

The

# Marconi Review

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## Introduction

THE essential requirements which must be met by the design of equipment designed to function as an effective aid to Air Traffic operations are adequate performance, and reliability in the engineering sense combined with economy in operation and maintenance.

In the case of radar equipment it is a sad fact that the pursuit of threshold signals has prejudiced these essential engineering requirements and has resulted in a service figure of merit which has fallen far short of that with which we are familiar in the established field of ground communication equipment.

The Marconi S.232 radar is the outcome of an engineering project which has sought to correct this position by applying radio engineering principles already firmly established in the communications field to the design of a radar of greatly improved reliability intended to meet the needs of a busy Airport Terminal.

## Selection of Parameters

The intensification of the air control problem at major terminals due to the increasing volume of traffic suggested the need for a general purpose radar capable of providing in one equipment the requirements of the Surveillance Radar Element of G.C.A., the Terminal Area Radar and the Long Range Radar. To meet

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## MARCONI COHERENT M.T.I. RADAR ON 50 cms.

BY E. EASTWOOD, PH.D., M.Sc., M.I.E.E.,

T. R. BLAKEMORE, and B. J. WITT

*The following paper was originally delivered at a symposium on Flight, Weather and Astro-Radiolocation held by the Committee for Radiolocation under the auspices of Gesellschaft zur Forderung des Verkehrs. V. in June 1955 at Munich. We are indebted to Professor Leo Brandt, editor of the Journal " Strecken-Anflug-und Landefunkortung " for permission to publish it in these pages.*

### Introduction

THE essential requirements which must be fulfilled by any piece of equipment designed to function as an effective aid to Air Traffic operations are adequate performance, and reliability in the engineering sense combined with economy in operation and maintenance.

In the case of radar equipment it is a sad fact that the pursuit of threshold signals has prejudiced these essential engineering requirements and has resulted in a service figure of merit which has fallen far short of that with which we are familiar in the established field of ground communication equipment.

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### Selection of Parameters

The intensification of the air control problem at major terminals due to the increasing volume of traffic suggested the need for a general purpose radar capable of providing in one equipment the requirements of the Surveillance Radar Element of G.C.A., the Terminal Area Radar and the Long Range Radar. To meet

such a comprehensive cover specification for the wide variety of aircraft now operating would require good resolving power, high gain and large transmitter power, and the use of an X- or S-band magnetron system obviously suggests itself. Experience of precipitation effects at these frequencies, however, i.e. obscuration by back scatter from rain or cloud, immediately opposes such a choice of wavelength. Scatter follows the Rayleigh inverse fourth power of wavelength law, and this suggests that a wavelength considerably in excess of 10 cm. is to be preferred, for although some reduction of rain returns can be obtained by use of polarization effects, such devices produce signal attenuation and it is obviously better to avoid rain echoes altogether. The requirement for simultaneous observation at short range requires that the radar be free from ground clutter. Ground returns normally increase with reduction of wavelength and show, moreover, marked fluctuation in phase and amplitude. The difficulty of eliminating such effects by use of M.T.I. techniques is therefore greater at shorter wavelengths. This consideration points to the use of a longer wavelength where the use of fully coherent M.T.I. technique as opposed to the microwave Coho Stalo method becomes possible; variation in the equipment factor is then reduced and the random variation in the clutter is reduced with a consequent substantial improvement both in cancellation and subclutter visibility.

The requirements for aerial gain and resolution however militate against the use of a really long wavelength since the aerial structure would become too large and expensive; it was concluded that a useful compromise choice for  $\lambda$  would be in the order of 50 cm.

For this wavelength an aerial aperture of  $20 \lambda$  would yield a  $4^\circ$  beamwidth which experience has proved to be effective for both control and tracking. This selection led to a simple and economical antenna design consisting of a paraboloidal reflector fed by an offset horn (Fig. 4). By suitable choice of the height of the array and its tilt angle the vertical polar diagram developed by partial reflection from the airfield surface can be arranged to provide comprehensive cover for both short and long range work.

The choice of 50 cm. was further prompted by the circumstance that power triodes suitable for pulse operation are available for this wavelength, and this has permitted the development of a pulse transmitter comparable in performance and reliability to a communication unit and constitutes a marked improvement on a magnetron system.

In an article by Emslie and McConnell<sup>1</sup> (1947), it is shown that coherence for M.T.I. purposes is best achieved by I.F. Locking with I.F. addition. This feature is adopted in the present system with the additional advantage that I.F. Locking by the oscillator instead of by the transmitter, as in the Coho Stalo arrangement, is now possible since the present triode transmitter is simply a crystal controlled power amplifier.

The choice of  $\lambda$  and the crystal driven triode amplifier have allowed the use of hard valve modulators and crystal control of the recurrence frequency. Both these design features have tended to reduction in pulse to pulse jitter, which, taken in combination with the virtual elimination of frequency fluctuation from local and reference oscillators; have led to substantial improvements in M.T.I. performance.

The duplexer and receiver system have also profited greatly from the use of a longer wavelength, since receiver valves capable of R.F. amplification and possessing good noise figures are available to replace the crystal of the microwave system, giving an overall noise factor of 9 db. The protection of such a valve receiver makes less

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<sup>1</sup>Vol. I. Radiation Laboratory Series, McGraw Hill.

exacting demands on the duplexer while the fully coherent system of transmission and reception eliminates the need for A.F.C. and ensures receiver simplicity and constancy of performance.

It will be seen that the choice of the 600 Mc/s band has had far reaching design consequences which have resulted in excellent radar cover both at short and long range and in all weather conditions. Stable radar and M.T.I. performance free from apparatus fluctuations have also been achieved. General engineering and the conservative rating of valves and other components have been similarly facilitated with consequent improvement in reliability, ease of maintenance and economy of operation.

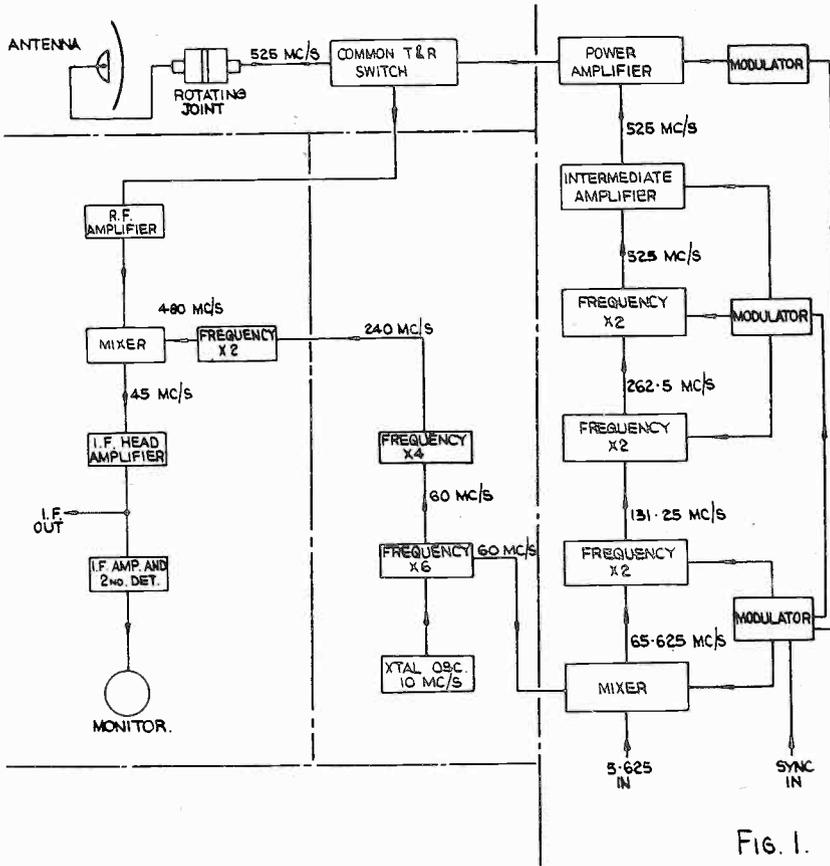


FIG. 1

S.232 Transmitter/Receiver System.

### Description of System

The elements of the system may be briefly and conveniently described by reference to Figs. 1 and 2 which give schematic arrangements for the Transmit/Receive and Cancellation portions of the system respectively.

The coherent I.F. reference signal which is the basis of the system, is obtained by multiplication ( $\times 8$ ) from the Master Crystal Oscillator of 5.625 Mc/s located in the

Radar Distribution Unit. A second output from the Master Oscillator is fed into an anode modulated, balanced mixer which is the first stage of the transmitter proper. This mixer also receives a signal of 60 Mc/s, which determines the Stable Local Oscillator frequency, derived from a 10 Mc/s crystal contained in the Receiver Type SR.300. The 65.625 Mc/s derived from the mixer is then fed into a succession of three anode modulated doublers, the final output at 525 Mc/s serving as drive to a pair of power amplifiers in series. These latter units and the pair of preceding doublers employ CV.436 valves, in a grounded grid connection using coaxial lines which are  $\lambda/4$  long for the first doubler, and  $3\lambda/4$  for the final high frequency doubler and amplifiers. Matching between stages is achieved by the use of line stretchers which are tapped into the cathode input circuit of the succeeding stage at the appropriate impedance points.

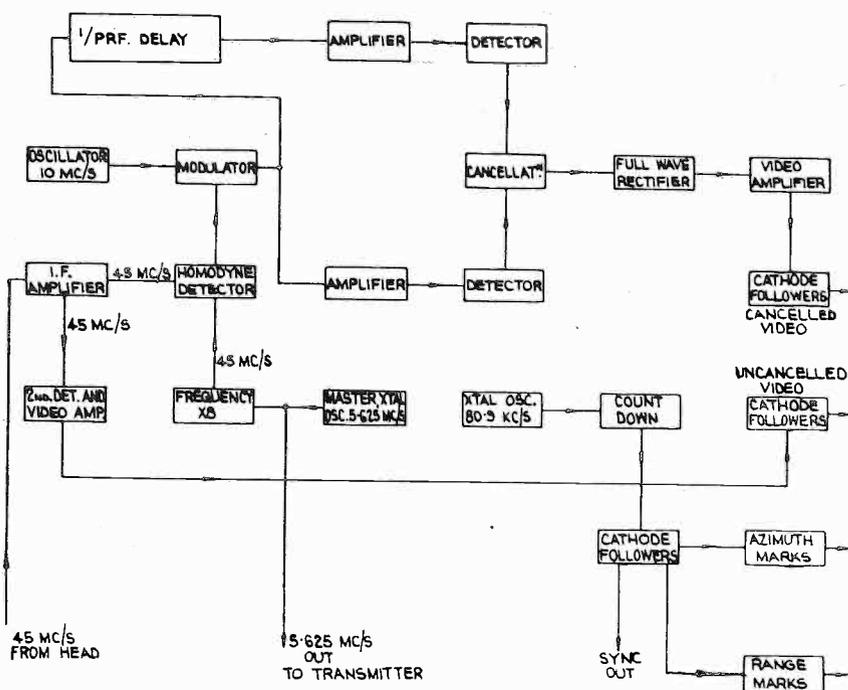


FIG. 2

S.232 Cancellation System.

The modulators indicated in Fig. 1 employ hard valves (paralleled CV.398s giving 12 K.V. pulse) as switching elements which are opened by pulses delivered from an L.C. pulse-forming line contained in the circuit of a CV.173 triggered by a synchronizing pulse from the Radar Distribution Unit (SJ 400). Three pulses are derived from the line of duration 9, 6 and 4 or 2  $\mu$ secs. respectively, the first of which is applied to the mixer and first doubler and provides for the  $1\frac{1}{2}$   $\mu$ sec. build up time required by these units. The 6  $\mu$ sec. pulse is contained within the previous 9  $\mu$ sec. and serves to switch the modulator driving the three following stages. The 4  $\mu$ sec. pulse is similarly wholly contained within the 6  $\mu$ sec. pulse and applied to the modulator feeding the final amplifier. Thus the build up time of the radiated pulse is

dependent only on the output amplifier; a good pulse shape is therefore obtained free from jitter and the cancellation achievable is improved accordingly. The final pulse length is switchable between 2 and 4  $\mu$ secs.

