRANSMISSION LINE FILTERS

By M. Telford, B.Sc.(Eng.).

The method of "T"-stub calculation presented below provides an alternative to the existing method when the stubs in a filter are true reciprocals but is of particular use when this condition does not apply and the normalized input reactances of separate stubs are not equal to unity at the "cross-over" frequency. It is also applicable to cases where another element, such as an isolating stub, is combined with a "T"-stub and the combination must have a given frequency response.

"T" connected stubs

"T" connected stubs[1] are two terminal networks built up entirely of lengths of transmission line. Their purpose is to provide very high and very low impedances at closely adjacent frequencies, with adjustment of the mean impedance level. As such they find many uses in filter applications where pass and stop bands of varying widths are required, e.g. in television transmission combining filters and in notch vestigial sideband filters. The exact equivalent circuits of direct and reciprocal "T"-stubs are shown in Figs. 1(a) and 1(b), and their co-axial embodiments in Figs. 1(c) and 1(d). It will be seen that they both consist of three transmission line elements, denoted by Z_1, Z_2, Z_3, or Z_{1R}, Z_{2R}, Z_{3R}. Z_1 and Z_{2R} are open circuited while Z_2 and Z_{1R} are short circuited. Z_3 and Z_{3R} are the connecting arms. The direct "T"-stub has its anti-resonant frequency below the resonant frequency and the opposite applies to the reciprocal "T"-stub, the spacing between these frequencies being chiefly dependent on Z_2 or Z_{2R} while the mean impedance level is principally controlled by Z_3 or Z_{3R}. The major part of "T"-stub design consists in evaluating the transmission line lengths required for the elements for given filtering characteristics.

Calculation Methods

" T "-stubs incorporated in previous filter designs[2, 3] etc. have been calculated on the basis of the theory given in the appendix to reference 2. All these filters employed one " T "-stub and one reciprocal " T "-stub, the transmission line lengths of corresponding elements (Z_1, Z_{1R}, etc.) being the same and the input impedances reciprocal over the frequency band. The calculations were based on these facts and the fact that the normalized input reactances of both stubs had to

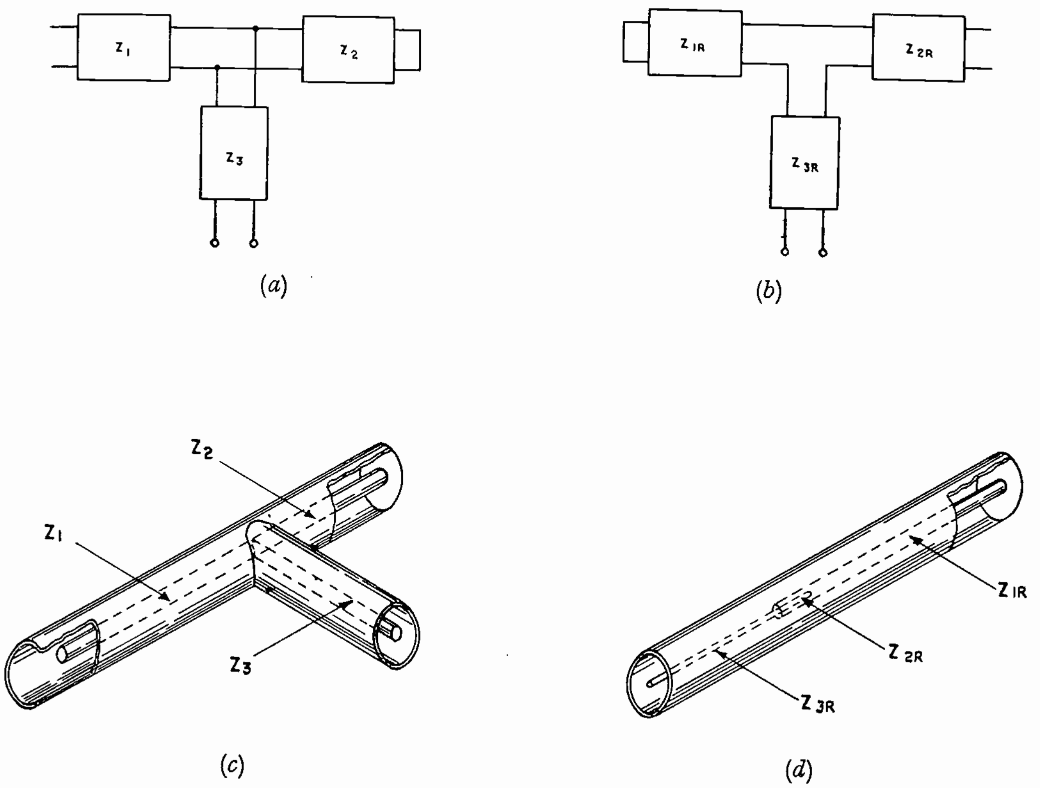


FIG. 1

" T "-connected Stubs.

(*a*) *Direct.*
(*b*) *Reciprocal.*
(*c*) *and* (*d*) *Co-axial Embodiments of* (*a*) *and* (*b*) *respectively.*

be unity at the cross-over (3 db insertion loss) frequency. The need for a new method of calculating the elements became apparent during the investigation into the design of the vestigial sideband filter for the Crystal Palace television station, where it was found necessary to base the design on the constant resistance network[4] shown in Fig. 2(a), which has four reactance elements, two of normalized impedance z_A and two of normalized impedance z_B such that:—

$$z_A \, z_B = \frac{1}{2} \qquad (1)$$

i.e. the normalized impedance of one pair of elements must be equal to one-half the normalized admittance of the other pair over the required frequency band.

The filters based on the constant resistance network of Fig. 2(a) and satisfying the above condition are subsequently referred to as " double notch " type filters, since they have four reactance elements instead of two, and perform a similar function to that of two of the notch vestigial sideband filters of reference 4 connected in cascade.

In order to endow the complete network with correct filtering characteristics, the elements must have resonances and anti-resonances at closely adjacent, fixed,

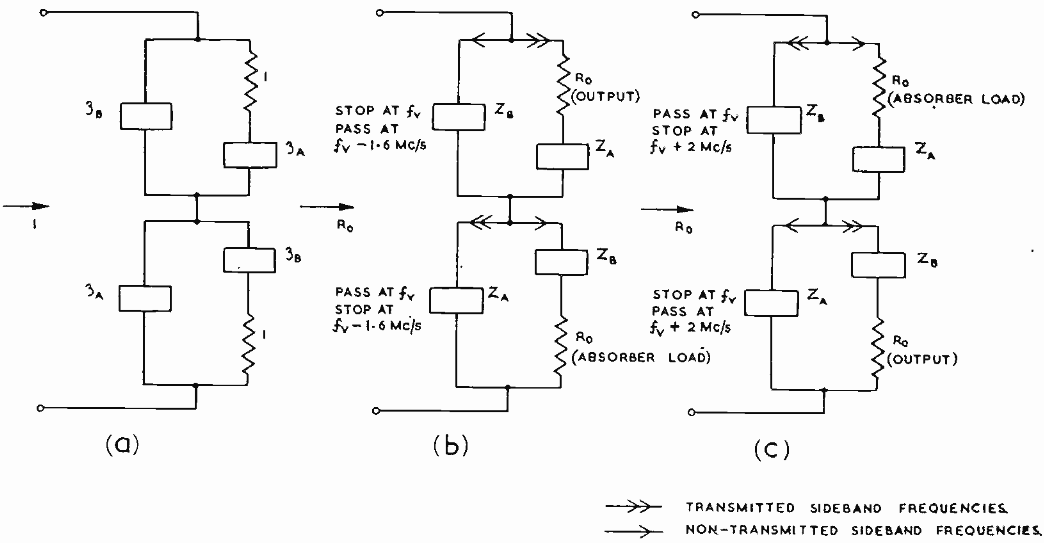


FIG. 2

Double Notch Filters.

(*a*) *Basic Network.*
(*b*) *Upper Sideband Transmitted.* (*C.C.I.R. Specn.*).
(*c*) *Lower Sideband Transmitted* (*B.B.C. Specn.*).

frequencies and they must be designed for a certain mean impedance level. These conditions may be fulfilled by using correctly designed direct and reciprocal " T "-stubs, arranged as shown in Fig. 2(b) for the case of upper sideband transmission and as shown in Fig. 2(c) for the case of lower sideband transmission. In Figs. 2(b) and 2(c) the elements Z_A are direct " T "-stubs and the elements Z_B are reciprocal " T "-stubs. The condition of equation (1) may be realized by either of two basic methods of stub design. These are (a) to design for the required frequency response assuming that the stubs will be true reciprocals, then to construct z_A with characteristic impedance one-half that of z_B (the geometric mean of the characteristic impedances being equal to the working impedance of the filter); (b) to design both pairs of stubs with elements of the same characteristic impedance and to satisfy equation (1) by making these elements of different lengths. In practice this results in a substantial difference only between the lengths of the stub connecting arms Z_3 Z_{3R} and minor differences between the lengths of other corresponding elements.

Design (a) was attractive in that only one major calculation was involved and exact reactance " reciprocity " was theoretically obtainable. However, the realization, on this basis, of a complete co-axial line filter presented considerable mechanical difficulties and ruled out the use of existing standard tubes and tuning adjustments. Design (b) avoided these difficulties but the calculations involved would have been somewhat complicated if carried out on the basis of reference 2. This led to the development of the new method outlined below. Unity input resistance is thereby obtainable at three spot frequencies only—e.g. the " stop ", " pass " and " cross-over " frequencies—but the resulting deterioration in filter characteristics has proved

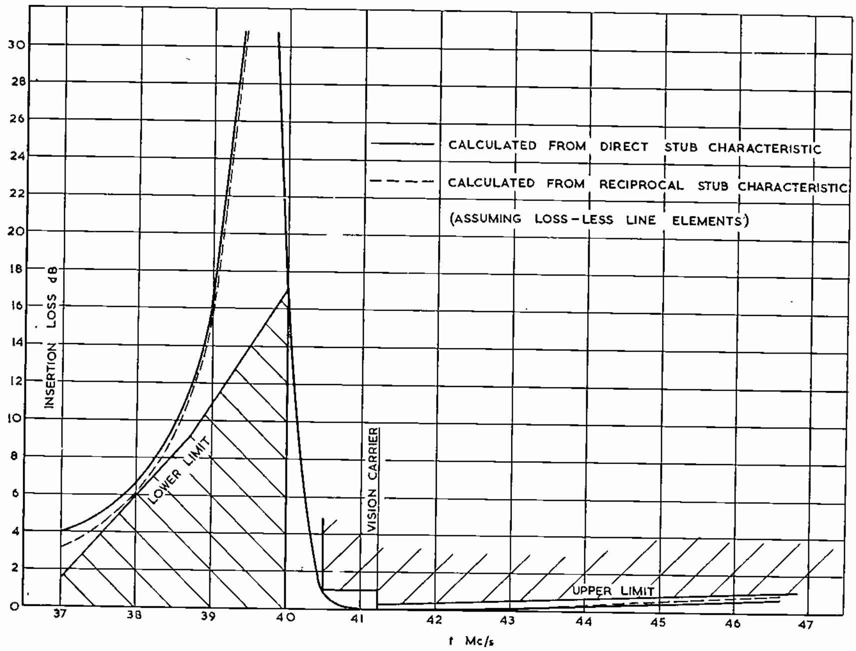


FIG. 3

Insertion Loss Specification and calculated Insertion Loss for C.C.I.R. Band I Vestigial Sideband Filters.

to be negligible except at the edges of the required frequency band. Besides allowing filters of the " double notch " type to be designed more speedily and accurately, the method also provides an alternative way of calculating the elements of direct and reciprocal " T "-stubs as used in combining filters, single notch vestigial sideband filters, and the like.

The great advantage of the new method lies in cases where the insertion loss frequency characteristic of a filter must be " tailored " to fit a tight specification. Such a specification for a sound-above-vision television vestigial sideband filter is shown in Fig. 3, as are the calculated response curves for a filter designed to meet this specification. Details of the design calculation leading to one of these curves, that based on the reciprocal " T "-stub, are given in Example 2. From Fig. 3 it will be observed that the insertion loss must be less than 1 db at 40·5 Mc/s and greater than 17 db at 40 Mc/s. The original calculation method would have involved estimating " stop " and " cross-over " frequencies from the above figures, together with a suitable " pass " frequency either coinciding with or close to the vision carrier frequency and then proceeding by a process of trial and error until a suitable characteristic shape was obtained. In the given example the new method takes the specified insertion loss figures at 40·5 and 40 Mc/s (f_a and f_b) as its starting point, together with an estimated filter pass frequency. Since each type of " T "-stub has three elements its characteristics may be fixed at three frequencies only. Further, the pass frequency must be close to the vision carrier frequency in the case of a television vestigial sideband filter, so that, using the new method, one design calculation, followed by a response calculation, will generally show whether or not the specification is met completely. If not, the only possible step is to move the filter pass frequency within narrow limits. The limitations for a particular application of any notch type filter employing " T "-stubs are quickly shown up—one or two calculations are all that are normally required.

The flexibility of the method is considerable—it has been used, for example, in the design of a reciprocal " T "-stub incorporating a short-circuited Z_{2R} of greater than 90° transmission line length at the frequencies concerned, in place of the usual short-length open circuited Z_{2R}. Further, in the Crystal Palace vestigial sideband filter design it was necessary to incorporate an isolating stub (approximately 90° and short-circuited) in the filter. This appears effectively in parallel with one of the direct " T "-stubs, and it was found possible (see page 126 below and Example 3) to modify the calculation method so that the combined input admittance of " T "-stub and isolating stub would be similar to that of the other " T "-stub in the network. The effect is to change slightly the line lengths of the elements of the " compensated " " T "-stub from those of the one which has no isolating stub in parallel. Another useful modification of the method may be used when it is mechanically advisable to have the characteristic impedance of the connecting arm Z_3 or Z_{3R} different from those of the other elements of the " T "-stub.

Theory and Description of New Calculation Method

The theory given below is derived initially for a straightforward " T "-stub design, as might be used in a single notch vestigial sideband filter. This is the simplest case—an example of the calculation method and layout for such a design is given in Example 1 below. It may be shown (equation 20 of Ref. 2) that the normalized input susceptance of a lossless direct " T "-stub of uniform characteristic impedance is given by the equation:—

$$s = \cot \theta_3 \frac{(\tan \theta_3 + \tan \theta_1 - \cot \theta_2)}{\cot \theta_3 - \tan \theta_1 + \cot \theta_2} \tag{2}$$

Cross multiplying and factorizing:—

$$(s + \cot \theta_3)(\tan \theta_1 - \cot \theta_2) = s \cot \theta_3 - 1$$

$$\tan \theta_1 - \cot \theta_2 = \frac{s \cot \theta_3 - 1}{s + \cot \theta_3} \tag{3}$$

At the " stop " frequency f_s, $s = 0$

At the " cross-over " frequency f_x, $s = 1$.

Hence, from (3)

$$t_s - c_s = -\tan \theta_{3s} \tag{4}$$

$$t_x - c_x = \frac{\cot \theta_{3x} - 1}{\cot \theta_{3x} + 1} \tag{5}$$

From reciprocity considerations it is apparent that similar formulae apply for the reciprocal " T "-stub; replacing s by x_R, θ_1 etc. by θ_{1R} etc.

The notation used above is identical with that of reference 2 with the abbreviations $t_s = \tan \theta_{1s}$, $c_s = \cot \theta_{2s}$, $t_x = \tan \theta_{1x}$, $c_x = \cot \theta_{2x}$.

From (4) and (5) the values of $t_s - c_s$ and $t_x - c_x$ may easily be computed for given θ_3 and by subtraction $(t_x - t_s) + (c_s - c_x)$ is obtained. In general $t_x - t_s$ is considerably greater than $c_s - c_x$, thus little error is introduced at this stage by taking $c_s - c_x$ as having either zero value or some small value derived from previous experience. Hence $t_x - t_s$ is evaluated, and providing $\theta_{1x} - \theta_{1s}$ is known the separate values of these angles may be obtained by inspection from suitable tangent tables. In practice $\theta_{1x} - \theta_{1s}$ is not known exactly, but $f_x - f_s$ is, and some such assumption as $\theta_{1x} = 80°$ at f_x may be used to obtain an initial value of $\theta_{1x} - \theta_{1s}$.

Having obtained the values of θ_{1x} and θ_{1s}, it is a simple matter to obtain c_s, c_x and θ_2 from the values for $t_s - c_s$ or $t_x - c_x$ (equations (4) and (5)). Hence the assumed $c_s - c_x$ and $\theta_{1x} - \theta_{1s}$ may now be corrected and a second calculation using the corrected values will give new values for θ_1 and θ_2. Convergence is very rapid, provided reasonable initial assumptions are made, and the second values obtained for θ_1 and θ_2 are generally accurate enough for all practical purposes.

The above process is repeated for various θ_3—the whole operation being conveniently set down in tabular form as shown on page 128. θ_{1p}, θ_{2p}, θ_{3p} are determined, f_p being known from the design criteria (in the case of a television vestigial sideband filter for example, f_p for a stub in series with the output has a similar value to the vision carrier frequency). At f_p $s = \infty$ so from (2):—

$$\cot \theta_{3p} - \tan \theta_{1p} + \cot \theta_{2p} = 0 \tag{6}$$

Hence the tabulation is continued so as to lead to values for the above function, and the zero value either appears directly or is obtained by interpolation, or extrapolation. Hence the correct θ_1, θ_2, θ_3 are obtained for given f_s, f_p, f_x.

It should be noted that the above process may be modified, using equation (3), so as to make the admittance-frequency characteristic pass through any three given points, subject to certain limitations. It is simplest to use either the stop or the pass frequency of the filter, depending on whether the stub concerned is in parallel or series with the output, so that an equation similar to (6) applies, but the other two design frequencies may be chosen freely so as to fit a given specification (see pages 127 and 139, Example 2).

For the " double notch " type filters, the insertion loss of the network is given by formulae such as:—

$$L = 10 \log_{10} \left(1 + \frac{s^4}{4}\right) \text{db} \tag{7}$$

s being the susceptance of the " T "-stub in parallel with the output branch of the network.

At the cross-over frequency ($L = 3$ db)

$$s = \sqrt{2} \tag{8}$$

Substituting (8) in (3)

$$t_x - c_x = \frac{\sqrt{2} \cot \theta_{3x} - 1}{\cot \theta_{3x} + \sqrt{2}} \tag{9}$$

" Double notch " calculations may be based on equations such as (9) and (4) except where an isolating stub is paralleled across a " T "-stub. In this case, s_1 being the combined admittance of " T " and isolating stubs:—

$$L = 10 \log_{10}\left(1 + \frac{s_1^4}{4}\right)$$

and at the cross-over frequency:—

$$s_1 = s - \cot \theta'_x = \sqrt{2}$$

$$s = \sqrt{2} + \cot \theta'_x \qquad (10)$$

θ' being the electrical length of the isolating stub.

At the stop frequency:

$$s_1 = 0 = s - \cot \theta'_s$$

$$s = \cot \theta'_s \qquad (11)$$

The calculations are then based on equations derived from the substitution of (10) and (11) in (3) (see Example 3). Equation (6) is used unchanged.

Example 1

Design of " T "-stub and reciprocal " T "-stub for given stop, pass and cross-over frequencies.

Design frequencies:—

$$\left.\begin{aligned} f_s &= 39{\cdot}75 \text{ Mc/s} \\ f_x &= 40{\cdot}4 \quad ,, \\ f_p &= 41{\cdot}25 \quad ,, \end{aligned}\right\} f_s \text{ and } f_p \text{ are reversed for reciprocal " T "-stub.}$$

Design equations:—

$$t_x - c_x = \frac{\cot \theta_{3x} - 1}{\cot \theta_{3x} + 1} \qquad (A)$$

$$t_s - c_s = -\tan \theta_{3s} \qquad (B)$$

$$\cot \theta_{3p} - \tan \theta_{1p} + \cot \theta_{2p} = 0 \qquad (C)$$

Initial assumption:—

$$f_0 = 45 \text{ Mc/s}$$

$$\theta_{1x} - \theta_{1s} = 1{\cdot}3°$$

Example 2

Design of Reciprocal " T "-stub for C.C.I.R. Band I V.S.F. (" Double Notch "). Upper Sideband Transmitted.

Vision carrier frequency 41·25 Mc/s.

Specified insertion loss characteristic:—See Fig. 3.

Design frequencies:—

$f_a = 40{\cdot}5$ Mc/s (1 db insertion loss)

$f_b = 40{\cdot}0$ Mc/s (17 db insertion loss)

$f_s = 41{\cdot}35$ Mc/s (0 db insertion loss)

(Stub is paralleled with output branch).

Extract from Design Calculation for Example 1. (Equations A, B, C.)

θ_{3x}	c_{3x}	$c_{3x}-1$	$c_{3x}+1$	(A) $t_x - c_x$	θ_{3s}	(B) $t_s - c_s$	(A)—(B)	Assumed $c_s - c_x$	$t_x - t_s$
29	1·804	·804	2·804	·286	28·54	—·544	·830	·091	·739
30	1·732	·732	2·732	·268	29·52	—·566	·834	·092	·742

θ_{1s}	t_s	c_s	θ_{2s}	θ_{2x}	c_x	Corrected $c_s - c_x$	$t_x - t_s$	θ_{1s}	t_s
79·24	5·262	5·806	9·77	9·93	5·711	·095	·735	79·21	5·247
79·26	5·272	5·838	9·72	9·88	5·742	·096	·738	79·23	5·257

c_s	θ_{2s}	θ_{1p}	θ_{2p}	θ_{3p}	t_{1p}	c_{2p}	c_{3p}	(C) $c_{3p} + c_{2p} - t_{1p}$
5·791	9·80	82·19	10·17	29·62	7·291	5·575	1·759	+·043
5·823	9·75	82·21	10·12	30·63	7·309	5·603	1·688	—·018

$$\left.\begin{aligned}\text{By interpolation } \theta_{1p} &= 82{\cdot}21^\circ\\ \theta_{2p} &= 10{\cdot}14^\circ\\ \theta_{3p} &= 30{\cdot}33^\circ\end{aligned}\right\}\text{ Same for reciprocal stub.}$$

Extract from Design Calculation for Example 2. (Equations B, C, D.)

θ_{3Ra}	$c_{3Ra}-1$	$c_{3Ra}+1$	(B) $t_a - c_a$	θ_{3Rb}	c_{3Rb}	$\cdot267\, c_{3Rb}-1$	$\cdot267 + c_{3Rb}$	(C) $t_b - c_b$	(B)—(C)	Assumed $c_b - c_a$
37	·327	2·327	·140	36·54	1·348	—·640	1·615	—·396	·536	·067
38	·280	2·280	·123	37·54	1·301	—·652	1·568	—·416	·539	·071

$t_a - t_b$	θ_{1Ra}*	t_a	c_a	θ_{2Ra}	θ_{2Rb}	c_b	Corrected $c_b - c_a$	$t_a - t_b$	θ^*_{1Ra}	t_a
·469	79·47	5·380	5·240	10·8	10·67	5·308	·068	·468	79·45	5·370
·468	79·46	5·375	5·252	10·78	10·65	5·320	·068	·471	79·49	5·390

c_a	θ_{2Ra}	θ_{1Rs}	θ_{2Rs}	θ_{3Rs}	t_{Rs}	c_{2Rs}	c_{3Rs}	(D) $c_{3Rs} + c_{2Rs} - t_{1Rs}$
5·230	10·82	81·12	11·05	37·78	6·401	5·121	1·290	+·010
5·267	10·75	81·16	10·98	38·80	6·430	5·154	1·244	—·032

By interpolation $\theta_{1R_s} = 81{\cdot}13^\circ$ $\theta_{2R_s} = 11{\cdot}03^\circ$ $\theta_{3R_s} = 38{\cdot}03^\circ$

*Obtained from previous column by use of tangent tables.