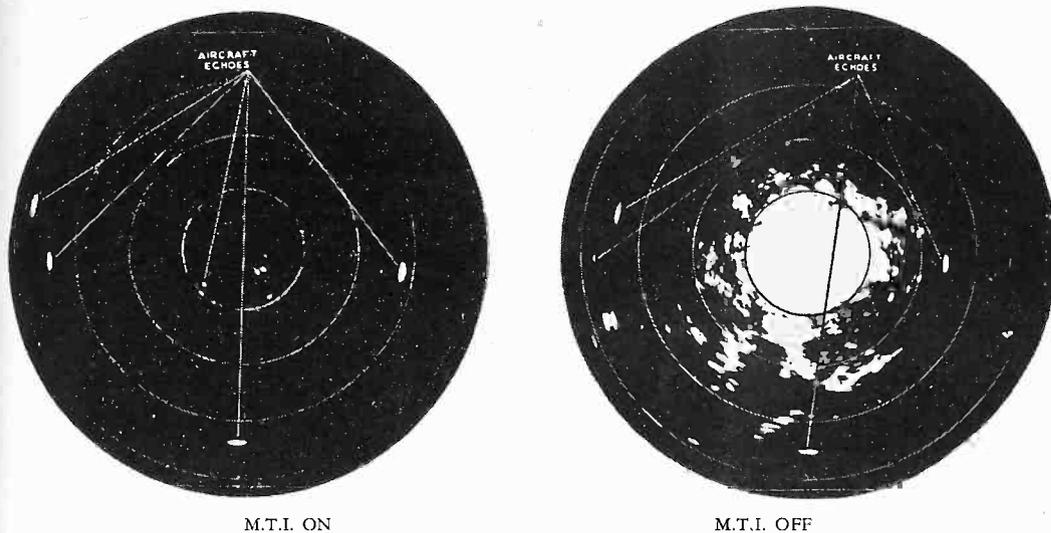


FIG. 3

Marconi Radar Type S.232 P.P.I. Display Photographs. (Range rings at 5 nautical mile intervals.)

The 60 KW output pulse is fed through the coaxial switch unit comprising T.R. and A.T.R. sections each using a 1 B 23 cell, to the coaxial rotating joint and so passes by cable to the probe drive of the feed horn.

The 60 Mc/s Stable Local Oscillator frequency after multiplication by 8 is fed to the mixer which receives signals from the R.F. Amplifier (CV.354 grounded grid circuit, Noise Figure 8.5 db) and so delivers 45 Mc/s I.F. both to the A-scope Monitor of the Receiver rack and to the cancellation units of the Radar Distribution Unit. After I.F. amplification the signal is rectified to provide uncancelled video for selection by the various displays. The amplified I.F. from the main I.F. amplifier is also fed into the homodyne detector where it is mixed with the output from the coherent reference oscillator, also at 45 Mc/s. The homodyne detector functions both as an amplitude and phase detector and yields an output which is a function both of the phase and amplitude of the input signal. For a permanent echo the output signal is constant but the echo from a moving target is modulated in amplitude at the Doppler frequency which is preserved in the heterodyne process.

The main I.F. amplifier is provided with swept gain to prevent overloading by strong signals at short range and also with I.A.G.G. which provides some measure of signal compression that assists in preserving the normal noise background on the P.P.I. within a region of cancelled clutter. This improves the subclutter visibility of the system.

The output from the homodyne detector modulates the carrier from a 10 Mc/s oscillator which is then fed to the driver valve of a supersonic delay line. The output from this unit after amplification and detection is added in opposite polarity to a second and undelayed signal delivered by the homodyne mixer through a parallel

amplifier-detector chain. The resulting residual signal from an aircraft is of bipolar type and fluctuates in amplitude at the Doppler frequency but the signal from a stationary target is of fixed amplitude. A full wave rectifier is therefore employed to provide the final "cancelled" signal to the video amplifier for distribution through

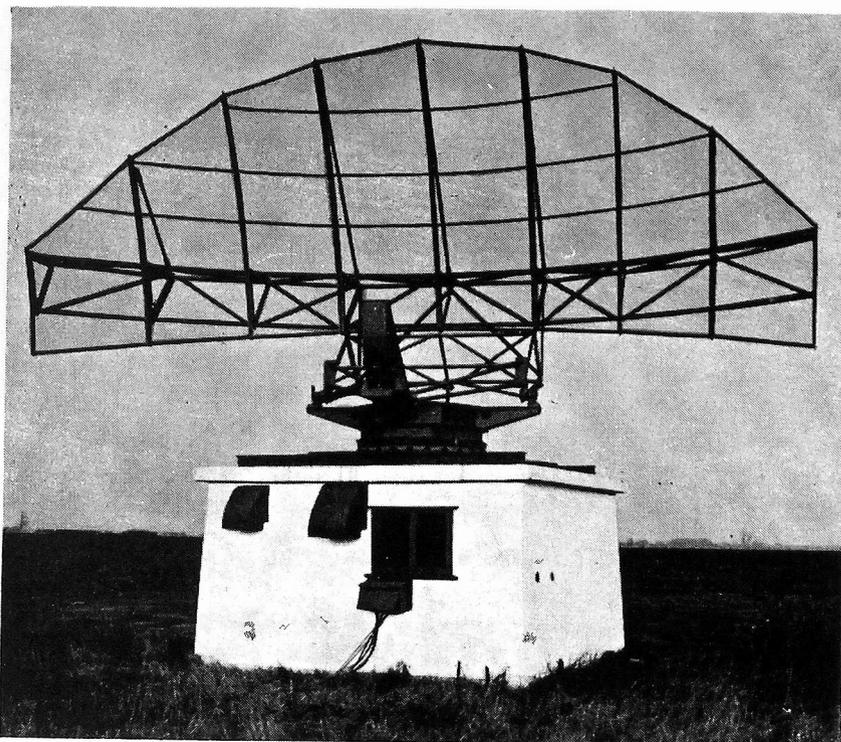


FIG. 4  
*Marconi Radar Type S.232. General view of aerial head.*

cathode followers to the displays. Both cancelled and uncanceled signals are available at every display unit.

The supersonic delay line uses X-cut quartz crystals as driver and receiver elements of the longitudinal wave transmitted through water contained in metal tubes of  $1\frac{1}{4}$ " diameter. This water line is maintained at the temperature of  $72^{\circ}$  by a simple thermostat; this temperature is the turnover point of the temperature/delay characteristic and ensures maximum stability. The cell operates at a frequency of 10 Mc/s and possesses a bandwidth of 350 kc/s. The slight pulse distortion which occurs is matched by appropriate design of the amplifier and filter associated with the undelayed channel and ensures zero residual of a cancelled test signal.

The recurrence frequency of the system is determined by a crystal oscillator of 80.9 kc/s which is counted down to yield an interpulse period and the transit time of the delay cell is adjusted to match. Experiments have been made with pulse regeneration circuits that lock the recurrence pulse to the cell itself but no superiority over the present simple system has been observed. The present arrangement permits full utilization of the variable p.r.f. feature and has been found to deliver a jitter free

pulse that permits optimum cancellation to be obtained. The synchronizing pulse itself is derived from the back edge of the 40  $\mu$ sec. pulse delivered by the Recurrence Pulse Generator. The leading edge of this pulse is used to operate the A.G.C. circuits associated with the delayed and undelayed 10 Mc/s channel amplifiers.

**Discussion of Performance**

The general performance of the system is well illustrated by the P.P.I. pictures reproduced in Fig. 3 which permits direct comparison with the M.T.I. and non-M.T.I. records. The cancellation of clutter is seen to be good while the visibility of the moving targets is in no way impaired. Tracking down to a range of 1 km has proved to be possible while aircraft of medium radar cross section are continuously plotted to ranges of over 150 km. General performance is greatly assisted by the favourable perception factor of the display unit SD.700 due to the 12-inch fluoride tube type CV.2162 and the associated circuit design. The signal is readily distinguishable with the aerial beamwidth and pulse duration employed, which, nevertheless, give adequate discrimination.

Measurement of the various residual instrumental instabilities, including T.R. cell effects, have shown that the limitation on cancellation due to these causes may be set at 40 db. The scanning limitation for the worst condition of operation of the system is of the order of 32 db corresponding to an aerial scan rate of 10 r.p.m., recurrence frequency 560 p.p.s. and 60 pulses per beamwidth. It will be seen that the scanning effect represents the major limitation on M.T.I. performance but this figure can be greatly improved by using a scan rate of 6 r.p.m. together with the increased repetition rate of 800 p.p.s. which is permitted by the recurrence frequency generator and the variable delay cell. Experiments have been performed with a double cancellation arrangement but for the present beamwidth this expensive refinement is unnecessary and it has proved more effective to design the antenna for low side lobes (24 db) rather than to seek for reduced beamwidth since the present resolution is already adequate. No difficulty has been experienced with blind speeds although the variable P.R.F. and moveable radio frequency permits the selection of this speed to suit any desired operational condition.

Experiments on the clutter returns from terrain associated with normal airfield siting shows that phase and amplitude fluctuation is considerably less than for a microwave system. This circumstance has assisted the visibility of aircraft in the clutter and a Subclutter Visibility of 30 db has been estimated for 560 p.p.s. and 10 r.p.m. proving that the valuable facility of plotting through intense regions of clutter has been achieved.

**S.232 Radar Parameters**

*Aerial*

Frequency	...	...	...	...	...	...	= 585—610 Mc/s.
Horizontal aperture	...	...	...	...	...	...	= 18 wavelengths.
Horizontal beamwidth—3 db one way	...	...	...	...	...	...	= 4 degrees.
Sidelobes	...	...	...	...	...	...	= -24 db.
Vertical aperture	...	...	...	...	...	...	= 6 wavelengths.
Vertical beamwidth—3 db one way	...	...	...	...	...	...	= 12 degrees.
Mean height, depending on installation...	...	...	...	...	...	...	= 15 feet (4.6 m.)

*Transmitter*

Power output	...	...	...	...	...	...	= 60 KW.
Pulse length	...	...	...	...	...	...	= 4 or 2 $\mu$ s.
Repetition rate	...	...	...	...	...	...	= 500—800 p.p.s.

*Marconi Coherent M.T.I. Radar on 50 cms.*

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*Receiver*

Overall bandwidth—3 db	...	...	...	...	= 350 kc/s.
Noise factor	...	...	...	...	= 9 db.
I.A.G.C. threshold above noise	...	...	...	...	= 20—26 db.

*Cancellation*

Residual at 560 p.p.s. Aerial rotation 10 r.p.m.	...	...	...	...	= - 32 db.
Residual at 800 p.p.s. Aerial rotation 6 r.p.m.	...	...	...	...	= - 39 db.
Instrumental limitation of cancellation...	...	...	...	...	= + 40 db.
Sub-clutter visibility at 560 p.p.s. 10 r.p.m.	...	...	...	...	= - 30 db.

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### BOOK REVIEW

*Television Receiver Servicing*, Volume II, Receiver and Power Supply Circuits, by E. A. W. Spreadbury, M.Brit., I.R.E. Published by Iliffe & Sons, Ltd., London. Price 21/-

Volume I of this two-part book, reviewed in a recent issue of this Journal (No. 116, Vol. XVIII), was concerned, in the main, with the Cathode Ray Tube and the Time Bases, leaving the Vision and Sound Receivers and the Power Supplies to be dealt with in Volume II.

The emphasis in Volume I, which dealt with that part of the receiver responsible, we are told, for the largest proportion of faults, was naturally on fault finding; in Volume II, however, it shifts to explanations of the working of individual circuits and the methods used to align them. In addition to chapters on the Video Stage, Tuning Circuits, the Sound Channel and Power Supplies, there is a comprehensive discussion of Aerials and Feeders which deals with everything from the meaning of characteristic impedance to how to make a Balun, from the design of attenuators to the best way to anchor a feeder to a tiled roof.

One chapter is devoted to Multi-channel Tuners for Bands I, II and III. The subject is a complex one and has been capably handled here, but one gains the impression that the author was handicapped by lack of operating experience in Band III, since the book was published before the opening of the first station in that Band. Some expansion of the subject will no doubt be undertaken in a later edition.

The high standard of excellence set by Volume I has been fully maintained in this later book and the two together form a well integrated and entirely readable whole. To the reader, whatever his interest in television, they bring a clear exposition of how the circuits chosen for a cross-section of the Industry's receivers work, while the essential similarities and differences between makers' products are clearly brought out.

As was noted in the case of Volume I, it is evident that the author is a practical man, and the information that he presents is obviously first-hand and completely reliable. A valuable feature of the book is the large number of hints and tips which, one feels, are the outcome of the author's personal experience in the field of television servicing.

# DESIGN OF A SIMPLE LINEAR FREQUENCY DISCRIMINATOR

By MRS. K. G. FAN COURT, B.A. and J. K. SKWIRZYNSKI, B.Sc., A.R.C.S.

*A design procedure is described for the construction of a simple frequency discriminator, consisting essentially of two tuned circuits in such a way as to minimize the harmonic content of the resulting demodulated signal. In particular the circuit parameters are chosen to ensure minimum 2nd and 3rd harmonics. Special care is taken to present the design procedure in a form readily used in practice; all necessary information is embodied in graphs which are very simple to consult.*

*The method presented is suitable for a large field of applications, as the resulting system is found to be relatively insensitive to frequency instability, provided it is kept within reasonable limits (i.e., variations not greater than the maximum frequency deviation of the input FM signal from the carrier frequency); these effects of detuning are shown on graphs.*

*It can be shown that an ordinary tuned circuit placed in the grid leads of the discriminator valves would completely spoil the performance of the system by introducing a comparatively high 2nd and 3rd harmonic component. The use of a transitionally coupled grid network is, therefore, advocated, whose amplitude response is sufficiently flat.*

*Finally, the effect of the 2nd harmonic of the carrier frequency is studied, and the use of the design procedure is exemplified for a particular case.*

## Introduction and Statement of the Problem

WITH the rapid development of multichannel communication by means of frequency modulated signals it has become necessary to investigate the possibility of using the simplest techniques and systems. At the same time all the available degrees of freedom should be used.

The problem considered here is the construction of a simple frequency discriminator with the best possible linearity, i.e., with the least possible harmonic content. A typical circuit of such a discriminator is shown in Fig. 1. Whilst it is possible to arrange such a circuit in several different ways, the basic principles of operation remain the same and are explained in Fig. 2.

The two anode circuits are tuned respectively to  $\alpha_1 f_0$  and  $\alpha_2 f_0$  where

$$\alpha_1 > 1 > \alpha_2$$

and where  $\omega_0 = 2\pi f_0$  is the centre angular frequency. This centre frequency (carrier) is modulated over the available band limited by

$$\omega_0 \pm \Delta$$

where  $\Delta$  is the maximum angular frequency deviation. Thus, the signal applied to the grids of both valves can be written as

$$V = V_0 \sin \left( \omega_0 t - \frac{\Delta}{p} \cos pt \right) \quad (1)$$

where  $p$  is the angular modulation frequency.

*Design of a Simple Linear Frequency Discriminator*

The signals are rectified and applied out of phase to the combining network. The resultant frequency-amplitude conversion curve can be constructed by subtracting the selectivity curves of both networks. Such a response curve is shown in Fig. 2, where the relative heights of the two selectivity curves are expressed in terms of the ratio  $h$ . This ratio can be adjusted by means of the potential divider  $r$  (Fig. 1) which regulates the grid bias of each valve. Furthermore, each of the tuned circuits is characterized by its  $Q$  value, where

$$Q = \frac{R}{\omega_0 L} \tag{2}$$

and  $R$  (i.e.  $R_1$  or  $R_2$  in Fig. 1) are the damping resistances of the two circuits respectively (including the valve internal resistances). Similarly  $C_1$  and  $C_2$  (Fig. 1) include the pertinent valve interelectrode capacities.

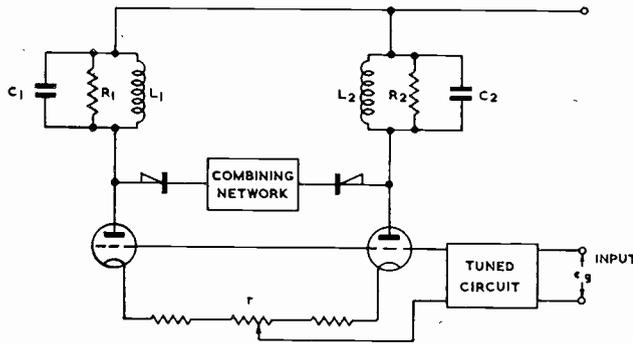


FIG. 1

*Schematic Diagram of the Simple Frequency Discriminator.*

In Fig. 2 it is seen that, provided the separation between  $\alpha_1 \omega_0$  and  $\alpha_2 \omega_0$  is suitably chosen, the shape of the resultant response curve will be almost linear near the carrier frequency. It has been the practice until now to take  $Q_1 = Q_2 = Q$  and also to tune both circuits at equal distances from the centre frequency:

$$\alpha_1 - 1 = 1 - \alpha_2$$

These  $\alpha$ 's were made to satisfy the condition

$$\sqrt{\frac{3}{2}} \frac{1}{Q} \leq \alpha_1 - 1 \leq \frac{3}{2} \frac{1}{Q}$$

Also it has been assumed that, provided  $h = 1$ , the overall response at  $\omega = \omega_0$  will vanish producing no audio-response at the carrier frequency. The last condition is necessary for satisfactory noise rejection in the discriminator stage. We shall see

that the above conditions are in fact generally incompatible, unless the value of the  $Q$  of the tuned circuits is so high as to warrant the use of arithmetical approximation of the network response function.

The selectivity curve of a simple parallel network, as shown in Fig. 1, is given but for a multiplying constant, by

$$|Z| = Z = (1 + X^2)^{-\frac{1}{2}} \quad (3)$$

where

$$X = Q \left( \frac{\omega}{\omega_0} - \frac{\omega_0}{\omega} \right) \quad (4)$$

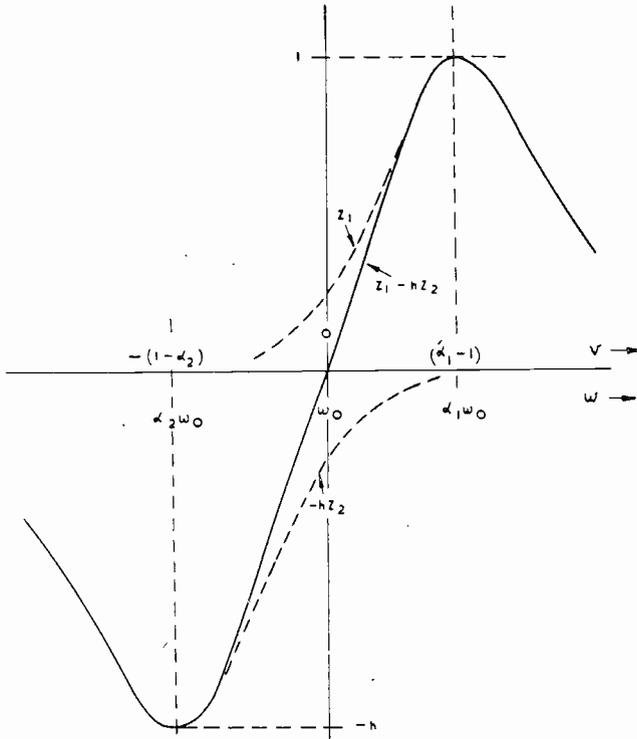


FIG. 2

*The Frequency Amplitude Response Conversion Curve of a Discriminator.*

The discriminator characteristic  $F$  is obtained by subtracting the selectivity curves of both tuned networks:

$$\begin{aligned} F &= Z_1 - hZ_2 \\ &= (1 + X_1^2)^{-\frac{1}{2}} - h(1 + X_2^2)^{-\frac{1}{2}} \end{aligned} \quad (5)$$

where

$$X_{1,2} = Q_{1,2} \left( \frac{\omega}{\omega_0 \alpha_{1,2}} - \frac{\omega_0 \alpha_{1,2}}{\omega} \right) \quad (6)$$

It is useful to introduce at this stage the fractional angular frequency deviation  $v$  from the centre angular frequency  $\omega_0$

where, from (1)  $\omega = \omega_0 (1 + v)$  (7)

$v = \frac{\Delta}{\omega_0} \sin pt$  (8)

Thus

$X_{1,2} = Q_{1,2} \left( \frac{1+v}{\alpha_{1,2}} - \frac{\alpha_{1,2}}{1+v} \right)$  (9)

The response will be linear around the carrier frequency if

$\left. \frac{\partial^2 F}{\partial v^2} \right|_{v=0} = 0$  (10)

This condition can be realized in several ways by relating the five available parameters

$h, \alpha_1, \alpha_2, Q_1, Q_2$

However, (10) will not be true if we assume equal  $Q$ -values and equal detunings for both networks, even when  $h = 1$ .

Suppose we expand (5):

$F = F_0 + F_I v + F_{II} v^2 + F_{III} v^3 + \dots$  (11)

then, using (8) and neglecting the powers of  $v$  higher than the fourth, we obtain

$F = H_0 + H_I \sin pt - H_{II} \cos 2 pt$   
 $- H_{III} \sin 3 pt + H_{IV} \cos 4 pt$   
 $+ H_V \sin 5 pt - \dots$  (12)

where

$$\left. \begin{aligned} H_0 &= F_0 + \frac{1}{2} \left( \frac{\Delta}{\omega_0} \right)^2 F_{II} + \dots \\ H_I &= \left( \frac{\Delta}{\omega_0} \right) F_I + \frac{3}{4} \left( \frac{\Delta}{\omega_0} \right)^3 F_{III} + \dots \\ H_{II} &= \frac{1}{2} \left( \frac{\Delta}{\omega_0} \right)^2 F_{II} + \frac{1}{2} \left( \frac{\Delta}{\omega_0} \right)^4 F_{IV} + \dots \\ H_{III} &= \frac{1}{4} \left( \frac{\Delta}{\omega_0} \right)^3 F_{III} + \frac{5}{16} \left( \frac{\Delta}{\omega_0} \right)^5 F_V + \dots \\ H_{IV} &= \frac{1}{8} \left( \frac{\Delta}{\omega_0} \right)^4 F_{IV} + \dots \end{aligned} \right\} \quad (13)$$

Inspection of the two equations (12) and (13) enables us to determine the conditions which the parameters must satisfy to produce a reliable linear discriminator.

*Condition A.* There is no signal when  $v = 0$ , i.e.  $F_0 = 0$ . It is shown below that  $F_2$  may be made to vanish. Then the steady state (i.e., at the carrier frequency) response becomes

$H_0 \sim \left( \frac{\Delta}{\omega_0} \right)^4$  (A)

where the sign  $\sim$  is taken to mean "is of the order of".

From (10), the condition of linearity is that  $F_{II} = 0$ ; there are still some degrees

of freedom available and these can be used to make  $F_{III} = 0$ .

This can be done in two ways. From (5) we see that each coefficient in (11) consists of two parts, one due to each network; thus

$$\left. \begin{aligned} F_I &= F_{I,1} - h F_{I,2} \\ F_{II} &= F_{II,1} - h F_{II,2} \text{ etc.} \end{aligned} \right\} \quad (14)$$

where

$$\left. \begin{aligned} F_{I,1} &= F_{I,1}(\alpha_1, Q_1) \\ F_{I,2} &= F_{I,2}(\alpha_2, Q_2) \text{ etc.} \end{aligned} \right\} \quad (15)$$

Thus, we can make  $F_{II,1} = 0$  and  $F_{II,2} = 0$  separately, thereby relating each  $Q$  and  $\alpha$  so that both networks have the points of inflection of their response curves coinciding with the centre frequency; then we can try to equate

$$F_{III} = F_{III,1} - h F_{III,2} = 0,$$

thereby relating both detunings  $\alpha_1$  and  $\alpha_2$  so that the cubic coefficient also vanishes. In fact this last condition cannot be fulfilled exactly.

Alternatively, we can first set

$$F_{III,1} = F_{III,2} = 0$$

separately, whereby each network is tuned so that its cubic coefficient vanishes at the carrier frequency. Then we can try to equate  $F_{II} = F_{II,1} - h F_{II,2} = 0$ , thereby cancelling the curvatures of both networks. The second method is preferable as it gives a much more stable system, i.e., less sensitive to accidental detunings. We thus propose:

*Condition B.* The third harmonic is minimized first by making

$$F_{III,1} = F_{III,2} = 0 \quad (B)$$

Then the 3rd harmonic ratio becomes

$$\frac{H_{III}}{H_I} \sim \left( \frac{\Delta}{\omega_0} \right)^4$$

This condition relates each  $Q$  with the corresponding  $\alpha$ .

*Condition C.* Subject to condition (B)

$$F_{II} = F_{II,1} - h F_{II,2} = 0 \quad (C)$$

Then the 2nd harmonic ratio becomes

$$\frac{H_{II}}{H_I} \sim \left( \frac{\Delta}{\omega_0} \right)^3$$

This condition relates  $\alpha_1$  and  $\alpha_2$ .

The three conditions A, B and C will constrain  $1 + 2 + 1 = 4$  degrees of freedom from the 5 available. This leaves one parameter (in our case  $\alpha_1$ ) which can be varied at will according to the bandwidth requirements.