Design equations:—

$$L = 10 \log_{10}\left(1 + \frac{s_R^4}{4}\right) \text{ db} \qquad \text{(A)}$$

where s_R = input susceptance of Reciprocal " T "-stub.

At f_a, $L = 1$ and hence from (A)

$$x_R = s_R \doteqdot 1$$

From (3) modified for Reciprocal " T "-stub

$$t_a - c_a = \frac{\cot \theta_{3Ra} - 1}{\cot \theta_{3Ra} + 1} \qquad \text{(B)}$$

At f_b, $L = 17$ and hence from (A).

$$s_R = 3{\cdot}74$$

$$x_R = 0{\cdot}267$$

From (3)

$$t_b - c_b = \frac{{\cdot}267 \cot \theta_{3Rb} - 1}{{\cdot}267 + \cot \theta_{3Rb}} \qquad \text{(C)}$$

At f_s, $s_R = 0$

From (2)

$$\cot \theta_{3Rs} - \tan \theta_{1Rs} + \cot \theta_{2Rs} = 0 \qquad \text{(D)}$$

For response curves calculated from modified equation (2) and insertion loss formula (A), see Fig. 3.

Example 3

Design of Compensated " T "-stub for Crystal Palace Vestigial Sideband Filter (" Double Notch ").

Vision carrier frequency=45 Mc/s.

∴ Required insertion loss characteristic:—

At 42	Mc/s	$< 0{\cdot}5$ db
45·75	Mc/s	$< 1{\cdot}5$ db
46·5	Mc/s	> 10 db
48	Mc/s	> 10 db

Design frequencies:—

$$\left.\begin{aligned} f_s &= 44{\cdot}8 \text{ Mc/s} \\ f_x &= 46{\cdot}2 \text{ Mc/s} \\ f_p &= 47 \text{ Mc/s} \end{aligned}\right\} \text{Stub is in parallel with output branch.}$$

Design equations:—

$$L = 10 \log_{10}\left(1 + \frac{s_1^{\,4}}{4}\right) \qquad \text{(A)}$$

$$t_x - c_x = \frac{mn - 1}{m + n} \qquad \text{(B)}$$

where $m = \sqrt{2} + \cot \theta'_x$ and $n = \cot \theta_{3x}$

$$t_s - c_s = \frac{pq - 1}{p + q} \qquad \text{(C)}$$

where $p = \cot \theta'_s$, $q = \cot \theta_{3s}$

$$\cot \theta_{3p} - \tan \theta_{1p} + \cot \theta_{2p} = 0 \qquad \text{(D)}$$

The isolating stub length is taken as 85° at f_x = 82·42° at f_s hence:—

$$m = \sqrt{2} + \cdot 087 = 1\cdot 501$$

$$p = \cdot 133$$

EXTRACT FROM DESIGN CALCULATION FOR EXAMPLE 3. (EQUATIONS B, C, D.)

θ_{3x}	n	$mn - 1$	$m + n$	(B) $t_x - c_x$	θ_{3s}	q	$pq - 1$	$p + q$	(C) $t_s - c_s$	(B)—(C)
49	·869	·304	2·370	·128	47·52	·916	—·878	1·049	—·837	·965
50	·839	·259	2·340	·111	48·49	·886	—·882	1·019	—·866	·977

Assumed $c_s - c_x$	$t_x - t_s$	θ_{1s}	t_s	c_s	θ_{2s}	θ_{2x}	c_x	Corrected $c_s - c_x$	$t_x - t_s$	θ_{1x}
·152	·813	75·73	3·932	4·769	11·84	12·21	4·621	·148	·816	75·75
·153	·824	75·81	3·955	4·821	11·72	12·09	4·669	·152	·826	75·83

t_s	c_s	θ_{2s}	θ_{1p}	θ_{2p}	θ_{3p}	t_{1p}	c_{2p}	c_{3p}	(D) $c_{3p} + c_{2p} - t_{1p}$
3·938	4·775	11·82	79·47	12·40	49·85	5·380	4·543	·844	·007
3·961	4·827	11·70	79·55	12·28	50·87	5·420	4·592	·814	—·014

By interpolation θ_{1p} = 79·51°, θ_{2p} = 12.34°, θ_{3p} = 50·36°.

List of Symbols

Z_1, Z_2, Z_3	Direct " T "-stub elements.
Z_{1R}, Z_{2R}, Z_{3R}	Reciprocal " T "-stub elements.
θ_1, θ_2, θ_3 θ_{1R}, θ_{2R}, θ_{3R}	Angular transmission line lengths of above at any frequency.
p	Subscript used with θ_1, θ_{1R}, f etc. to denote " pass " condition.
s	Subscript used with θ_1, θ_{1R}, f etc. to denote " stop " condition.
x	Subscript used with θ_1, θ_{1R}, f etc. to denote " cross-over " condition.
a b	Subscript used with θ_1, θ_{1R}, f etc. to denote conditions at particular design frequencies other than above.
x, x_R, s, s_R	Normalized input reactance or susceptance of " T "-stub or Reciprocal " T "-stub.
t_s, t_x, t_a, t_b	Abbreviations for tan θ_{1s} or tan θ_{1Rs} etc. used in design calculations.
t and c	are also used in abbreviations such as t_{1Rs}, c_{3Rs} for tangent and cotangent of particular angles.
L	Insertion loss of complete filter in db.
θ'	Angular transmission line length of isolating stub.

References

(1) Marconi's Wireless Telegraph Co. Ltd. British Patent No. 698509.

(2) Sosin, B. M. "A Combining Filter for Vision and Sound Transmission." J.I.E.E. 1952 Part IIIA.

(3) Sosin, B. M. " V.H.F. Power Transmission Equipment for Band III Television Broadcast," Marconi Review 114, p.88.

(4) Norton, E. L. "Constant Resistance Networks with Applications to Filter Groups." B.S.T.J. Vol. 16, 1937, pp.178-193.

SOME DESIGN CONSIDERATIONS FOR LINKS CARRYING MULTICHANNEL TELEPHONY

PART I

By S. Fedida, B.Sc.(Eng.),(Hons.), A.C.G.I., A.M.I.E.E.

A survey of some of the important factors affecting the design of links conveying multichannel telephony signals is given in the following article. This survey covers the amplitude and phase linearity of terminal modulators, demodulators and associated I.F. amplifiers, the phase linearity of the repeaters, the effect of weak echoes due to feeder mismatches or multipath propagation and some of the testing methods which have been found useful in the design of suitable equipment.

The calculation of inter-modulation distortion is treated in the second part of this article.

Overall Performance Requirements

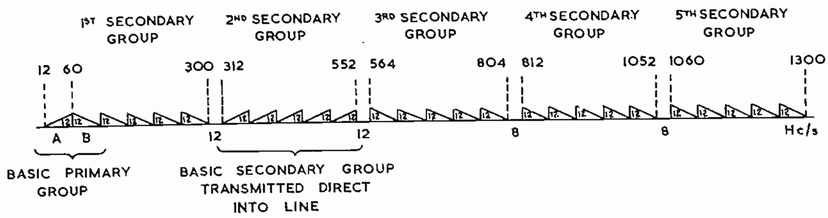
THE performance requirements of multichannel radio links are based on the C.C.I.F. standards, recommended for inter-continental trunk telephone circuits, carrying several supergroups.

If these performance requirements are satisfied, radio links may form a part or the entirety of a trunk telephone network of a quality comparable to that of metallic circuits.

Because of this requirement of inter-changeability with coaxial cables and in order to make use of the channelling equipment, which is already available for use over metallic circuits, the baseband signal, or intelligence, applied to the radio link, conforms to the standards adopted by the C.C.I.F.[1] The composition of the baseband signal for systems carrying several supergroups is as shown in Fig. 1 which has been obtained from Fig. 26 of reference 1.

Here a speech channel is contained in the band 300-3,400 c/s and a primary group, or simply a group, is made up of twelve single-sideband channels spaced 4 kc/s apart and placed side by side so as to occupy a band 48 kc/s wide. The basic group extends from 12-60 kc/s. A secondary group, or supergroup, is made up of five groups of twelve channels each covering a band of 240 kc/s. The basic supergroup which is normally transmitted over a metallic circuit, carrying sixty channels only, is placed in the band extending from 312 to 552 kc/s. Other supergroups may be added according to the plan of Fig. 1.

The performance expected of a long route is defined by reference to a fictitious circuit the "*circuit fictif de reference*," which may be part of an international cable connection. This circuit is 2,500 km long, carries 120 channels or more and is normally divided into nine sections each of 280 km, at the terminals of each of which a frequency translation of one type or another is contemplated. The plan of the reference circuit, for coaxial cables, is shown in Fig. 2. It may be seen that the actual voice channels are accessible at only two intermediate points (apart from the terminals). Frequency translations from basic groups to basic supergroups and from basic supergroups to baseband, are made at various other points for the purpose of channel dropping.

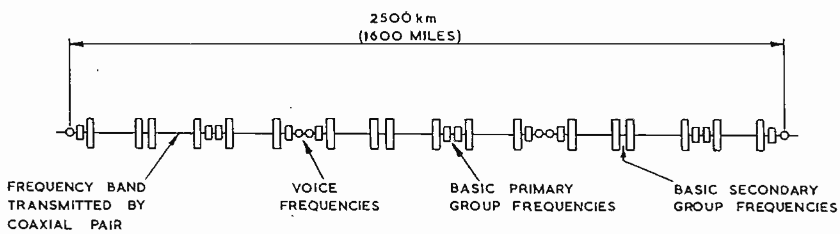


Represents a 12 channel primary group, wherein the effective carriers are spaced 4 kc/s apart and in which the voice frequencies are in the right order in the various telephone channels.

Represents a 12 channel primary group, wherein the effective carriers are spaced 4 kc/s apart and in which the voice frequencies are in the reverse order in the various telephone channels.

FIG. 1. *Arrangement of telephone channels transmitted in wide band international cables.*

The standard recommended by the C.C.I.F. for this reference circuit is in terms of the noise power measured at a " zero reference point " in a channel, at the receiving end of the system, which must not exceed 10,000 pW, for more than 1% of the busiest hour. The nine sets of frequency translation equipment, including terminals, are allocated a total of 2,500 pW, while the rest may be allowed to the cable route. Thus an average of 3 pW per km may be contributed by the cable system on the assumption that all noise contributions are power adding.



Voice modulation equipment (translation of voice frequency to basic primary groups and vice versa).

Primary group modulation equipment (translation of basic primary groups to basic secondary groups and vice versa).

Secondary group modulation equipment (translation of basic secondary group to the transmitted band and vice versa).

FIG. 2. *General arrangement of the C.C.I.F. " circuit fictif de reference " for coaxial cables.*

Since the baseband signal is available at points 280 km apart along the cable route, any section between two frequency translation terminals could be replaced by a radio link. Thus a reference radio link could be defined as having a length of 280 km (175 miles) and such that its noise contribution due to all causes, i.e. valve noise plus intermodulation noise, should not exceed 840 pW, as measured with a psophometer, in a channel at a zero reference level point.

If the system noise is measured in a bandwidth extending from 0 to 4 kc/s, instead of a channel bandwidth of 3·1 kc/s and with an unweighted quadratic meter, instead of a psophometer, the above noise contribution may be corrected by +3 db to 1,680 pW.

Subdivision of Noise Contributions, among the elements of the Reference Radio Link

The subdivision of the noise contributions made by the various parts of the reference radio link is a matter for the equipment designer. This subdivision is generally made in such a way as to lead to an economical system design, i.e., so as to place equal demands on the several parts of the system.

Noise in a radio link has two origins. A part of it is made up of the valve noise occurring in the early stages of the microwave receivers at the terminals and at the repeaters and this is, in the absence of fading, a direct function of the available transmitter power, distance between hops, aerial gains, etc., and of the frequency deviation. The remainder is made up of intermodulation between the various components of the multichannel signal.

The multichannel signal is composed of the addition of several hundred speech channels, each of which is made up of very large numbers of sinusoids having random amplitude and phase relationships. The resultant signal closely resembles random noise and the cross-modulation between the various terms, in this signal, is also very similar to random noise. Cross-modulation noise increases, in general, with the frequency deviation applied to the system, while the basic valve noise decreases, as the deviation is increased. It follows that a compromise has to be struck between valve noise and cross-modulation noise.

A reasonable arrangement is to make them both equal. If, on the other hand, an excess of power is available it is advisable to reduce the deviation and so increase the valve noise contribution in relation to the intermodulation noise.

The reference radio link is made up of a number of repeaters with terminals at each end. A reasonable hop size, between two consecutive repeaters, is about 35 miles, but more hops may have to be allowed to take account of peculiarities in the terrain configuration between any two given stations, and of fading.

If we allow an average hop length of 35 miles, i.e., five hops in the reference link, we may subdivide the noise contributions as follows:—

Distribution A

	Crosstalk	*Noise*
Terminal Modulator and Demodulator	210 pW	210 pW
Five Receivers and five Transmitters (126 pW per pair)	630 pW	630 pW
Total	840 pW	840 pW

The crosstalk contribution of the terminal modulators and demodulators is likely to be greater than that of the R.F. portions of the system, as these elements are subject to amplitude, as well as phase non-linearity; this is allowed for in the above table.

An alternative noise subdivision suggested elsewhere could be made in the proportion of 1/3 for valve noise and 2/3 for crosstalk, on the basis that it is unlikely that fading periods, when valve noise is at its peak, should coincide with the busiest hour, when traffic and consequently intermodulation noise, is at its peak.

It is also assumed that the reference link is made up of seven sections of 25 miles each and that the crosstalk contribution of a pair of terminals is equal to one half the crosstalk contribution of the whole link.

Thus we have:—

	Crosstalk
Terminal Transmitter and Receiver (including modem)	560 pW
Six repeaters	560 pW

We can deduce the following table on the basis of the above figures:

DISTRIBUTION B

	Crosstalk	*Noise*
Terminal Modulator and Demodulator	467 pW	234 pW
Seven Receivers and seven Transmitters (93 pW of crosstalk or 46 pW of valve noise per pair)	653 pW	326 pW
	1,120 pW	560 pW

In distribution B the modulators and demodulators are allowed to generate 3·5 db more crosstalk and 0·5 db more noise than in distribution A. On the other hand each transmitter/receiver pair must generate about 1·5 db less crosstalk and 4·3 db less valve noise for a shorter hop length. The reduction of hop length from 35 to 25 miles gives an increase in signal to noise ratio of nearly 3 db, so that distribution B implies a better noise and crosstalk performance, from the microwave equipment, of 1·5 db.

The signal to noise ratios in a channel, in the absence of traffic and modem noise due to all causes, for a single hop link for distributions A and B respectively should be 69 and 73·5 db.

The Total Multichannel Signal

Some uncertainty exists as to the power of the total multichannel signal which is normally available for transmission along the radio link. A knowledge of this power is important, since it determines the linearity (amplitude and phase) of every component part of a link.

The overall C.C.I.F. performance requirement is stated in terms of a limiting noise level which is not to be exceeded for more than 1% of the busiest hour and since this noise depends on the traffic loading, it is necessary to determine the power P of the multichannel signal applied to the terminals of the radio link which is not exceeded for more than 1% of the busiest hour. The link should not cause the noise in a channel to exceed the limit stated, when loaded continuously with a signal of power P. The ratio of P to the channel test tone level p is sometimes referred to as the loading factor L.

The actual speech power level p_1 in a telephone channel is a variable quantity. Average values of $-9{\cdot}9$[2] dbm, $-15{\cdot}4$[3] and -12 dbm have been quoted, the measurement being made at a point of zero reference level.

In systems carrying large number of channels, a small proportion of these carry speech at any one time. Measurements taken by the Bell Telephone Laboratories[2] provide the basis for the calculation of the utilization factor k of a multichannel system. This is defined as the ratio of the number of active channels n to the total number of available channels N which is not exceeded for more than 1% of the time.

Values of k for typical systems, based on the busiest hour utilization, are as follows:—

TABLE I

UTILIZATION FACTOR OF MULTICHANNEL SYSTEMS

Number of Channels N	Utilization factor k	No. of active Channels n
60	0·38	23
120	0·34	41
240	0·32	77
600	0·30	180
960	0·30	288

The mean power of the multichannel speech signal P_1 not exceeded for more than 1% of the busiest hour is obviously np_1, if the speech levels in all the channels are controlled to the mean value.

In general, the mean speech power in a channel is not controlled, so that the mean power of the total multichannel signal is rather higher than np_1. Values of P_1, obtained from reference 2 are given below.

TABLE II

Number of Channels N	Mean power of multichannel speech signal P_1 not exceeded for more than 1% of busiest hour P_1
60	5·6 dbmo*
120	7·3 ,,
240	8·8 ,,
600	12·0 ,,
960	13·6 ,,

* dbmo. db reference 1 mW, measured at a zero reference level point.

The values of P_1 listed above have been based on a mean channel level of -12 dbmo.

An important point to note in connection with the multi-channel signal is that power peaks, several times greater than the mean power P_1, may occur occasionally. Although the rate of occurrence of these peaks is very small, nevertheless the radio system should be able to withstand these peaks without "overloading."

If we assume the multichannel signal power P_1 to be sustained continuously, peaks PP_1 not exceeded for more than 1% of the time, and peaks PP_0 almost never exceeded may be obtained from the table below. These figures are calculated on the assumption that the multichannel signal is equivalent to a uniform band of random noise.

TABLE III

Number of Channels	Ratio of peak power PP to mean power P_1	
N	PP_1/P_1	PP_0/P_1
60	9·3 db	15·1 db
120	8·9 ,,	14·3 ,,
240	8·6 ,,	13·7 ,,
600	8·4 ,,	13·3 ,,
960	8·3 ,,	13·2 ,,

The overload point of a telephone system has been variously defined. According to one definition it is the level at which an increase of 1 db in the input power produces an increase of 0·75 db in the output power. According to another definition[4] it is the level at which an increase of 1 db in the input power, produces an increase of 20 db in the power of the third harmonic. It can be proved that the first definition specifies a level at which a third harmonic margin of about 30 db is obtained, assuming the system to be represented by a power series with terms up to the third.

The total multichannel signal power P is made up of the total multichannel speech power P_1, to which should be added frequency and level pilots, carrier leaks, signalling tones, voice frequency telegraph tones, etc. The allowance for these additional signals varies as between systems. A figure sometimes used for high capacity systems is 4 db.

Thus the total multichannel signal power, for a mean channel level of -12 dbmo, is given by the following table:—

TABLE IV

Number of Channels N	Mean multichannel signal power P not exceeded for more than 1% of the busiest hour P
60	9·6 dbmo
120	11·3 ,,
240	12·8 ,,
600	16·0 ,,
960	17·6 ,,

The loading factor L is according to the definition given above, equal to the value of P in the above table. In the case of a frequency modulated system the loading factor specifies the ratio of the multichannel signal frequency deviation to the channel test tone frequency deviation.

The Multichannel Signal Frequency Deviation

The rms channel test tone deviation, for system without preemphasis, has been provisionally agreed at 200 kc/s.

It follows that the multichannel signal frequency deviations should be as tabulated below.

TABLE V

Number of Channels N	Top mod. frequency f_2	Multichannel signal frequency deviation (rms) ΔF
60	300 kc/s	604 kc/s
120	552 ,,	735 ,,
240	1,052 ,,	875 ,,
600	2,540 ,,	1,260 ,,
960	4,028 ,,	1,520 ,,

Amplitude Linearity Requirements

Crosstalk in a frequency modulated radio link arises partly in the terminal modulators, demodulators and line amplifiers, on account of the non-linearity in the transfer characteristics, and partly in the I.F. and R.F. portions of the systems, on account of the non-linearity in the phase characteristic of these portions.

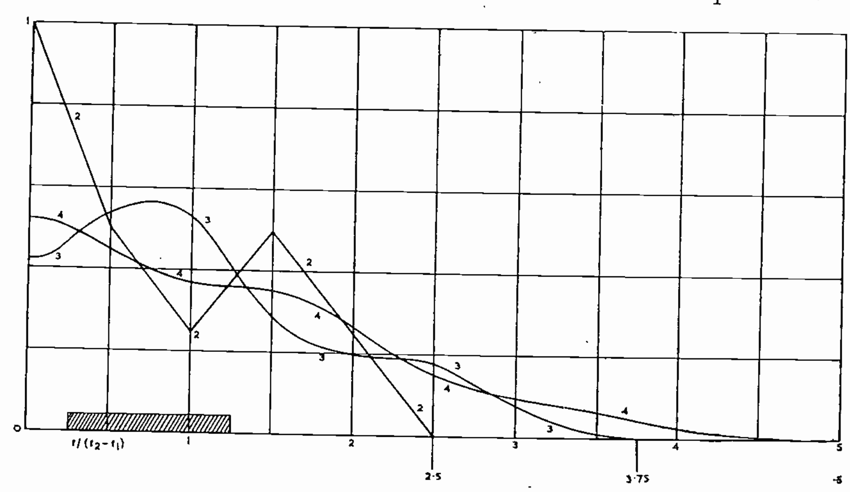


FIG. 3

Amplitude intermodulation spectra for $f_2/f_1 = 5$ (60 channel system).

Let us assume, for the purpose of calculating the effect of amplitude non-linearity, an overall transfer characteristic of the type:

$$V_0 = a_1 V_i + a_2 V_i^2 + a_3 V_i^3 + \ldots\ldots \qquad (1)$$

where the coefficients a_1, a_2, a_3, are single valued and independent of frequency and let

N = total number of channels.
f_1, f_2 = lowest and highest frequencies in the working band.
t_n = n^{th} harmonic power, in mW, produced by that sinusoidal input, which gives a fundamental output power of 1 mW.
T_n = total n^{th} order distortion power in mW.
P = total fundamental power output in mW.
$F_{n(\omega)}$ = frequency distribution of intermodulation powers of order n.

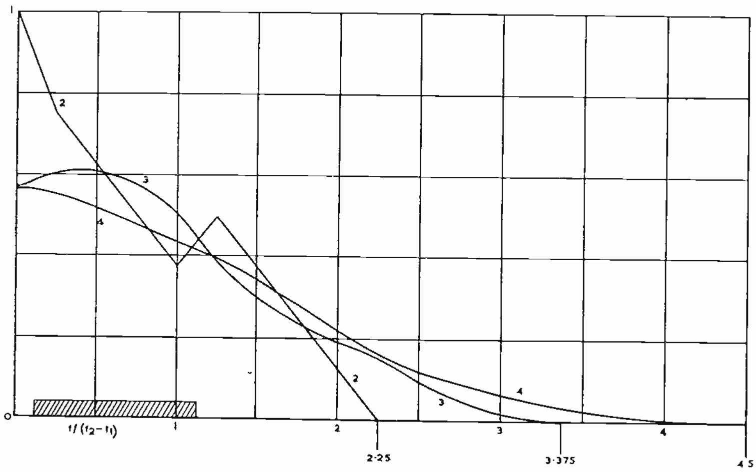


FIG. 4

Amplitude intermodulation spectra for $f_2/f_1 = 9$ (120 *channel system*).

It is shown in equation 48 of Part II that the total distortion power of order n is:

$$T_n = 2^{n-1}\, n!\, t_n\, P^n \tag{2}$$

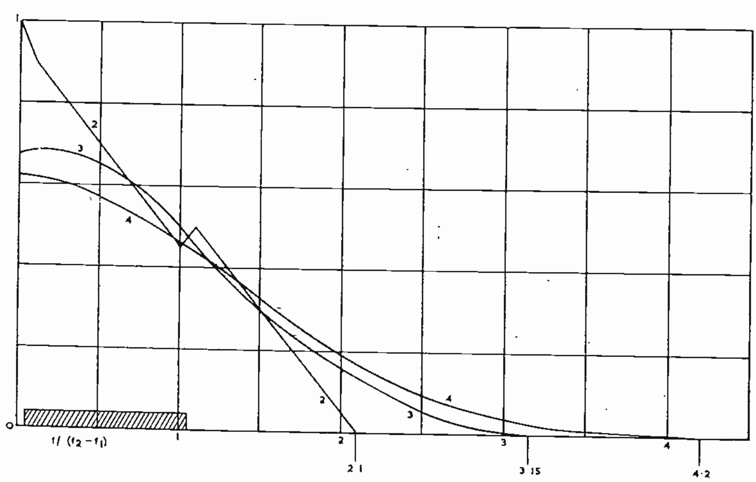


FIG. 5

Amplitude intermodulation spectra for $f_2/f_1 = 21$ (300 *channel system*).