### Expansion of the Discriminator Characteristic Function

Consider the general bandpass parameter expression given in (9).

$$\begin{aligned} X &= Q \left( \frac{1+v}{\alpha} - \frac{\alpha}{1+v} \right) \\ &= \frac{Q}{\alpha} \left[ (1+v)^2 - \alpha^2 \right] (1+v)^{-1} \end{aligned} \quad (16)$$

Thus

$$X^2 = \frac{Q^2}{\alpha^2} \sum_{s=0}^{\infty} \xi_s v^s \quad (17)$$

where

$$\left. \begin{aligned} \xi_0 &= (1 - \alpha^2)^2 \\ \xi_I &= 2(1 - \alpha^4) \\ \xi_{II} &= 1 + 3\alpha^4 \\ \xi_{III} &= -4\alpha^4 \\ \xi_{IV} &= 5\alpha^4 \\ \xi_V &= -6\alpha^4 \\ &\text{etc.} \end{aligned} \right\} \quad (18)$$

The individual selectivity function is given in (3):

$$\begin{aligned} Z &= \frac{1}{\sqrt{1+X^2}} \\ &= \left[ 1 + \frac{Q^2}{\alpha^2} \sum_{s=0}^{\infty} \xi_s v^s \right]^{-\frac{1}{2}} \end{aligned} \quad (19)$$

$$= \left[ 1 + \frac{Q^2}{\alpha^2} (1 - \alpha^2)^2 \right]^{-\frac{1}{2}} \left[ 1 + \frac{Q^2}{\alpha^2} \sum_{s=1}^{\infty} \frac{\xi_s v^s}{1 + \frac{Q^2}{\alpha^2} (1 - \alpha^2)^2} \right]^{-\frac{1}{2}} \quad (20)$$

Let

$$1 + \frac{Q^2}{\alpha^2} (1 - \alpha^2)^2 = \beta^2 \quad (21)$$

$$\frac{\alpha^2}{Q^2} + (1 - \alpha^2)^2 = \frac{\alpha^2}{Q^2} \beta^2 = \epsilon^2 \quad (22)$$

Then (20) becomes

$$Z = \frac{1}{\beta} \left( 1 + \sum_{s=1}^{\infty} \frac{\xi_s v^s}{\epsilon^2} \right)^{-\frac{1}{2}} \quad (23)$$

$$\begin{aligned} &= \frac{1}{\beta} \left\{ 1 - \frac{\xi_I}{2\epsilon^2} v + \left[ \frac{3\xi_I^2}{8\epsilon^4} - \frac{\xi_{II}}{2\epsilon^2} \right] v^2 \right. \\ &\quad \left. - \left[ \frac{5\xi_I^3}{16\epsilon^6} - \frac{3\xi_I \xi_{II}}{4\epsilon^4} + \frac{\xi_{III}}{2\epsilon^2} \right] v^3 \right. \end{aligned} \quad (24)$$

$$\begin{aligned}
 & + \left[ \frac{35 \xi_1^4}{128 \epsilon^6} - \frac{15 \xi_1^2 \xi_{II}}{16 \epsilon^6} + \frac{3 \xi_{II}^2 + 6 \xi_1 \xi_{III}}{8 \epsilon^4} - \frac{\xi_{IV}}{2 \epsilon^2} \right] v^4 \\
 & - \left[ \frac{63 \xi_1^5}{256 \epsilon^{10}} - \frac{35 \xi_1^3 \xi_{II}}{32 \epsilon^8} + \frac{15 \xi_1 (\xi_I \xi_{III} + \xi_{II}^2)}{16 \epsilon^6} \right. \\
 & \left. - \frac{3(\xi_I \xi_{IV} + \xi_{II} \xi_{III})}{4 \epsilon^4} + \frac{\xi_V}{2 \epsilon^2} \right] v^5 + \dots \}
 \end{aligned}$$

This expansion is true for both networks. In subsequent work each is distinguished by a subscript 1 or 2. Expressions having no subscripts apply equally to both networks.

Equation (24) gives the quantities  $F_{s,1}$  and  $F_{s,2}$  ( $s = 0, I, II, \text{etc.}$ ), i.e., the coefficients of  $v^s$  for the two networks. Using (14) and (15) these define the coefficients of  $v^s$  in the discriminator characteristic function

$$F_s = F_{s,1} - h F_{s,2} \tag{25}$$

**The Conditions on the Initial Value and the Slope of the Discriminator Characteristic**

As stated on page 64, the condition for no output at  $v = 0$  is

$$F_0 = 0, \text{ or } F_{0,1} - h F_{0,2} = 0 \tag{26}$$

Hence, using (24)

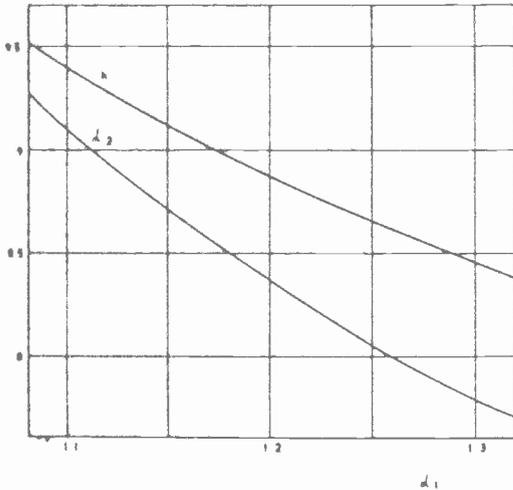


FIG. 3

The Relation between the two Detuning Ratios, and the Gain Ratio,  $h$ , as functions of  $\alpha_1$ .

$$\frac{1}{\beta_1} = \frac{h}{\beta_2} \tag{27}$$

Therefore,

$$h^2 = \frac{1 + \frac{Q_2^2}{\alpha_2^2} (1 - \alpha_2^2)^2}{1 + \frac{Q_1^2}{\alpha_1^2} (1 - \alpha_1^2)^2} \tag{28}$$

The gain ratio  $h$ , is thus defined in terms of four other parameters, whose values are found on pp. 68 and 69. It is plotted in Fig. 3.

In practice, after  $\alpha_1, \alpha_2, Q_1$  and  $Q_2$  are determined  $h$  can be simply adjusted by means of the potentiometer  $r$  (see Fig. 1).

The fundamental component of the output signal is (from (13)) nearly proportional to  $F_1$ , i.e., to

$$F_1 = F_{1,1} - h F_{1,2} = \frac{1}{2\beta_1} \left[ \frac{\xi_{I,2}}{\epsilon_2^2} - \frac{\xi_{I,1}}{\epsilon_1^2} \right] \tag{29}$$

after use has been made of (24) and (27). The function  $F_1$  is plotted in decibels in

Fig. 4, after inserting the values of  $\alpha_2$ ,  $Q_1$ ,  $Q_2$  found on pp. 68 and 69. To estimate the gain of the discriminator this function is added to  $20 \log_{10} \left( \frac{\Delta}{\omega_0} \right)$  obtained from

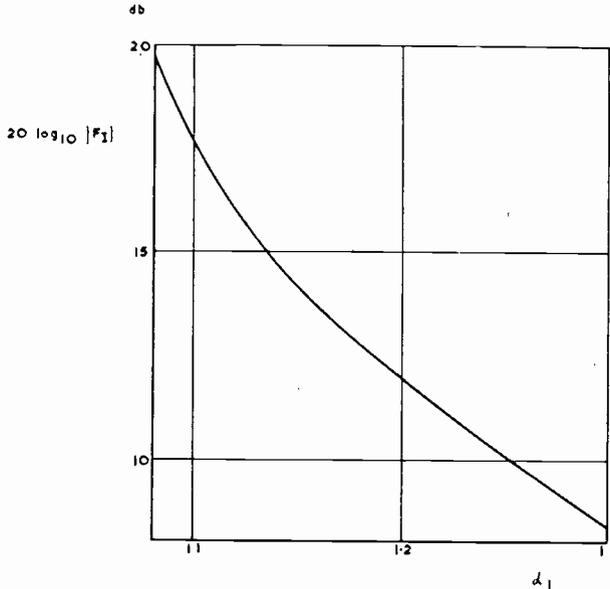


FIG. 4  
*The Discriminator Gain Function.*

Fig. 8. This gives the gain with respect to the maximum response of the tuned circuit denoted by the suffix 1.

Otherwise, if  $e_g$  is the input voltage (see Fig. 1) and  $V_a$  the voltage across the combining network, the slope of the discriminator characteristic becomes

$$V_a = e_g g_{m1} R_1 F_1 \text{ volts per unit fractional angular deviation}$$

$$= \frac{e_g g_{m1} R_1 F_1}{f_0} \text{ volts per Mc/sec.} \quad (30)$$

where  $f_0$  = centre frequency in Mc/sec.

$g_{m1}$  = mutual conductance of valve 1.

### The Minimization of the Third Harmonic

Referring to (24) the fulfilment of condition B (page 65) requires that

$$\frac{5 \xi_I^3}{16 \epsilon^6} - \frac{3 \xi_I \xi_{II}}{4 \epsilon^4} + \frac{\xi_{III}}{2 \epsilon^2} = 0 \quad (31)$$

for each tuned network separately.

Hence

$$\frac{1}{Q^2} = \frac{x-1}{8x^3} \left[ x^3 + 17x^2 + 3x + 3 \pm (x+1) \sqrt{x^4 + 134x^2 + 9} \right] \quad (32)$$

where, for convenience,

$$x = \alpha^2 \quad (33)$$

Using this relation the positive real values of  $Q$  are plotted in the neighbourhood of  $\alpha = 1$  (Fig. 5). The branches drawn in a continuous line are obtained using the negative sign in (32). These give the values used in subsequent work. The dotted curve  $Q'$  is obtained using the positive sign. It is rejected as giving  $Q$  values too low to be of practical use.

**The Minimization of the Second Harmonic**

Condition C (page 65) requires that

$$\frac{3\xi_{I,1}^2}{8\varepsilon_1^4} - \frac{\xi_{II,1}}{2\varepsilon_1^2} - \frac{3\xi_{I,2}^2}{8\varepsilon_2^4} + \frac{\xi_{II,2}}{2\varepsilon_2^2} = 0 \tag{34}$$

On substitution this becomes

$$F(x_1) - F(x_2) = 0 \tag{35}$$

where

$$\left. \begin{aligned} F(x) &= \frac{4x^2}{G(x)} \left[ \frac{24x^2}{G(x)} - \frac{1+3x^2}{1-x^2} \right] \\ G(x) &= \sqrt{x^4 + 134x^2 + 9} - 3(3x^2 + 1) \end{aligned} \right\} \tag{36}$$

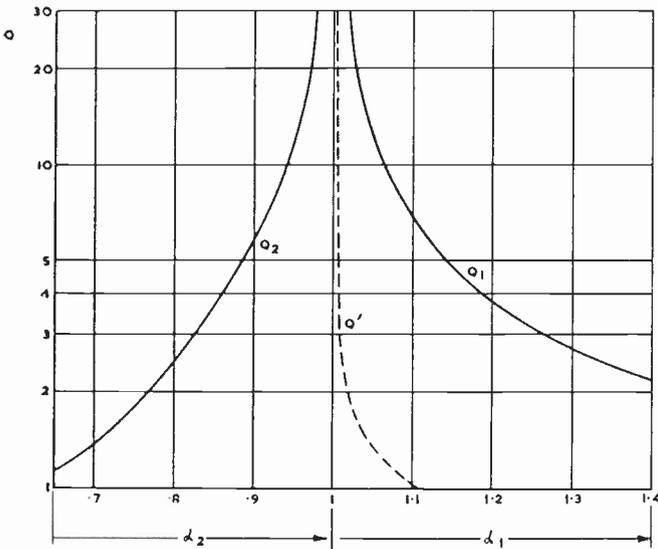


FIG. 5

*Q<sub>1</sub> and Q<sub>2</sub> as functions of the Relative Detuning.*

For any given value of  $x_1$  near, and greater than, unity (35) has a root of  $x_2$  near, and less than, unity. This relation between  $x_1$  and  $x_2$  (i.e. between  $\alpha_1^2$  and  $\alpha_2^2$ ) is of the type required. The relation between  $\alpha_1$  and  $\alpha_2$  is found by plotting  $F(x_1)$  and  $F(x_2)$ , against  $\alpha_1$  and  $\alpha_2$  respectively, in the neighbourhood of unity, and reading off pairs of values of  $\alpha_1$  and  $\alpha_2$  for which the functions are equal.

The curve showing the relationship between  $\alpha_1$  and  $\alpha_2$  is plotted in Fig. 3.

The circuit parameters have now been subjected to four conditions. Thus, if one parameter, e.g.  $\alpha_1$ , is chosen, the remaining four are obtained from the above data.