The proportion of this power falling within the working band depends on the ratio $s = f_2/f_1$, always assuming that the input signal spectrum is uniform between f_1 and f_2.

The distortion spectra for various values of s are given in Figs. 3 to 6.

The distortion power of a given order, falling within a channel of bandwidth 4 kc/s, is given by equation 84 of Part II.

$$d_n = \frac{T_n F_{n(\omega)}}{N} \tag{3}$$

Second and Third Order Intermodulation Powers

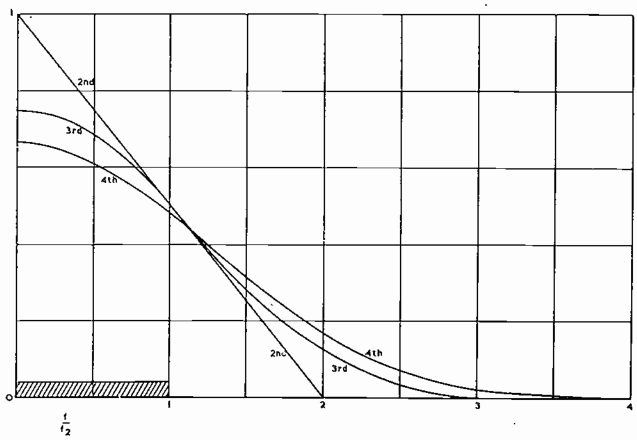


Fig. 6

Amplitude intermodulation spectra for $f_2/f_1 = \infty$.

If we assume that the input signal spectrum extends from zero frequency to frequency f_2, then we have by combining equations (2) and (3)

$$d_2 = \frac{4\, t_2 P^2}{N} F_{2(\omega)} \tag{4}$$

The maximum of the second order spectrum is at zero frequency, when $F_{2(\omega)} = 1$. Therefore the second order intermodulation noise in the noisiest channel is:

$$d_2 = \frac{4\, t_2 P^2}{N} \tag{5}$$

The second harmonic margin $\left(\frac{F}{H_2}\right)_p$ of a 1 mW test tone or the equivalent channel test tone deviation, in a system producing a second order distortion power of d_2, in the noisiest channel, when delivering a total fundamental power output P (i.e., at full deviation), is:

$$\left(\frac{F}{H_2}\right)_p = \frac{1}{t_2} = \frac{4\, P^2}{d_2 N} \tag{6}$$

If the system is tested with a tone, the power output of which is qmW, the second harmonic margin of this test tone is:

$$\left(\frac{F}{H_2}\right)_q = \frac{4\, P^2}{d_2\, Nq} \tag{7}$$

and when $q = P$ we obtain

$$\left(\frac{F}{H_2}\right)_P = \frac{4\, P}{d_2 N} \tag{8}$$

When the system carries traffic, P represents the total multichannel signal power in mW, measured at a point of zero level, and this power is proportional to the loading factor L.

In the case of third order distortion powers, the total amount produced is $T_3 = 24\, t_3\, P^3$, with a peak in the spectrum distribution at zero frequency ($f_1 = 0$) of value 0·75. Thus the 3rd order distortion power, falling into the lowest frequency channel (4 kc/s wide), is:

$$d_3 = \frac{18\, t_3\, P^3}{N} \tag{9}$$

The third harmonic margin of a 1 mW test tone in a system producing a third order distortion power d_3, in the noisiest channel, when delivering a fundamental power output P is:

$$\left(\frac{F}{H_3}\right)_p = \frac{1}{t_3} = \frac{18\,P^3}{d_3\,N} \tag{10}$$

If the system is tested with a tone, the power output of which is qmW, the third harmonic margin of this test tone is:

$$\left(\frac{F}{H_3}\right)_q = \left(\frac{F}{H_3}\right)_q \times \left(\frac{p}{q}\right)^2$$

$$= \frac{18\,P^3}{d_3\,N} \times \frac{1}{q^2} \tag{11}$$

when $q = P$ we have

$$\left(\frac{F}{H_3}\right)_P = \frac{18\,P}{d_3\,N} \tag{12}$$

Again P represents the total multichannel signal power in mW, measured at a point of zero level.

The above formulae are summarized below:

Test tone	Deviation	2nd order distortion margin	3rd order distortion margin
1 mW	Channel deviation	$\frac{4\,P^2}{d_2\,N}$	$\frac{18\,P^3}{d_3\,N}$
P mW	Multichannel deviation	$\frac{4\,P}{d_2\,N}$	$\frac{18\,P}{d_3\,N}$

(13)

Modulator and Discriminator Amplitude Linearity

We have allowed a total intermodulation noise power of 210 pW for a terminal modulator/demodulator, including associated line amplifiers.

We shall here assume that half the intermodulation power, i.e., 105 pW, is caused by amplitude non-linearity and the other half by phase non-linearity. Again we may allocate 52·5 pW each to the modulator and discriminator with their associated amplifiers, to allow for amplitude non-linearity. Finally we may assume that the second and third order contributions are equal. Thus we get finally:

$$d_2 = d_3 = 26 \text{ pW } (-76 \text{ dbm})$$

The distortion margins to be expected from modulators and demodulators, for various systems and calculated according to equation (13), are given below:

TABLE VI

SECOND AND THIRD ORDER HARMONIC MARGINS OF MODULATORS AND DEMODULATORS FOR SIGNAL/L.F. INTERMODULATION RATIO OF 76 DB IN EACH

Number of Channels			Channel deviation r.m.s.	Multichannel deviation r.m.s.	Distortion margin on the Multichannel deviation	
$10 \log_{10} N$	N	L db	δf	ΔF	F/H_2	F/H_3
17·8	60	9·6	200 kc/s	604 kc/s	73·8 db	80·3 db
20·8	120	11·3	,,	735 ,,	72·5	79·0
23·8	240	12·8	,,	875 ,,	71·0	77·5
27·8	600	16·0	,,	1260 ,,	70·2	76·7
29·8	960	17·6	,,	1520 ,,	69·8	76·3

If we adopt the noise distribution B, the above figures may all be reduced by 3·5 db.

Phase Linearity Requirements

The non-linear phase characteristics of networks, whether passive or active, transmitting frequency modulated signals is responsible for producing intermodulation distortion similar to that caused by amplitude non-linearity.

For the purpose of calculating this distortion we may use an approximate formula developed by Van der Pol.[6].

$$g_0(t) = g(t) + \frac{d}{dt} g(t) \left[\phi'(\omega) + \Delta\omega g(t)\, \phi''(\omega) + \frac{\Delta\omega^2 g(t)^2}{2} \phi'''(\omega) + \ldots \right] \quad (14)$$

where the input signal $\Delta\omega g(t) = \Delta\omega \cos \omega t$, $g_0(t)$ is the output signal and $\phi^n(\omega)$ the n^{th} differential coefficient of the phase-frequency characteristic.

Let τ_n = n^{th} order delay coefficient of network.

N = total number of channels.

$n^2 t'_n$ = n^{th} harmonic power in mW produced by that sinusoidal input, which gives a fundamental output power of 1 mW, at the top frequency in the signal band (normalized angular frequency of unity).

T'_n = total n^{th} order distortion power in mW produced by phase distortion.

P = total fundamental power output in mW due to all channels.

$F'_n(\omega)$ = frequency spectrum of n^{th} order intermodulation powers caused by phase distortion.

It is shown (Part II, equation 122) that the total intermodulation power of order n generated by phase non-linearity is:

$$T'_n = \frac{2^{n-1}(n-1)!\, n^2 t'_n P^n}{3} \quad (15)$$

on the assumption that the input signal spectrum is uniform from zero frequency to a maximum modulation frequency f_2 (in normalized units 0 to 1).

The proportion of this power, falling within a 4 kc/s channel, in a system carrying N channels is (equation 124, Part II):

$$d'_n = \frac{T'_n F'_{n(\omega)}}{N} \qquad (16)$$

Intermodulation spectra $F'_{n(\omega)}$ for a flat input spectrum are given in Fig. 7.

Second and Third Order Intermodulation Powers

By combining equations (15) and (16) we obtain for the 2nd order intermodulation power in a channel.

$$d'_2 = \frac{8\,t'_2 P^2 F'_{2(\omega)}}{3N} \qquad (17)$$

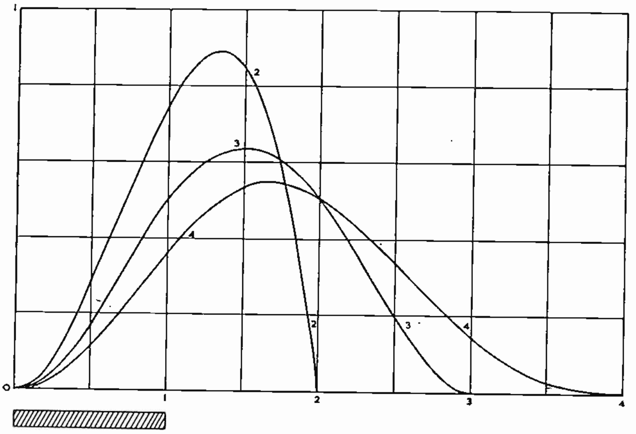


FIG. 7

Intermodulation spectra due to phase distortion $f_2/f_1 = \infty$.

The maximum of the 2nd order intermodulation power, within the signal band, is at the top modulation frequency, where from Fig. 7 $F'_{2(\omega)} = 0{\cdot}75$, thus we may write

$$d'_2 = \frac{1}{2}\,4\,t'_2 P^2 \qquad (18)$$

The second harmonic margin of a 1 mW test tone signal (or equivalent channel test tone deviation) of frequency equal to the top modulation frequency in a system producing a second order intermodulation power d'_2 in the noisiest channel, when delivering a total fundamental power output P (i.e., at full deviation), is:

$$\left(\frac{F}{H_2}\right)'_p = \frac{1}{4\,t'_2} = \frac{P^2}{2\,d'_2 N}\cdot \quad \left(\frac{f}{f_2} = 1\right) \qquad (19)$$

It is shown in Part II, equation 16, that the harmonic power is proportional to the square of the frequency of the fundamental. Thus, for a test tone of 1/3 the top modulation frequency, the harmonic margin is:

$$\left(\frac{F}{H_2}\right)'_p = \frac{9}{2}\,\frac{P^2}{d'_2 N}\cdot \quad \left(\frac{f}{f_2} = \frac{1}{3}\right) \qquad (20)$$

The harmonic margins of test tones producing fundamental output powers of q mW and P mW, at frequencies f_2 and $f_2/3$, respectively, are:

$$\left(\frac{F}{H_2}\right)'_q = \frac{1}{2}\,\frac{P^2}{d'_2 N q} \quad \left(\frac{f}{f_2} = 1\right) \qquad (21)$$

$$\left(\frac{F}{H_2}\right)'_q = \frac{9}{2}\,\frac{P^2}{d'_2 N q} \quad \left(\frac{f}{f_2} = \frac{1}{3}\right) \qquad (22)$$

$$\left(\frac{F}{H_2}\right)'_P = \frac{1}{2}\,\frac{P}{d'_2 N} \qquad \left(\frac{f}{f_2} = 1\right) \tag{23}$$

$$\left(\frac{F}{H_2}\right)'_P = \frac{9}{2}\,\frac{P}{d'_2 N} \qquad \left(\frac{f}{f_2} = \frac{1}{3}\right) \tag{24}$$

In the case of third order distortion powers, the total amount produced is

$$T'_3 = \frac{72}{3}\,t'_3 P^3 \tag{25}$$

The amount falling in a 4 kc/s channel, at the top of the signal band ($F'_{3(\omega)} = 0{\cdot}35$) is

$$d'_3 \doteqdot \frac{72}{10}\,t'_3 P^3 \tag{26}$$

The third harmonic margin of a 1 mW test tone of frequency equal to the top modulation frequency in the signal band in a system producing a third order distortion power d'_3, in the noisiest channel, due to phase non-linearity, when delivering a total fundamental power output P, is:

$$\left(\frac{F}{H_3}\right)'_P = \frac{1}{9\,t'_3} = 0{\cdot}8\,\frac{P^3}{d'_3 N} \quad \left(\frac{f}{f_2} = 1\right) \tag{27}$$

For test tones delivering power outputs of q mW and P mW, at frequencies f_2 and $f_2/3$, the third harmonic margins are:

$$\left(\frac{F}{H_3}\right)'_q = 0{\cdot}8\,\frac{P^3}{d'_3 N q^2} \quad \left(\frac{f}{f_2} = 1\right) \tag{28}$$

$$\left(\frac{F}{H_3}\right)'_q = 7{\cdot}2\,\frac{P^3}{d'_3 N q^2} \quad \left(\frac{f}{f_2} = \frac{1}{3}\right) \tag{29}$$

$$\left(\frac{F}{H_3}\right)'_P = 0{\cdot}8\,\frac{P}{d'_3 N} \quad \left(\frac{f}{f_2} = 1\right) \tag{30}$$

$$\left(\frac{F}{H_3}\right)'_P = 7{\cdot}2\,\frac{P}{d'_3 N} \quad \left(\frac{f}{f_2} = \frac{1}{3}\right) \tag{31}$$

The above formulae are summarized below:—

Test tone	Deviation	Frequency	2nd order margin	3rd order margin
1 mW	Channel	f_2	$\frac{1}{2}\,\frac{P^2}{d'_2 N}$	$0{\cdot}8\,\frac{P^3}{d'_3 N}$
P mW	Multichannel	f_2	$\frac{1}{2}\,\frac{P}{d'_2 N}$	$0{\cdot}8\,\frac{P}{d'_3 N}$

For tests made at one third the top modulation frequency, the distortion margins should be increased by about 9·5 db.

Phase Linearity of Modulators, Demodulators and Associated I.F. Amplifiers

We have allowed a total intermodulation noise power per channel of 105 pW for phase distortion occurring in the modulator, demodulator and two associated I.F. amplifiers (high level and low level), shown in Fig. 15, ref. 6.

The noise allowance per unit is therefore 26 pW and if we assume that second and third order terms are equal to each other and predominant, we may write:

$$d'_2 = d'_3 = 13 \text{ pW } (-79 \text{ dbm}) \tag{32}$$

The distortion margins to be obtained in these units and on the above basis are given below.

TABLE VII

SECOND AND THIRD ORDER HARMONIC MARGINS FOR MODULATORS, DEMODULATORS AND I.F. AMPLIFIERS, FOR SIGNAL TO PHASE DISTORTION INTERMODULATION RATIO OF 79 DB, IN EACH UNIT

		Deviations kc/s rms.		Distortion margins db for ΔF at f_2		Allowable group delay variation	
N	f_2	δf	ΔF	F/H_2	F/H_3	$\tau_2/(B/2)$	$\tau_3/(B/2)^2$
60	300	200	604	67·7	70·0	0·417 mμ/sec Mc/s	1·360 mμ/sec Mc/s²
120	552	200	735	66·4	68·7	0·216	0·580
240	1052	200	875	65·0	67·2	0·112	0·255
600	2540	200	1260	64·1	66·4	0·035	0·055
960	4028	200	1520	63·7	66·0	0·020	0·025

Group Delay Variations

The performance of the various elements of the terminal equipment, as regards phase distortion, has been specified in terms of the harmonic margins of a test signal.

It is also possible to specify the required performance in terms of the group delay coefficients defined in equation (86) of Part II.

The coefficient τ_2 specifies the " linear " variation of group delay, within the half bandwidth B/2 (skewness of group delay curve) while τ_3 specifies the quadratic variation of delay over the band.

We may use equations (92) and (93) of Part II to correlate the distortion margins, referred to above, with the permissible values of the coefficients τ_2 and τ_3.

$$\frac{F}{H_2} = \frac{1}{\tau_2\,\omega\left(\frac{\Delta f}{B}\right)} \tag{33}$$

and

$$\frac{\tau_2}{(B/2)} = \frac{2}{\omega\,\Delta f\,(F/H_2)} \tag{34}$$

Similarly

$$\frac{\tau_3}{(B/2)^2} = \frac{4}{\omega \Delta f^2 (F/H_3)} \tag{35}$$

The permissible limits of τ_2 and τ_3, as calculated from (34) and (35), are given in the last two columns of Table VII.

Distortion Due to Weak Echoes

The case of intermodulation distortion arising out of feeder mismatches or weak echoes due to multipath propagation is analyzed in Part II for short echo delays. It is shown, therein, that intermodulation distortion arising out of feeder mismatches closely resembles distortion arising out of the phase non-linearity of transmission circuits.

The analysis of feeder distortion closely follows that given by Lewin[7] and it is extended to the case of intermodulation caused by large numbers of tones. The total intermodulation power of order n is found to be:

$$T''_n = \frac{2^{n-1}(n-1)!\, n^2 t''_n P^n}{3} \tag{36}$$

assuming an input signal spectrum extending from zero to a maximum frequency f_2 (in normalized units of angular frequency, 0 to 1) and where

$n^2 t''_n = n^{\text{th}}$ harmonic power in mW, produced by that sinusoidal input, which gives a fundamental output power of 1 mW, at the top frequency of the signal band.

P = total fundamental power output in mW.

The proportion of this power, falling within a 4 kc/s channel in a system carrying N channels, is

$$d''_n = \frac{T''_n F''_n(\omega)}{N} \tag{37}$$

$F''_{n(\omega)}$ is the spectrum distribution of intermodulation powers of order n and it is, in this instance, identical with $F'_{n(\omega)}$ of Fig. 7. Similarly the distortion margins for sinusoidal test tones, are identical with those given on page 145.

Allowable Short Feeder Mismatches for Typical Systems

We shall assume here that half the total intermodulation noise power generated by a transmitter-receiver pair is caused by feeder mismatches (echo distortion). Each feeder is therefore responsible for 31 pW of intermodulation noise, which we shall assume is equally contributed to by second and third order products. Thus:

$$d''_2 = d''_3 = 15 \text{ pW } (-78 \text{ dbm}) \tag{38}$$

A table of test tone distortion margins, applicable to echo mismatches, may be obtained from Table VII, by reducing all margins by 1 db.

The permissible feeder mismatches (or echo amplitude u) may be obtained by making use of equation (142) of Part II.

$$\left(\frac{P_{H_n}}{P_F}\right)'' = \frac{4u^2 n^2 \omega_f^2 \tau^{2n} \Delta' \omega^{2n-2}}{(n!)^2} \begin{Bmatrix} \sin^2\theta \\ \cos^2\theta \end{Bmatrix} \tag{39}$$

where

$\left(\frac{P_{H_n}}{P_F}\right)''$ = harmonic margin of order n

u = $r_1 r_2 e^{-2\alpha l}$ = echo amplitude.

r_1, r_2 = reflection coefficients of load and generator respectively.

$e^{-\alpha l}$ = one-way feeder attenuation.

τ = one-way feeder delay. (Half echo delay).

$\Delta'\omega$ = peak angular frequency deviation of test tone.

$\Delta\omega$ = rms angular frequency deviation of test tone.

ω_f = angular frequency of test tone.

θ = phase angle depending on the carrier frequency, the phases of the generator and load reflection coefficients. θ will be assumed to have an average value of 45°.

We may obtain from equation (39) and Table VII, values for $\sqrt{u} = \sqrt{r_1 r_2 e^{-2\alpha l}}$ which is equal to the reflection coefficients at both ends of the feeder, on the assumption that they are equal and that the feeder attenuation is negligible.

TABLE VIII

ALLOWABLE FEEDER MISMATCHES FOR SIGNAL TO SECOND AND THIRD ORDER ECHO DISTORTION RATIOS OF 78 DB

Echo delay 2τ = 100 mμ/secs.

N	L	f_2	Δf	F/H_2	F/H_3	$\sqrt{u}$
60	9·6	300	604	66·7	69·0	11·3%
120	11·3	552	735	65·4	67·7	8·1
240	12·8	1052	875	64·0	66·2	5·8
600	16·0	2540	1260	63·1	65·4	3·3
960	17·6	4028	1520	62·7	65·0	2·5

An echo delay of about 100 mμ/sec is produced by an air spaced coaxial line 50 ft. in length. For larger delays the reflection coefficient is reduced accordingly, since $r\tau$ is a constant, up to the point at which the approximations made in the derivation of equation (39) cease to hold.

Intermodulation Distortion caused by Mismatches in Long Feeders

If the echo delays are long, the approximations made above are not valid.

The general case is treated in Part II and the result is in agreement with Lewin's[8] who used a somewhat different approach. It is found that as the feeder length increases the total amount of intermodulation noise in a given channel at first increases and then decreases. This appears to be due to the fact that with increasing feeder length, or alternatively, deviation, the high order distortion terms become progressively predominant. Since however the spectrum distribution of distortion powers contains the factor ω^2, the bulk of the distortion is at high frequencies, and in the case of the higher orders, well above the signal band. Thus although the total amount of distortion power increases with increasing delay, or deviation, the proportion of this power, falling within the signal band is reduced.

The total interference power d'' of all orders, in a channel is:

$$d'' = \frac{u^2}{4}\left(\frac{f}{\Delta f}\right)^2 \frac{P}{N}\Phi(\tau\Delta\omega) \quad (40)$$

where: $u = r_1 r_2 e^{-2\alpha l}$ = echo amplitude.

f = mean frequency of channel in which intermodulation noise is measured.

P = total signal power.

N = number of channels.

$\Phi(\tau\Delta\omega)$ = Echo distortion function. Function of half echo delay τ, rms angular deviation $\Delta\omega$, and channel mean frequency f. $\Phi(\tau\Delta\omega)$ is plotted in Fig. 8 for the top channel, and for the middle channel.

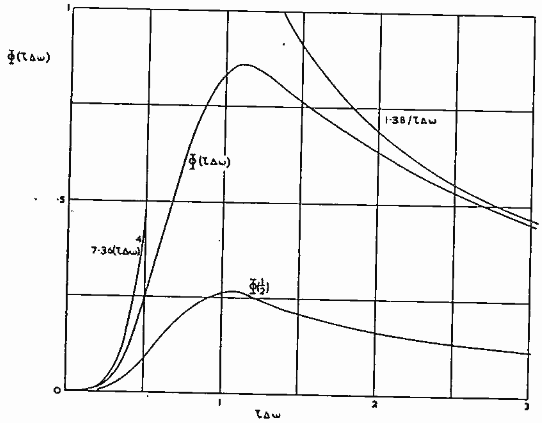


FIG. 8

Intermodulation power in the top channel and in the middle channel due to weak echoes.