It is of interest to find the harmonic ratios under these conditions. The ratios are

$$\left. \begin{aligned} \frac{H_{II}}{H_I} &= \frac{1}{2} \frac{F_{IV}}{F_I} \left( \frac{\Delta}{\omega_0} \right)^3 \\ \frac{H_{III}}{H_I} &= \frac{5}{16} \frac{F_V}{F_I} \left( \frac{\Delta}{\omega_0} \right)^4 \\ \frac{H_{IV}}{H} &= \frac{1}{8} \frac{F_{IV}}{F_I} \left( \frac{\Delta}{\omega_0} \right)^3 \end{aligned} \right\} \tag{37}$$

The decibel values are

$$\left. \begin{aligned}
 \text{2nd harmonic} &= 20 \log_{10} |B_{II}| + 60 \log_{10} \left( \frac{\Delta}{\omega_0} \right) \\
 \text{3rd harmonic} &= 20 \log_{10} |B_{III}| + 80 \log_{10} \left( \frac{\Delta}{\omega_0} \right) \\
 \text{4th harmonic} &= 20 \log_{10} |B_{II}| + 60 \log_{10} \left( \frac{\Delta}{\omega_0} \right) - 20 \log_{10} 4
 \end{aligned} \right\} \quad (38)$$

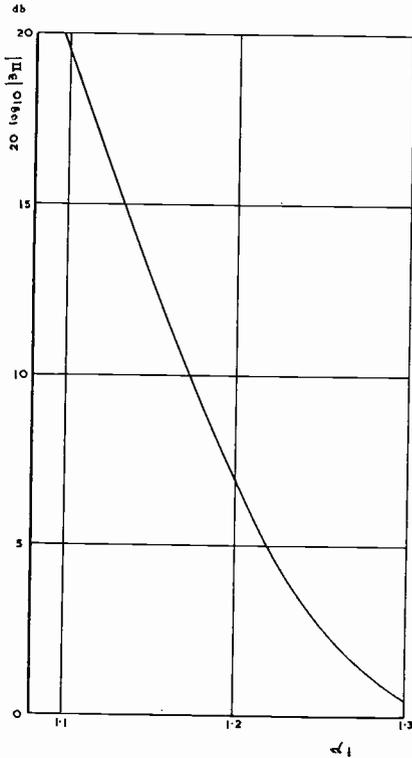


FIG. 6  
*The 2nd Harmonic Factor.*

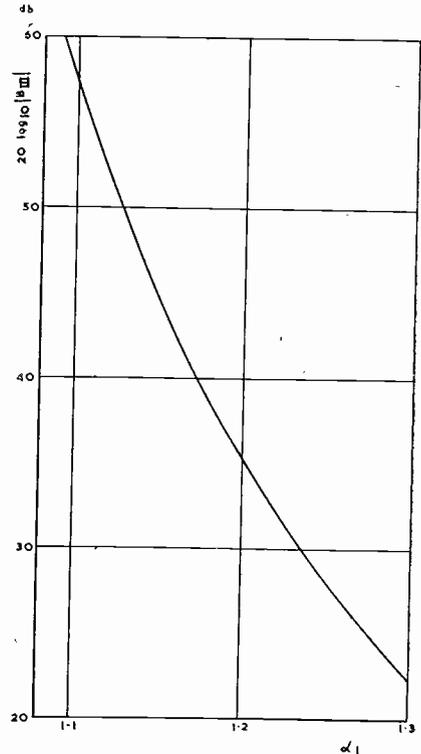


FIG. 7  
*The 3rd Harmonic Factor.*

where

$$\begin{aligned}
 B_{II} &= \frac{1}{2} \frac{F_{IV}}{F_I} \\
 B_{III} &= \frac{5}{16} \frac{F_V}{F_I}
 \end{aligned} \quad (39)$$

The values of the harmonics may be obtained from Figs. 6, 7 and 8 where the quantities involved are plotted.

### Estimation of the Effect of a Small Detuning

In practice, the spot-tuning of the individual discriminator circuits may be difficult to achieve, and what is more important, to maintain. The effect on the harmonic magnitudes of small deviations in the circuit arrangement is therefore investigated.

Suppose that the frequency deviation, instead of (8), is given by

$$v = \delta (f + \sin pt) \tag{40}$$

where  $\delta = \frac{\Delta}{\omega_0}$  (41)

and  $f$  is the relative detuning, where  $-1 \leq f \leq +1$ . Then (11) becomes

$$F(v) = F_0 + F_I \delta(f + \sin pt) + F_{II} \delta^2(f + \sin pt)^2 + \dots \tag{42}$$

Using the relations  $F_0 = F_{II} = F_{III} = 0$  the coefficients of the harmonic terms are then obtained as in (12) and (13):

$$\left. \begin{aligned} H'_I &= F \delta + F_{IV} \delta^4 (4f^3 + 3f) + \dots \\ H'_{II} &= F_{IV} \delta^4 \left( 3f^2 + \frac{1}{2} \right) + F_V \delta^5 \left( 5f^3 + \frac{5}{2}f \right) + \dots \\ H'_{III} &= F_{IV} \delta^4 f + F_V \delta^5 \left( \frac{5}{2}f^2 + \frac{5}{16} \right) + F_{VI} \delta^6 \left( 5f^3 + \frac{15}{8}f \right) + \dots \end{aligned} \right\} \tag{43}$$

The dashed  $H$ 's indicate that the harmonic coefficients are now functions of the detuning ratio  $f$ . It can be seen from (12) that

$$H'_n (f = 0) = H_n \tag{44}$$

The above equations can be written in more convenient form as

$$\left. \begin{aligned} \frac{H'_I}{H_I} &\doteq 1 + \frac{F_{IV}}{F_I} \delta^3 (4f^3 + 3f) + \dots \\ \frac{H'_{II}}{H_{II}} &\doteq 1 + 6f^2 + \frac{F_V}{F_{IV}} \delta (10f^3 + 5f) + \dots \\ \frac{H'_{III}}{H_{III}} &\doteq 1 + 8f^2 + \frac{16 F_{IV} f}{5 F_V \delta} + \frac{F_{VI}}{F_V} \delta (16f^3 + 6f) + \dots \end{aligned} \right\} \tag{45}$$

The ratios  $H'_{II}/H_{II}$  and  $H'_{III}/H_{III}$  are plotted on the db. scale, in Figs. 9 and 10 respectively.

It can be seen that, in the range considered, the second harmonic, including the detuning effect, is everywhere at least 40 db. down from the fundamental. Similarly for the third harmonic. Thus the effect on the harmonics of a small detuning is not serious.

### The Effect of the Grid Circuit Response on the Discriminator Performance

The circuit placed in the grids of both discriminator valves (see Fig. 1) may have a marked effect on the harmonic response of the system, especially if its amplitude

characteristic is not sufficiently flat over the band of frequency deviation. Any deviation from a flat amplitude response will introduce an amplitude modulated component of the grid voltage, causing corresponding variations of the slope of the discriminator characteristic, coherent with the main signal.

This amplitude modulated part of the grid voltage will generally have a preponderant 2nd harmonic of the main signal, but a detuning from the centre frequency may also introduce variations at the fundamental frequency of this signal.

Suppose that the amplitude response of the tuned circuit can be expressed in terms of the frequency deviation from the centre (carrier) frequency as follows:

$$f(v) = f_0 + f_I v + f_{II} v^2 + \dots \quad (46)$$

where  $v$  is defined as in (7) and (8). Then, as in (11) and (12)

$$f(v) = 1 + G_I \sin pt - G_{II} \cos 2pt - G_{III} \sin 3pt + \dots \quad (47)$$

where  $G_I, G_{II}, \dots$  are the amplitudes of the fundamental and harmonics. This expression represents the modulation of the grid voltage due to the variations of the amplitude response of the grid circuit. The effect of such variations on the total discriminator response may be obtained by multiplying (47) and (12) and collecting corresponding harmonic terms.

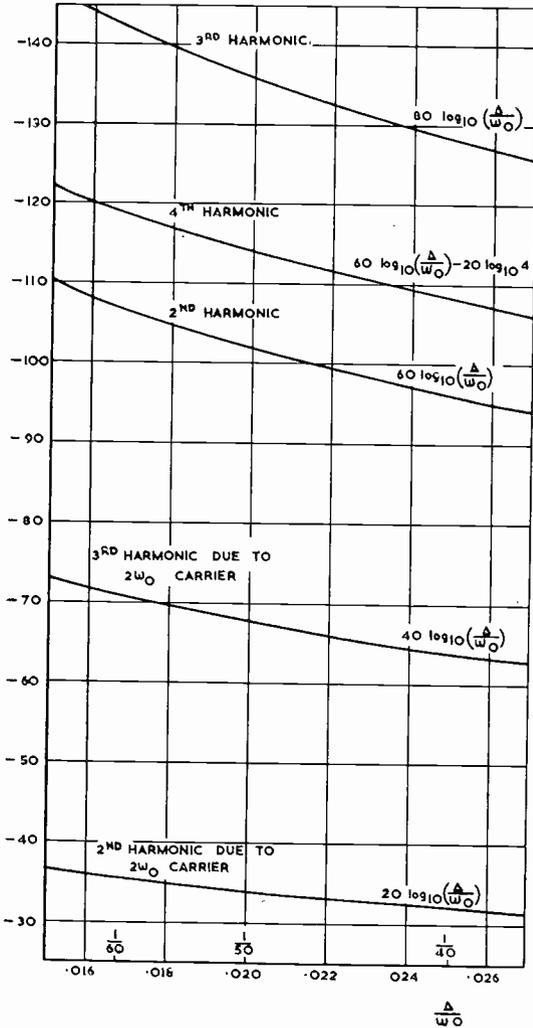


FIG. 8

Factors used to estimate the gain, in conjunction with Fig. 5, and the harmonic distortions in conjunction with Figs. 6, 7 and 11.

This is done for two types of networks:

- (a) single tuned circuit
- (b) transitionally coupled, double tuned circuit (see "The Double-Tuned Circuit with Transitional Coupling" Report 784, Radiation Laboratory, Massachusetts Institute of Technology, Cambridge, Massachusetts, U.S.A.).

Neglecting terms of higher order in  $(\Delta/\omega_0)$  than the coefficient in the original

*Design of a Simple Linear Frequency Discriminator*

discriminator response, the fundamental and harmonic coefficients of the modified response are

	Single tuned	Double tuned
Fundamental	$H_I$	$H_I$
2nd harmonic	$H_{II} + \frac{1}{2}H_I G_I + \frac{1}{2}H_I G_{III}$	$H_{II}$
3rd harmonic	$H_{III} + \frac{1}{2}H_I G_{II} + \frac{1}{2}H_I G_{IV}$	$H_{III} + \frac{1}{2}H_I G_{II} + \frac{1}{2}H_I G_{IV}$

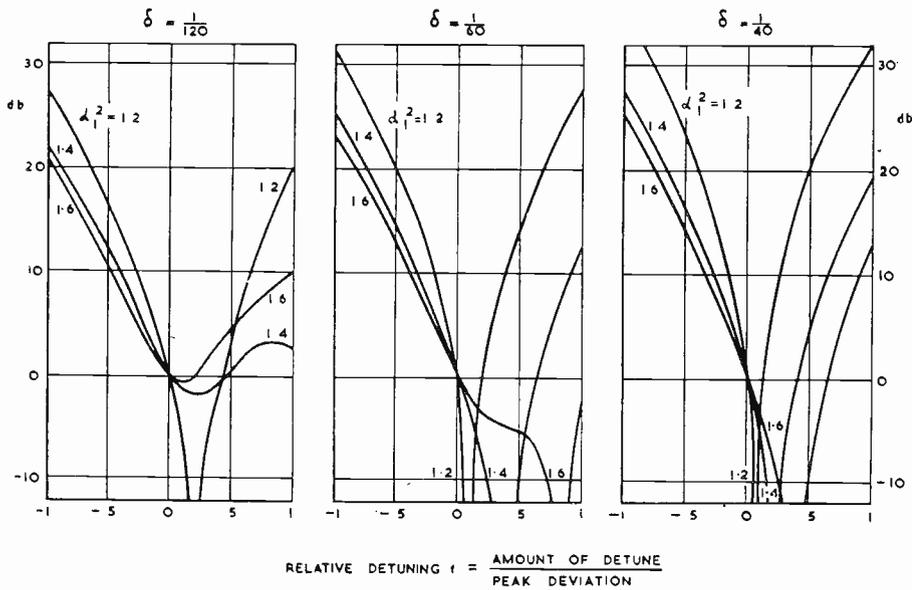


FIG. 9  
*2nd Harmonic Detuning Distortion Factors.*

where, in our case

$$H_I \sim \left(\frac{\Delta}{\omega_0}\right); H_{II} \sim \left(\frac{\Delta}{\omega_0}\right)^4; H_{III} \sim \left(\frac{\Delta}{\omega_0}\right)^5$$