Allowable feeder mismatches may now be calculated with the aid of equation (40), on the assumption that $d'' = -75$ dbm. The result of this calculation is given in Figs. 9 and 10, for echo delays of 100, 200 and 400 mμ/secs. The allowable feeder mismatches have, for a given system, a minimum value, which must not be exceeded,

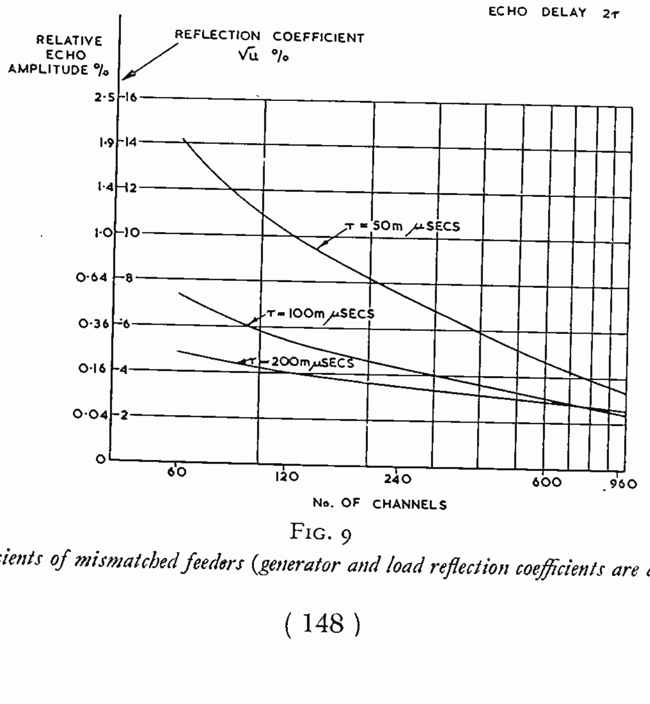


FIG. 9

Reflection coefficients of mismatched feeders (generator and load reflection coefficients are assumed equal).

when the echo delay has a certain critical value. This minimum value is given in Fig. 10, together with the corresponding echo delay. An increase or decrease of echo delay from this critical value should cause a reduction of intermodulation noise.

Apart from the dependence of d'' upon $\Delta\omega$ through the function $\Phi\ (\tau\Delta\omega)$, d'' is inversely proportional to $(\Delta\omega)^2$. Group delay distortion and amplitude distortion

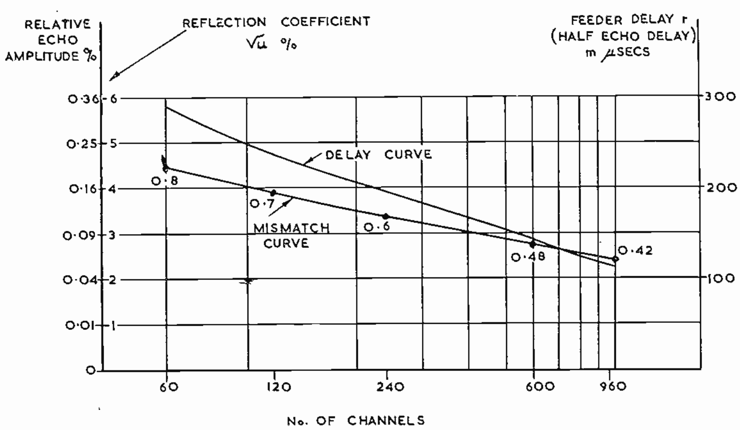


FIG. 10

Reflection coefficients of mismatched feeders for worst echo delay (generator and load reflection coefficients are assumed equal).

on the other hand increase with the deviation. Thus there exists, for any given system, an optimum deviation, at which the total intermodulation noise is at a minimum.

Reflection Coefficient " Bandwidth "

The calculations of Part II have been based on the assumption that the reflection coefficient at the generator and load ends are constant with frequency. In other terms it has been assumed that the echo amplitude and phase do not vary over a small bandwidth around the carrier frequency.

It is obvious however that the magnitude of the reflection coefficients of the elements terminating both ends of the feeder, measured at frequencies far away from the unmodulated carrier frequency, can have no influence on the intermodulation distortion.

It follows that the definition of an allowable feeder mismatch is not complete, unless the bandwidth over which this mismatch value holds is also specified.

Stewart[9] has calculated the power spectrum of a carrier, frequency modulated by random noise. We can, following his method, determine the bandwidth containing all but 1% or all but 0·1% of the total carrier energy, for given systems.

If we use the system parameters given in Table IV, we obtain the curves of Fig. 11.

Curve 2 of Fig. 11 may be taken as defining approximately the bandwidth over which the mismatch figures of the previous paragraph, have to be obtained. This conclusion may not be strictly true. As a check a direct calculation of intermodulation noise needs to be made, for the case where the reflection coefficient varies over the signal band according to a predetermined law.

Testing Methods

A precise measurement of intermodulation arising in systems and components is quite a complex operation which necessitates the development of a large amount of intricate test equipment.

The ultimate test of the completed link is under operation conditions with actual traffic loading. This test cannot, in general, be made under laboratory conditions as it would require the provisioning of the necessary channelling equipment. This equipment is very expensive. It follows that other methods of test must be provided.

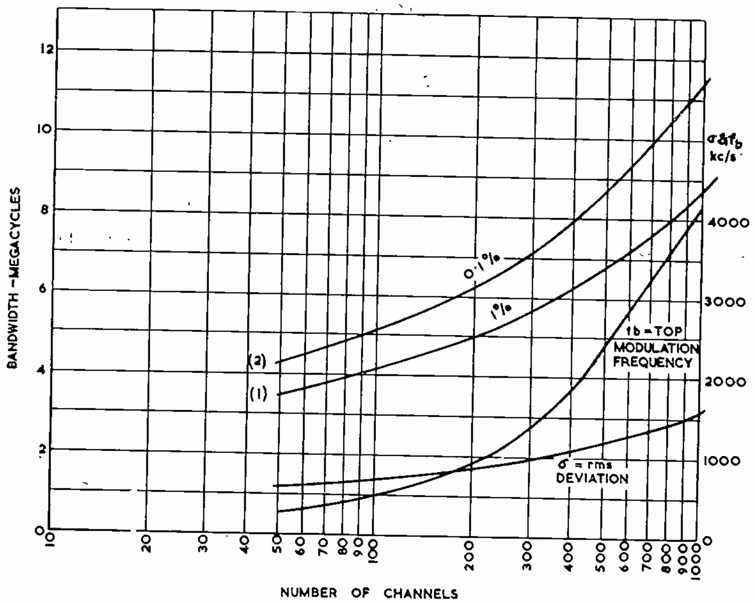


Fig. 11

Energy bandwidth of multichannel systems, modulated by random noise.

System Testing

System testing may be done by measuring the harmonics of sinusoidal tones injected into the system. Two tone methods may also be used with advantage if it is required to test the system at the highest modulation frequencies. The testing of a large system with sinusoidal tones is not satisfactory for several reasons. Echo distortion is characterized by a very rapid variation of the harmonics of a single tone with the frequency of the tone. It is also found, in practice, that the level of the harmonics of a single tone varies very rapidly with the detuning of the carrier frequency.

A better way of testing multichannel radio links consists in applying to the system a uniform band of noise (white noise) covering the whole of the baseband, to the exclusion of a few narrow intervals which simulate vacant channels. This method of testing multichannel systems is now widely accepted as simulating, as closely as is possible, actual traffic conditions.

A white noise tester (Fig. 12) consists of a wideband noise generator, low pass and high pass filters to limit the noise spectrum to the actual boundaries of the baseband, and band stop filters, inserted between the noise generator and the system under test, in order to remove the noise loading from a number of narrow bands, and to simulate vacant channels.

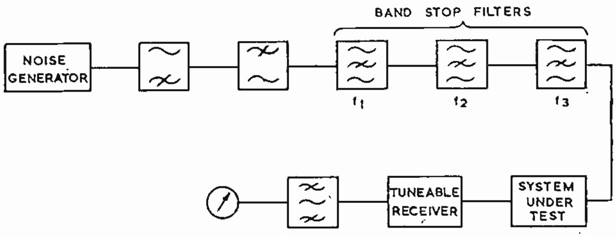


FIG. 12

Block diagram of white noise tester.

A tunable receiver, at the output of the system, measures the level of the applied noise and the noise level in the vacant bands. The bandwidth of the receiver is preferably made equal to 4 kc/s and independent of the signal frequency.

A white noise tester must be capable of measuring the valve and crosstalk noise produced by modem (modulator and demodulator back to back) as well as extensive systems. We have seen on page 134 that the ratio of intermodulation and valve noise to test tone for modem units may be as low as —64 db, reference 1 mW, while the valve noise alone is —67 db reference 1 mW (distribution A).

The noise input into the system has a total power P which depends on the number of channels. This noise is spread over the whole of the baseband, so that the applied noise per channel is P/N at a point of zero reference level. This noise P/N is given below as a function of the number of channels.

Number of Channels	Mean multi-channel signal power	Average Noise Power per Channel	Ratio R of noise in traffic Channel to noise in vacant Channel		
			Modem alone		Reference link
N	P	P/N	Without traffic	With traffic	
60	9·6 dbmo	— 8·2 dbmo	58·6 db	55·6 db	49·6 db
120	11·3 ,,	— 9·5 ,,	57·3 ,,	54·3 ,,	48·3 ,,
240	12·8 ,,	—11·0 ,,	55·8 ,,	52·8 ,,	46·8 ,,
600	16·0 ,,	—11·8 ,,	55·0 ,,	52·0 ,,	46·0 ,,
960	17·6 ,,	—12·2 ,,	54·6 ,,	51·6 ,,	45·6 ,,

The ratio R of noise outside the vacant channel to noise inside the vacant channel is given in columns 4 and 5 above, for noise distribution A.

If we wish to measure this ratio R with an accuracy of 0·1 db, the residual noise powers due to breakthrough in the bandstop filter and to the receiver valve noise should each be 20 db down. It follows that the bandstop filter attenuation should be 78·6 db.

The recommended channel level at the output of the radio system (CCIR Study Group IX, Document 69E revised), is -15 dbm, so that the noise level of the receiver, measured in a 4 kc/s band, should be less than about -102 dbm.

Component Testing: Two-tone testing

The performance of the main components of a link has been specified in terms of the harmonics of a single sinusoidal tone.

The measurement of the linearity of individual components, in terms of the harmonic power of a single tone, is not reliable, as it is difficult to separate the distortion produced by the component under test from the distortion produced by the associated equipment. For example, a discriminator may be tested for linearity by measuring the harmonics of a single tone. This method of test requires the use of a frequency modulator and the distortion of the latter affects the overall distortion.

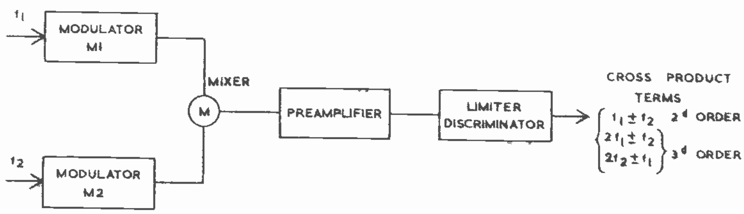


FIG. 13

Block schematic of instrument used for demodulation testing.

One way of overcoming these difficulties is to measure the crossmodulation products of a pair of tones. It is then possible to use two modulators, each one of which is modulated by a single tone. Harmonics of each tone produced by the modulators, provided they are not too large, do not affect the intermodulation products. A typical scheme for testing demodulators is shown in Fig. 13. The only components in the system, apart from the one under test, liable to produce crossmodulation terms are the mixer and the preamplifier and special precautions have to be taken to reduce this disturbing factor to a negligible level.

Again, the harmonics of single tones do not always fall within the signal band and consequently it is not always possible to measure them. Thus the maximum modulation frequency at which the second harmonic may be measured is half the top modulation frequency, whereas no such limitation exists with two tone testing.

We shall assume in what follows that two tone testing is done with two tones of equal amplitude v_2 equal to half the amplitude of the tone used in single tone testing. The total fundamental power in the latter case is P_F, while in the former the fundamental power of each tone is P_{F_1}, and the power of each crossmodulation term, of order n, is $(P_n)_1$.