Also, for the single tuned circuit

$$G_I \sim \left(\frac{\Delta}{\omega_0}\right)^2; G_{II} \sim \left(\frac{\Delta}{\omega_0}\right)^2; G_{III} \sim \left(\frac{\Delta}{\omega_0}\right)^3 \text{ etc.}$$

and for the double tuned circuit

$$G_{II} \sim \left(\frac{\Delta}{\omega_0}\right)^4 \sim G_{IV}$$

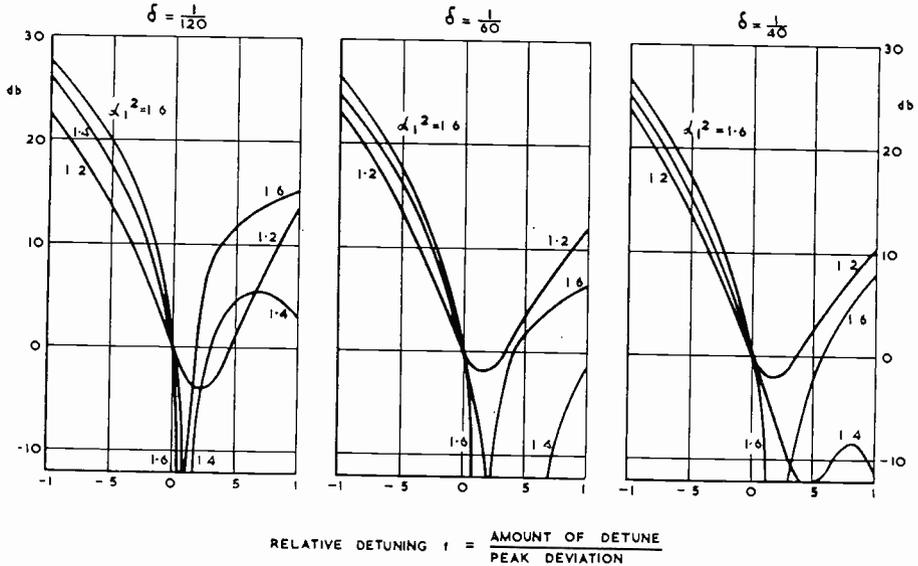


FIG. 10  
*3rd Harmonic Detuning Distortion Factors.*

It is seen, therefore, that the use of a single tuned circuit would produce 2nd and 3rd harmonics, both of the order of  $\left(\frac{\Delta}{\omega_0}\right)^3$ , which would completely destroy the linearity of the discriminator. The corresponding orders of magnitude for a double tuned circuit are  $\left(\frac{\Delta}{\omega_0}\right)^4$  and  $\left(\frac{\Delta}{\omega_0}\right)^5$  respectively.

Thus a double tuned circuit with transitional coupling placed across the input terminals of the discriminator will have no appreciable effect on the fundamental or second harmonic of the output and the change in the third harmonic is small enough to be unimportant. It is accordingly suggested that a carefully designed discriminator should always embody a double tuned grid circuit in order to preserve its linearity.

### The Second Harmonic Effect

So far the carrier frequency of the signal at the output terminals of the discriminator has been assumed to be without harmonics. In practice these may occasionally occur, and it will therefore be interesting to consider their effect on the discriminator response.

The 2nd harmonic of  $\omega_0$ , i.e.  $2\omega_0$ , will produce the fundamental as well as the harmonics of the modulation frequency which will mix with those of the main carrier. Hence, if the 2nd harmonic of the carrier is not sufficiently low, even a well designed discriminator will produce non-linear distortion. This effect is considered below.

## Design of a Simple Linear Frequency Discriminator

The working is similar to that on pp. 66 and 67 but the expansion is now made around the frequency  $2\omega_0$  (see the response curve of Fig. 2). This is effected by replacing the parameters  $\alpha$  and  $v$  by  $\alpha/2$  and  $v/2$  respectively. The  $\xi$  coefficients of (18) now become

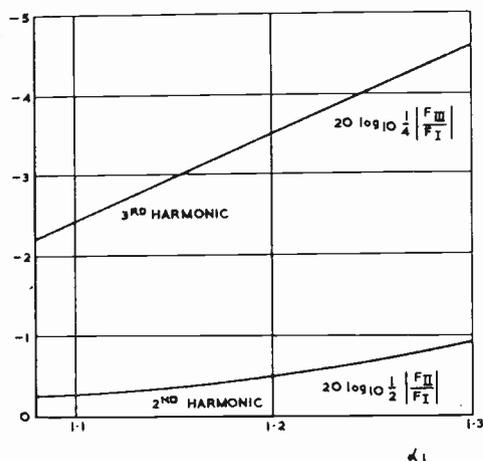


FIG. 11

*The Effect of the 2nd Harmonic of the Carrier on the 2nd and 3rd Harmonics of the output.*

$$\left. \begin{aligned} \xi_{20} &= \frac{1}{16} (4 - \alpha^2)^2 \\ \xi_{21} &= \frac{1}{8} (16 - \alpha^4) \\ \xi_{22} &= \frac{1}{16} (16 + 3\alpha^4) \\ \xi_{23} &= -\frac{1}{4} \alpha^4 \end{aligned} \right\} \quad (48)$$

From these values the 2nd and 3rd harmonic factors are calculated as before, and plotted in Fig. 11. The harmonics are given by

$$\left. \begin{aligned} \text{2nd harmonic} &= 20 \log_{10} \frac{1}{2} \left| \frac{F_{II}}{F_I} \right| + 20 \log_{10} \frac{\Delta}{\omega_0} \\ \text{3rd harmonic} &= 20 \log_{10} \frac{1}{4} \left| \frac{F_{III}}{F_I} \right| + 40 \log_{10} \frac{\Delta}{\omega_0} \end{aligned} \right\} \quad (49)$$

### An Example on the Use of the Results

Suppose we have a discriminator working round the frequency of 60 Mc/s with a peak deviation of 1 Mc/s. We tune one of the networks to 72 Mc/s.

Thus  $\alpha_1 = 1.2$

From Fig. 4.  $\alpha_2 = .837$

Fig. 3.  $Q_1 = 3.8$

Fig. 3.  $Q_2 = 3.3$

Obtain the 2nd harmonic content from Figs. 6 and 8 (using the values

$$\frac{\Delta}{\omega_0} = \frac{1}{60}; \quad \alpha_1 = 1.2)$$

$$\begin{aligned} 60 \log_{10} \left( \frac{\Delta}{\omega_0} \right) &= -106.5 \text{ db} \\ 20 \log_{10} |B_{II}| &= +7.0 \text{ db} \end{aligned}$$

Therefore

$$\text{2nd harmonic} = -99.5 \text{ db}$$

*Design of a Simple Linear Frequency Discriminator*

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Obtain the 3rd harmonic content from Figs. 7 and 8:

$$\begin{aligned} 80 \log_{10} \frac{\Delta}{\omega_0} &= -142.5 \text{ db} \\ 20 \log_{10} |B_{III}| &= + 35.5 \text{ db} \end{aligned}$$

Therefore

$$\text{3rd harmonic} = -107.0 \text{ db}$$

Obtain the 4th harmonic content from Figs. 6 and 8:

$$\begin{aligned} 60 \log_{10} \frac{\Delta}{\omega_0} - 20 \log_{10} 4 &= -119.0 \text{ db} \\ 20 \log_{10} |B_{II}| &= + 7.0 \text{ db} \end{aligned}$$

Therefore

$$\text{4th harmonic} = -112.0 \text{ db}$$

The above refers to the spot-tuning condition. To estimate the effect of detuning consult Figs. 9 and 10. Suppose the system is detuned from the carrier frequency by  $\cdot 5$  Mc/s, i.e.,  $f = \pm 0.5$ .

For the 2nd harmonic:

	$f = + 0.5$	$f = - 0.5$
Detuning effect	- 5.0 db	+ 14.5 db
Previous value	- 99.0 db	- 99.0 db
... Total	- 104.0 db	- 84.5 db

For the 3rd harmonic:

	$f = + 0.5$	$f = - 0.5$
Detuning effect	- 38.0 db	+ 16.0 db
Previous value	- 107.0 db	- 107.0 db
Total	- 145.0 db	- 91.0 db

The 2nd harmonic effect is now considered (see p. 74). It is assumed for the moment that the fundamental and 2nd harmonic in the carrier frequency are of equal magnitude. Then, consulting Figs. 8 and 11, the 2nd harmonic effect is found

to make the following contributions to the 2nd and 3rd harmonics:

$$20 \log_{10} \left( \frac{\Delta}{\omega_0} \right) = -35.6 \text{ db}$$

$$20 \log_{10} \frac{1}{2} \left| \frac{F_{II}}{F_I} \right| = \underline{\quad - .6 \text{ db}} \quad$$

Therefore

$$\text{2nd harmonic} = -36.2 \text{ db}$$

$$40 \log_{10} \left( \frac{\Delta}{\omega_0} \right) = -71.2 \text{ db}$$

$$20 \log_{10} \frac{1}{4} \left| \frac{F_{III}}{F_I} \right| = \underline{\quad - 3.5 \text{ db}} \quad$$

Therefore

$$\text{3rd harmonic} = -74.7 \text{ db}$$

The 2nd harmonic, due to the  $2\omega_0$  carrier is, from the above,  $-36.2$  db, and that due to the main carrier  $\omega_0$  (see page 75) is  $-99.5$  db. Hence, in order that the 2nd harmonic due to the  $2\omega_0$  carrier should be no greater than the 2nd harmonic due to the main carrier, the latter must be at least  $63.3$  db above the  $2\omega_0$  carrier level. For the value  $63.3$  db the resulting 2nd harmonic due to the  $2\omega_0$  carrier is attenuated to  $-99.5$  db, which is sufficiently small to be neglected.

The 3rd harmonic due to the  $2\omega_0$  carrier is still smaller, i.e.,  $-(74.7 + 63.3)$  db =  $-138.0$  db.

### Conclusion

In the preceding section we have shown the procedure to be used in applying the method given above for the design of a simple frequency discriminator. The values of  $\omega_0$  (i.e., the carrier frequency) and of  $\Delta$  (i.e., the maximum deviation frequency) are usually imposed and the only independent parameter to be chosen is  $\alpha_1$  (or for that matter  $\alpha_2$ ) giving the tuning frequency of the corresponding resonant circuit in the anode of one of the valves (Fig. 1). It is advisable that

$$\alpha_1 - 1 \geq 10 \frac{\Delta}{\omega_0}$$

to ensure that the discriminator operates on the linear part of the network characteristics. On the other hand  $\alpha_1$  should not be too large, as this would make the sensitivity of the system small (see page 68 and Fig. 4). The value  $\alpha_1 = 1.2$  chosen in the example above (page 75) seems to be the usual compromise.

### Acknowledgments

The authors wish to express their gratitude to Mr. S. Fedida for proposing the problem, and for advice and encouragement. Also to Mr. S. Clarke for help in algebraic manipulations and to Mrs. N. Huttly and Mr. J. C. Thackray for help in computation of the curves.

# DESIGN DATA FOR LADDER NETWORKS

BY E. GREEN, M.Sc.

In *Marconi Review* No. 108<sup>(1)</sup>, formulæ and curves were given for the design of ladder networks to give Butterworth or Chebyshev response in the pass band, and in No. 116<sup>(2)</sup> these were extended to cover the case of low pass networks with unequal resistance terminations. The following article gives some further useful data in a table covering the design of networks with up to nine branches and attenuation curves for networks with up to eleven branches.

FIG. 1A, B and C is reproduced from *Marconi Review* No. 108, p. 26, to show the types of network to which the data can be applied. A few symbols are repeated here for convenience. Thus in Fig. 1A:

$\omega$  = variable radian frequency.

$\omega_\beta$  = Radian frequency defining the normal limit of the pass band. For Butterworth response this may be at any level. For Chebyshev response it is fixed at valley level.

$x$  =  $\omega/\omega_\beta$ .

$V_p$  = Output voltage at the peaks of the response curve.

$V_\beta$  = Output voltage at limits of pass band.

$G_1 = \frac{1}{R_1}$  = conductance of output branch.

$C_n, G_n$  or  $R_n, L_n$  make up the input branch.

$d_1 = \frac{G_1}{C_1 \omega_\beta}$  = decrement of branch 1 at bandwidth  $\omega_\beta$  ( $R_1/L_1 \omega_\beta$  in case of series branch).

$d_n = \frac{R_n}{L_n \omega_\beta}$  or  $\frac{G_n}{C_n \omega_\beta}$  = decrement at  $\omega_\beta$  of input branch.

$D = d_n/d_1$  = ratio of input to output decrement.

$K_{12} = \frac{1}{\sqrt{L_2 C_1 \omega_\beta}}$  = coupling factor at bandwidth  $\omega_\beta$  between branches 1 and 2.

$K_{23} = \frac{1}{\sqrt{L_3 C_2 \omega_\beta}}$

$K_{r(r+1)}$  = coupling factor between  $r^{\text{th}}$  and  $(r+1)^{\text{th}}$  branch.