It is shown in Part II that, for the case of amplitude distortion only

$$\frac{(P_2)_1}{P_{F_1}} = 4\, t_2\, P_{F_1}$$

$$\frac{(P_2)_1}{P_{F_1}} = t_2\, P_F = \frac{P_{H_2}}{P_F}$$

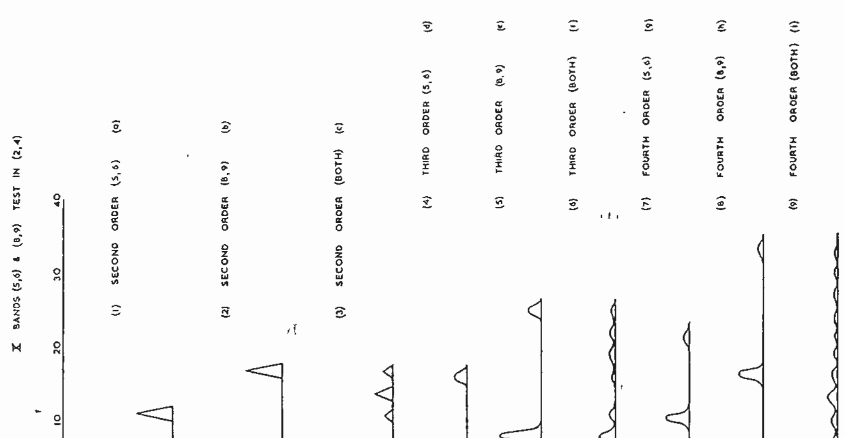


FIG. 14
Intermodulation spectra for two equal signal bands 5, 6 and 8, 9.

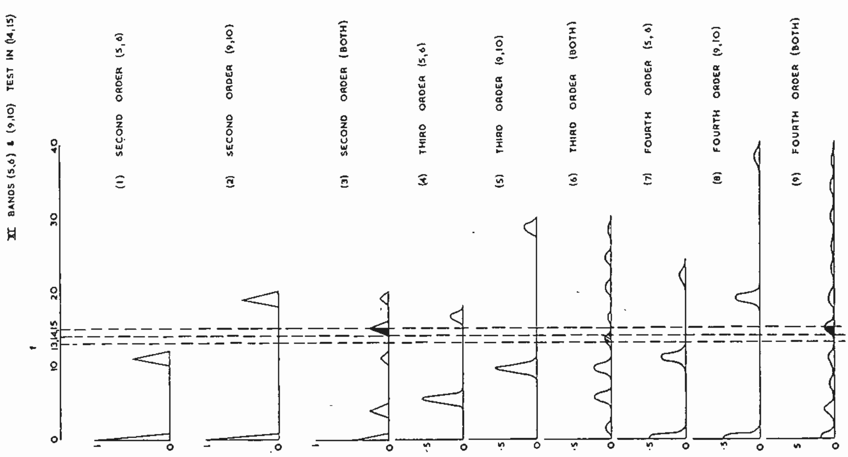


FIG. 15
Intermodulation spectra for two equal signal bands 5, 6 and 9, 10.

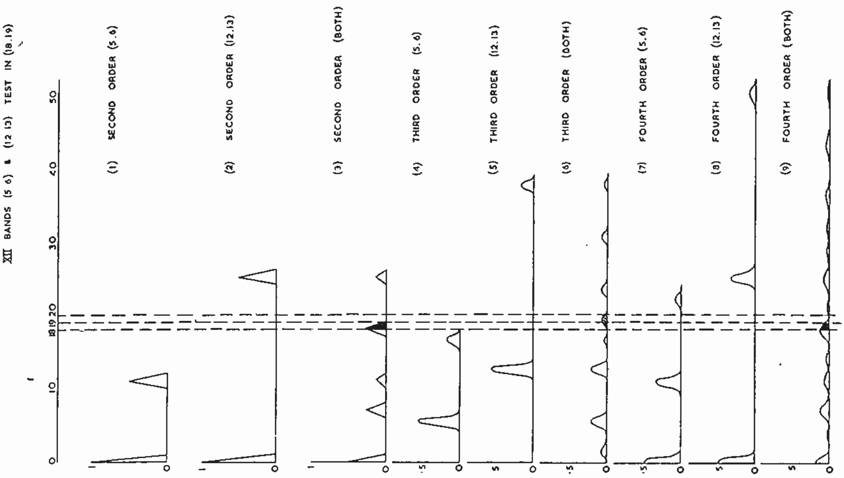


FIG. 16
Intermodulation spectra for two equal signal bands 5, 6 and 18, 19.

Thus the second order intermodulation margin is equal to the second order harmonic margin.

Similarly

$$\frac{(P_3)_1}{P_{F_1}} = 9\, t_3\, (P_{F_1})^2$$

$$\frac{(P_3)_1}{P_{F_1}} = \frac{9}{16}\, t_3\, P_F{}^2 = \frac{9}{16}\, \frac{P_{H_3}}{P_F}$$

The third order intermodulation margin is 2·5 db less than the third order harmonic margin.

The case of phase distortion is very similar. We have in this case

$$\frac{(P'_2)_1}{P_{F_1}} = 4\, \omega^2\, t'_2\, P_{F_1}$$

$$\frac{(P'_2)_1}{P_{F_1}} = \omega^2\, t'_2\, P_F = \frac{P'_{H_2}}{P_F}$$

For third order distortion we have

$$\frac{(P'_3)_1}{P_{F_1}} = 9\, \omega^2\, t'_3\, P_{F_1}{}^2$$

$$\frac{(P'_3)_1}{P_{F_1}} = \frac{9}{16}\, \omega^2\, t'_3\, P_F{}^2 = \frac{9}{16}\, \frac{P'_{H_3}}{P_F}$$

ω here is the angular frequency of the harmonic or the interaction term.

Two-tone methods of measuring the linearity of modulators, demodulators and transmission systems, such as repeaters, are fully explained in reference 6.

Two-tone testing with pure tones is not always reliable, because of cancellation within a given unit. It appears possible, however, when measuring the linearity of discriminators, or receivers in general, to use noise as a test signal, while at the same time avoiding the ambiguity caused by distortion arising in the modulator.

The method consists in using two oscillators, as in Fig. 13 which are both frequency modulated with white noise, covering two non-overlapping bands A and B, only one band being applied to each modulator. The noise bands A and B are selected in such a way that the distortion spectra of the modulators have overlapping distortion-free zones. The discriminator, which accepts noise modulation in both bands, generates intermodulation noise covering the distortion-free zones of the modulators.

It is shown in Part II that by selecting noise bands, covering frequencies 5 to 6 and 8 to 9, a distortion-free zone exists in the band 2 to 4, within which the intermodulation noise, arising out of amplitude non-linearity in the receiver, may be measured. The distortion spectra applicable to three sets of noise bands, and arising out of amplitude non-linearity, are given in Figs. 14 and 15, 16. The intermodulation noise falling within the band 2 to 4 is mainly second and fourth order.

Acknowledgment

The calculations on which Figs 3 to II and I4 to I6 have been based are detailed in Part II of the present article, which has been written by Mr. D.S. Palmer and the Author.

References

(1) C.C.I.F. XVI. Assemblee Plernière. Firenze, 22-27 Octobre 1951. Tome III bis. Transmission sur les lignes, Maintenance. Page 120.

(2) Load rating theory of Multi-channel Amplifiers by B. D. Holbrook and J. T. Dixon. B.S.I.J., 1939.

(3) Non-linear distortion in transmission systems by R. A. Brockbank and C. A. A. Wass. J.I.E.E. Vol. 92, Pt. 3, 1945.

(4) Ref. 1, p.261.

(5) The fundamental principles of frequency modulation by B. Van der Pol. J.I.E.E. 1946, 93, Pt. III.

(6) Wide band Microwave Radio Links, by S. Fedida, *Marconi Review* 118, 3rd Quarter, 1955.

(7) Phase Distortion in Feeders by L. Lewin, J. J. Muller and R. Basard, *Wireless Engineer*, Vol. 27, May, 1950.

(8) Interference in multi-channel circuits by L. Lewin, *Wireless Engineer*, Vol. 27, December, 1950.

(9) The Power Spectrum of a Carrier Frequency Modulated by Gaussian Noise by J. L. Stewart. Proc. I.R.E., Vol. 42, October, 1954.

(10) Circuit Technique in Frequency Modulated Microwave Links by H. Grayson, T. S. McLeod, R. A. G. Dunkley and G. Dawson, J.I.E.E., April, 1952.

BOOK REVIEW

* *Practical Electroacoustics*, by M. Rettinger. Published by Thames & Hudson. 63/-.

This book*, written by a member of the engineering department of R. C. A. Victor, covers a wide field of subjects all related to the science of electroacoustics. They are dealt with in a very practical and lucid way, mathematics being introduced where necessary to explain the function and aid in the design of equipment and installations.

The subjects covered range from microphones and loudspeakers to attenuators, public address systems and magnetic recording. The chapter on architectural acoustics, written with the author's evidently wide experience in this field, should prove invaluable to those who are not well up in this subject; while that on magnetic structures clarifies the rather bewildering assortment of magnetic units found in various text-books on the subject. Methods of measurement are discussed throughout the book, and practical advice is given on the interpretation of results. The subject matter is well supported by graphs and practical data.

The publication should prove very useful both as a text and reference book to engineers and students engaged in broadcasting, public address and sound reinforcement projects. The book is well indexed, and a good bibliography is included, but the price seems rather high for only 270 pages.

BOOK REVIEWS

Fundamentals of Transistors, by L. Krugman. Chapman and Hall, 21/-.

Transistor Audio Amplifiers, by R. F. Shea. Chapman and Hall, 52/-.

The first of these books, though of a type common twenty-five or thirty years ago when the subject matter was the applications of thermionic valves, may be something of a novelty to the present generation, for this John F. Rider publication is a successful attempt to instruct the reader in the art of using transistors in circuits in simple language without recourse to the jargon of physics and mathematics which sometimes completely overawes the prospective users. Except for a bare seven pages devoted to the elementary physics of the device, the book deals with the properties of transistors as circuit elements in the three basic modes of connection (depending upon which electrode be earthed) and proceeds then to show how they may best be used in amplifiers, oscillators, trigger circuits and, more briefly, in other applications.

Unlike a valve, the transistor, whatever the mode of connection, loads the circuit preceding it and further, the output load is " seen " at the input and vice versa; these features necessarily complicate the circuit arithmetic which becomes tedious rather than difficult, but one is led step by step through the calculations from the basic four-terminal equivalent network (using z or y parameters) to the determination of the operating characteristics, the calculation being supplemented by curves which present the circuit information perhaps more effectively than figures and formulae.

The author, L. Krugman, of the U.S. Signal Corps Engineering Laboratories, is to be congratulated on the way in which he has presented the subject matter. This is in no sense a " popular " publication but an extremely practical handbook for use in the laboratory or at the bench by all who may desire to use transistors. No particular products are referred to in the text but full use is made of generalized straight-line characteristics which can readily be adapted to specific types. The book has been well reproduced by a photo-litho process, both text and diagrams being good. At a guinea the value for money is undeniable.

Mr. Shea's book (a John Wiley publication) is in the nature of a companion volume to his Principles of Transistor Circuits and deals solely with practical aspects of the subject indicated by the title and is consequently much more specialized than Mr. Krugman's. The author favours, and uses almost exclusively, the representation of transistors by their hybrid (h) parameters, claiming they are less dependent upon one another, are generally easier to measure and show less variation with operating point and frequency. Sets of curves show the variations to be expected with operating conditions, with frequency and with temperature; these data depict the mean values of a large number of units of (American) General Electric manufacture whilst the tabulated specifications of similar units include anticipated tolerances based upon production spreads, the importance of which in any design are stressed repeatedly. Finally full specifications of units produced by five other (American) manufacturers complete this section of the book.

A discussion of the three basic configurations or modes of connection with particular stress upon parameter stability leads to a consideration of their combination with either transformer or R.C. coupling, the very full sets of curves presented being summarized in tabular form to stress the features of the various connections. The radical difference between valves and transistors is perhaps best exemplified where the low level stages of hearing aid, phonograph or microphone amplifiers are considered since special problems are met in relation to noise, impedance levels and distortion; this is brought out well in the section on volume controls which, whilst performing their function, are required to maintain impedance match, bias stability and reasonable frequency response. In the higher level stages, where the peak signals may approach the applied bias, stress is laid on the advantage of using graphical methods, especially in Class B operation which is particularly suited to transistors on account of the high efficiency attainable with low quiescent power drain, albeit at the expense of some slight distortion; although feedback is offered as a possible cure for the latter the author is rather cautious here. As examples of the applications of the techniques discussed a few typical audio amplifiers are finally described together with performance data. A bibliography devoted mainly to applications completes a book which includes nearly 200 figures and the format of which leaves nothing to be desired.

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