## Solutions for Type B Response

Equation (3) of reference 1 can be rewritten in a more convenient form for tabulation by substituting  $d_1^2 \sin^2 \theta = \gamma_n^2/(1+D)^2$  derived from equation (2) ( $\theta = \pi/2n$ ).

$$K_{r(r+1)}^2 = \frac{(\cos^2 r \theta + D^2 \sin^2 r \theta)}{\sin(2r-1)\theta \cos(2r+1)\theta (1+D)^2} \cdot \gamma_n^2 \quad (1)$$

$$= B_{r(r+1)} \gamma_n^2 \quad (1a)$$

where  $B_{r(r+1)}$  depends only on  $r$ ,  $n$ , and  $D$ , and  $\gamma_n$  depends only on the tolerance,  $V_p/V_\beta$ , in the pass band, and  $n$  the number of branches. This has been plotted in Fig. 2 for  $n = 2, 3, 4, 6$  and  $8$ . (Values for  $5$  and  $7$  can be interpolated.)

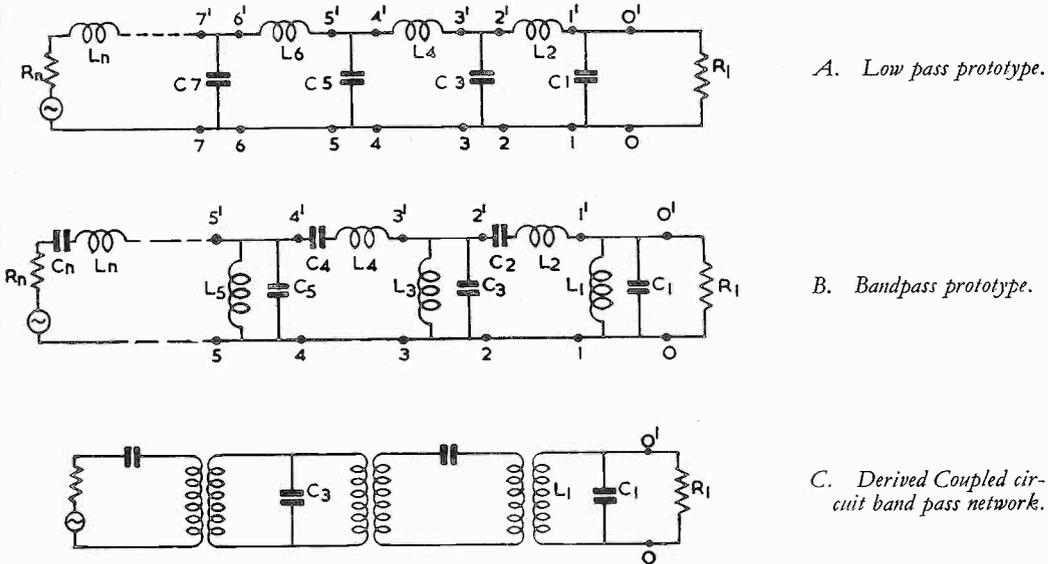


FIG. 1  
Typical Networks.

### Solutions for Type C Response

Using equation (5) of reference 1,  $d_1^2 \sin^2 \theta = S_n^2 / (1 + D^2)$  equation (6) can be rewritten.

$$K_{r(r+1)}^2 = \frac{\sin^2 r \theta \cos^2 r \theta}{\sin(2r-1)\theta \sin(2r+1)\theta} + \frac{\cos^2 r \theta + D^2 \sin^2 r \theta}{\sin(2r-1)\theta \sin(2r+1)\theta (1+D^2)^2} \cdot S_n^2 \quad (2)$$

$$= A_{r(r+1)} + B_{r(r+1)} S_n^2 \quad (2a)$$

where  $A_{r(r+1)}$  depends only on  $r$  and  $n$ ,  $B_{r(r+1)}$  is the same as for type B response, and  $S_n$  depends on the tolerance  $V_p/V_\beta$  and  $n$ , the number of branches.  $S_n$  has been plotted in Fig. 3.

### Type A Response

This is a single peaked type of response produced by undercoupling.<sup>(3)</sup> It has

$$K_{r(r+1)}^2 = -A_{r(r+1)} + B_{r(r+1)} C_n^2 \quad (3)$$

$A_{r(r+1)}$  and  $B_{r(r+1)}$  are the same as for Type C and  $C_n$  is a parameter depending on  $V_p/V_\beta$  and  $n$ , shown in Fig. 4. Also

$$d_1 (1 + D) = C_n / \sin \theta \quad (3b)$$

Table I gives values of  $A_{r(r+1)}$  and  $B_{r(r+1)}$  for values of  $n$  between 2 and 9, and  $D$  between 0 and 1.0.  $D$  can always be made less than 1.0 by numbering from the end with the larger decrement. Then, when the type of response, the value of  $D$ , and  $V_p/V_\beta$  are known, the  $K$ 's and  $d$ 's can be found, and from these, the values of the branches.

Design Data for Ladder Networks

TABLE I

n	Coupling Factor	A	B's for values of D shown										
			0	.1	.2	.3	.4	.5	.6	.7	.8	0.9	1.0
2	$K_{12}^2$	0.5	1.00	.835	.722	.645	.592	.556	.531	.516	.506	.501	.5
3	$K_{12}^2$	0.375	1.500	1.244	1.056	0.914	.806	.722	.656	.604	.562	.528	.5
	$K_{23}^2$	0.375	0.5	.426	.389	.376	.378	.389	.406	.427	.451	.475	.5
4	$K_{12}^2$	0.354	2.414	1.998	1.688	1.450	1.265	1.119	1.001	.906	.827	.762	.707
	$K_{23}^2$	0.293	.586	.489	.423	.378	.347	.325	.311	.302	.296	.294	.293
	$K_{34}^2$	0.354	.414	.362	.355	.374	.408	.452	.501	.553	.605	.656	.707
5	$K_{12}^2$	0.346	3.618	2.993	2.523	2.161	1.877	1.651	1.467	1.317	1.192	1.088	1.00
	$K_{23}^2$	0.279	.809	.672	.574	.501	.448	.407	.376	.352	.334	.320	.309
	$K_{34}^2$	0.279	.427	.360	.319	.296	.284	.280	.281	.285	.292	.300	.309
	$K_{45}^2$	0.346	.382	.345	.366	.419	.490	.572	.658	.746	.833	.918	1.00
6	$K_{12}^2$	0.342	5.098	4.217	3.551	3.036	2.631	2.306	2.043	1.826	1.646	1.495	1.366
	$K_{23}^2$	0.275	1.098	.911	.773	.669	.590	.529	.480	.442	.411	.386	.366
	$K_{34}^2$	0.268	.536	.447	.387	.346	.317	.298	.285	.276	.271	.269	.268
	$K_{45}^2$	0.275	.366	.312	.285	.275	.276	.285	.297	.313	.330	.348	.366
	$K_{56}^2$	0.342	.366	.345	.396	.488	.603	.729	.860	.991	1.120	1.245	1.366
7	$K_{12}^2$	0.340	6.859	5.671	4.773	4.078	3.528	3.088	2.729	2.434	2.188	1.980	1.806
	$K_{23}^2$	0.272	1.447	1.198	1.014	.874	.765	.680	.612	.557	.513	.476	.444
	$K_{34}^2$	0.264	.679	.565	.484	.425	.382	.350	.326	.308	.295	.285	.278
	$K_{45}^2$	0.264	.432	.363	.319	.292	.276	.267	.264	.264	.267	.272	.278
	$K_{56}^2$	0.272	.335	.289	.273	.275	.289	.309	.334	.360	.388	.416	.444
	$K_{67}^2$	0.340	.358	.353	.440	.578	.744	.923	1.107	1.290	1.469	1.642	1.806
8	$K_{12}^2$	0.339	8.875	7.335	6.172	5.270	4.556	3.983	3.516	3.130	2.809	2.537	2.307
	$K_{23}^2$	0.271	1.848	1.530	1.292	1.110	.969	.856	.766	.693	.633	.583	.540
	$K_{34}^2$	0.262	.848	.704	.599	.522	.464	.419	.384	.358	.337	.320	.307
	$K_{45}^2$	0.260	.520	.434	.375	.335	.308	.289	.276	.268	.263	.261	.260
	$K_{56}^2$	0.262	.379	.320	.287	.269	.262	.263	.267	.275	.284	.295	.307
	$K_{67}^2$	0.270	.316	.276	.271	.285	.312	.345	.382	.422	.461	.501	.540
	$K_{78}^2$	0.339	.352	.364	.491	.681	.904	1.142	1.385	1.626	1.862	2.089	2.307
9	$K_{12}^2$	0.337	11.160	9.230	7.762	6.620	5.723	5.000	4.408	3.921	3.512	3.170	2.877
	$K_{23}^2$	0.270	2.306	1.908	1.610	1.380	1.210	1.059	.944	.850	.772	.707	.653
	$K_{34}^2$	0.261	1.042	.864	.733	.635	.560	.502	.456	.419	.390	.367	.347
	$K_{45}^2$	0.258	.625	.520	.446	.393	.355	.326	.306	.291	.280	.272	.266
	$K_{56}^2$	0.258	.440	.369	.323	.293	.275	.265	.260	.258	.259	.262	.266
	$K_{67}^2$	0.261	.347	.296	.270	.261	.262	.270	.282	.297	.313	.330	.347
	$K_{78}^2$	0.270	.306	.272	.276	.304	.344	.392	.444	.497	.550	.602	.653
	$K_{89}^2$	0.337	.348	.380	.552	.800	1.089	1.395	1.705	2.013	2.311	2.600	2.877

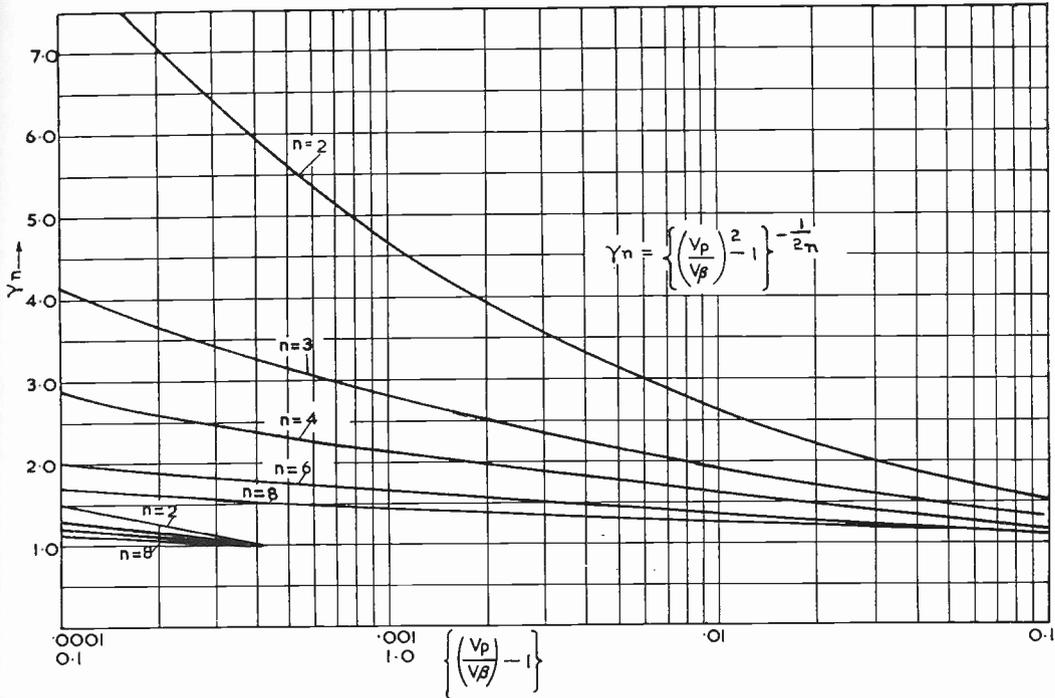


FIG. 2 Type B Response:—Parameter  $\gamma_n$ .

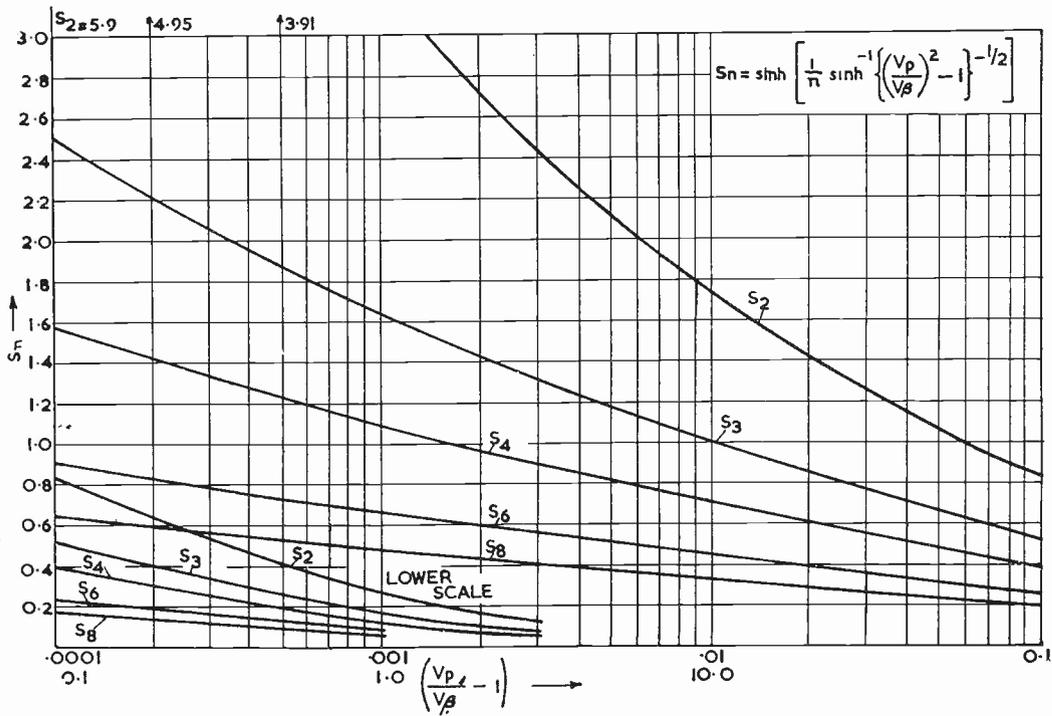


FIG. 3 Type C Response:—Parameter  $S_n$ .

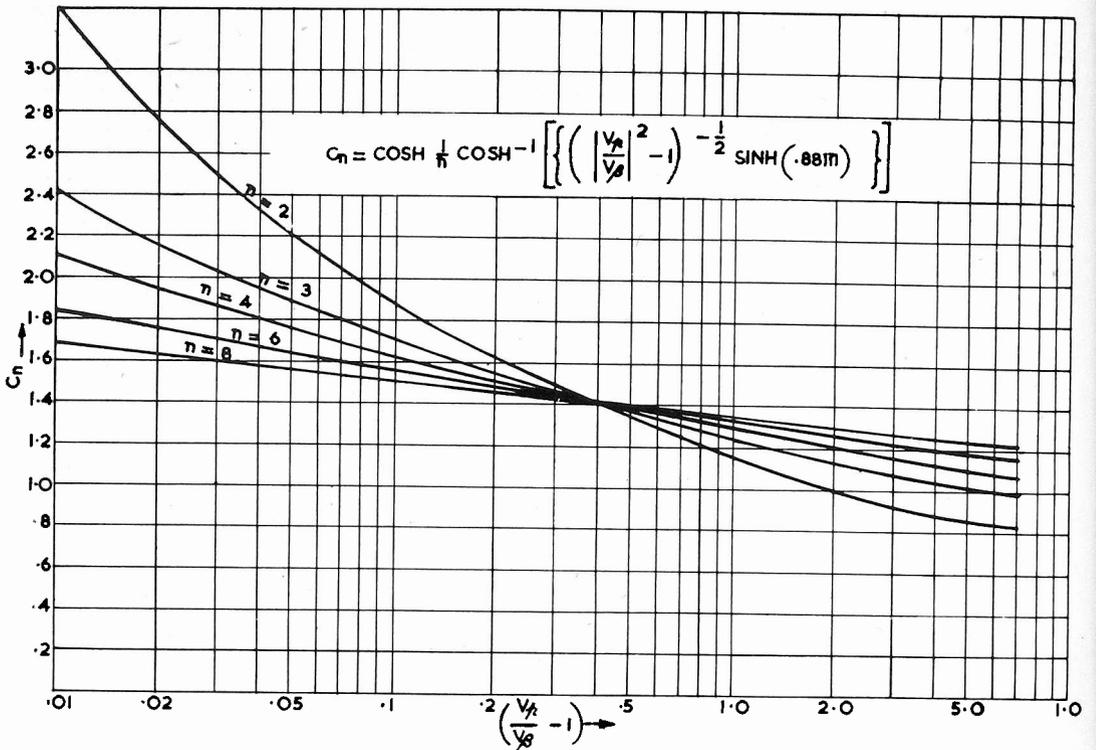


FIG. 4 Type A Response:—Parameter  $C_n$ .

**Example**

We may use the same example as in reference 2, *i.e.*, to design a network with 4 branches, with type C response in which

$$V_p/V_\beta = 1.02, R_n/R_1 = 0.6, n = 4.$$

The network is shown in Fig. 5.

Then from Fig. 3,  $S_n = 0.6$ ,  $S_n^2 = 0.36$  and from reference 2

$$D = 0.54$$

$$d_1 = \frac{S_n}{(1 + D) \sin \theta} = \frac{0.6}{1.54 \times .384} = 1.02$$

$$d_4 = D d_1 = .55$$

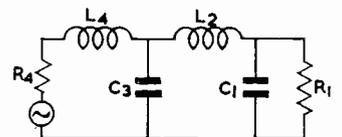


FIG. 5 Four Branch Network.

For the coupling factors we find from the Table:

$$K_{12}^2 = A_{12} + B_{12} S_n^2 = .354 + 1.06 \times .36 = .736$$

$$K_{23}^2 = A_{23} + B_{23} S_n^3 = .293 + .318 \times .36 = .407$$

$$K_{34}^2 = A_{34} + B_{34} S_n^2 = .354 + .476 \times .36 = .525$$

showing close agreement with the values in reference 2. It is of interest to note that we can find an alternative network to that of reference 2. If we reverse the

network the input will be at a shunt branch and

$$\frac{G_1}{G_4} = \frac{R_4}{R_1} = 0.6$$

This leads to the same values of  $d$ 's and  $K$ 's as before though now we have  $R_4$  associated with  $d_1$  and so on.

$$\frac{R_4}{L_4 \omega_\beta} = d_1 = 1.02 \quad \text{or} \quad L_4 \omega_\beta = .98 R_4 = .59 R_1 \quad (1.08)$$

$$\frac{1}{L_4 C_3 \omega_\beta^2} = K_{12}^2 = 0.736 \quad \text{or} \quad \frac{1}{C_3 \omega_\beta} = .736 L_4 \omega_\beta = .434 R_1 \quad (.564)$$

$$\frac{1}{L_2 C_3 \omega_\beta^2} = K_{23}^2 = 0.407 \quad \text{or} \quad L_2 \omega_\beta = \frac{1}{.407 C_3 \omega_\beta} = 1.06 R_1 \quad (1.41)$$

$$\frac{1}{C_1 L_2 \omega_\beta^2} = K_{34}^2 = 0.525 \quad \text{or} \quad \frac{1}{C_1 \omega_\beta} = .525 L_2 \omega_\beta = .558 R_1 \quad (1.03)$$

$$\frac{1}{R_1 C_1 \omega_\beta} = d_4 = 0.55 \quad \text{or} \quad R_1 = \frac{1}{.55 C_1 \omega_\beta} = 1.01 R_1$$

The branch values have been expressed in terms of  $R_1$  as in reference 2 to facilitate comparison. The values obtained there are bracketed alongside. The new version requires smaller coils and larger condensers, and might be preferable.

### Attenuation Curves of Low Pass and Bandpass Filters Giving Response Types B, C and A

A filter specification usually gives

$\omega_\beta$  = Radian frequency defining edge of pass band.

$A_p$  = Maximum allowable variation inside the pass band in db.

$\omega_a$  = Radian frequency where attenuation band begins.

$A_a$  = Minimum allowable attenuation in the attenuation band.

$k$  =  $\omega_a/\omega_\beta$ . (In the derived bandpass case,  $\omega_\beta$  and  $\omega_a$  are bandwidths.)

The curves of Figs. 6, 7 and 8 are more convenient and comprehensive than those of Figs. 24 and 25 in reference 1.

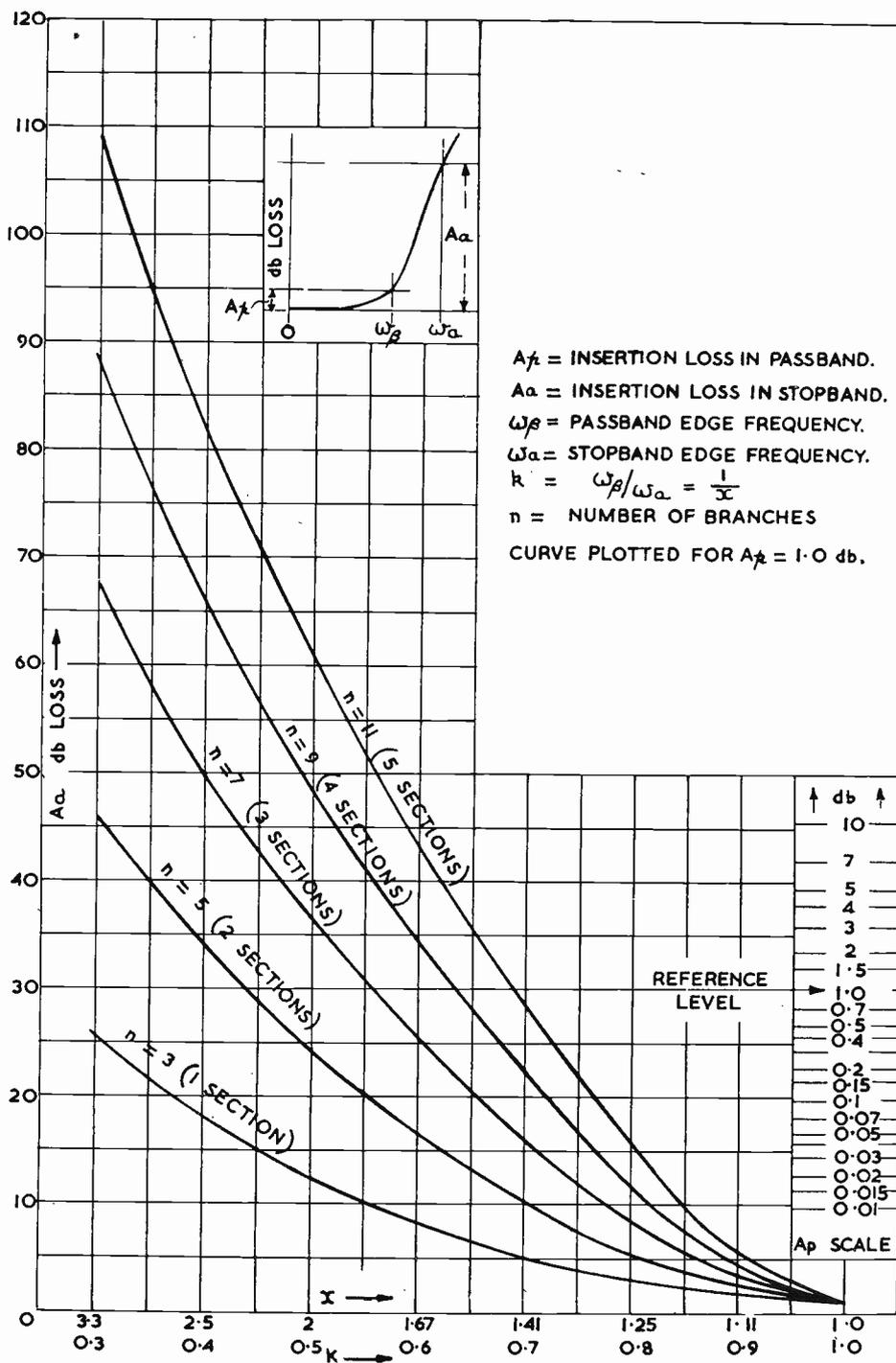
From them we can find the number of branches required when the above data are given. They show the relation between  $A_a$  and  $k$  assuming  $A_p = 1.0$  db. If  $A_p$  is not 1.0 db, the scale at the side gives the correction to  $A_a$ . Thus if  $A_p = 0.1$  db we must subtract 10 db from the value of  $A_a$  given by the curves; and for  $A_p = .3$  db we must add 6 db. The curves should not be used for values of  $A_a$  less than 10 db when the adjustment is upwards or for final values less than 10 db when the adjustment is downwards.

*Example:*— $k = 0.5$ ,  $A_p = 0.5$  db,  $A_a = 60$  db.

From Fig. 6 we see that with Type B response and 11 branches, for  $A_p = 1.0$  db,  $k = 0.5$ ,  $A_a = 60$  db, and for  $A_p = 0.5$  we must subtract 3 db from  $A_a$  leaving 57 db.

Fig. 7, for Type C response and 7 branches shows  $A_a = 68$  db on the curve, giving 65 db when  $A_p = 0.5$  db. This is the better solution.

To find the values of the branch elements, for  $A_p = 0.5$  db,  $V_p/V_\beta = 1.06$ , and Fig. 2 by interpolation gives  $S_7 = 0.26$ ,  $S_7^2 = .0676$ .



Design Data for Ladder Networks

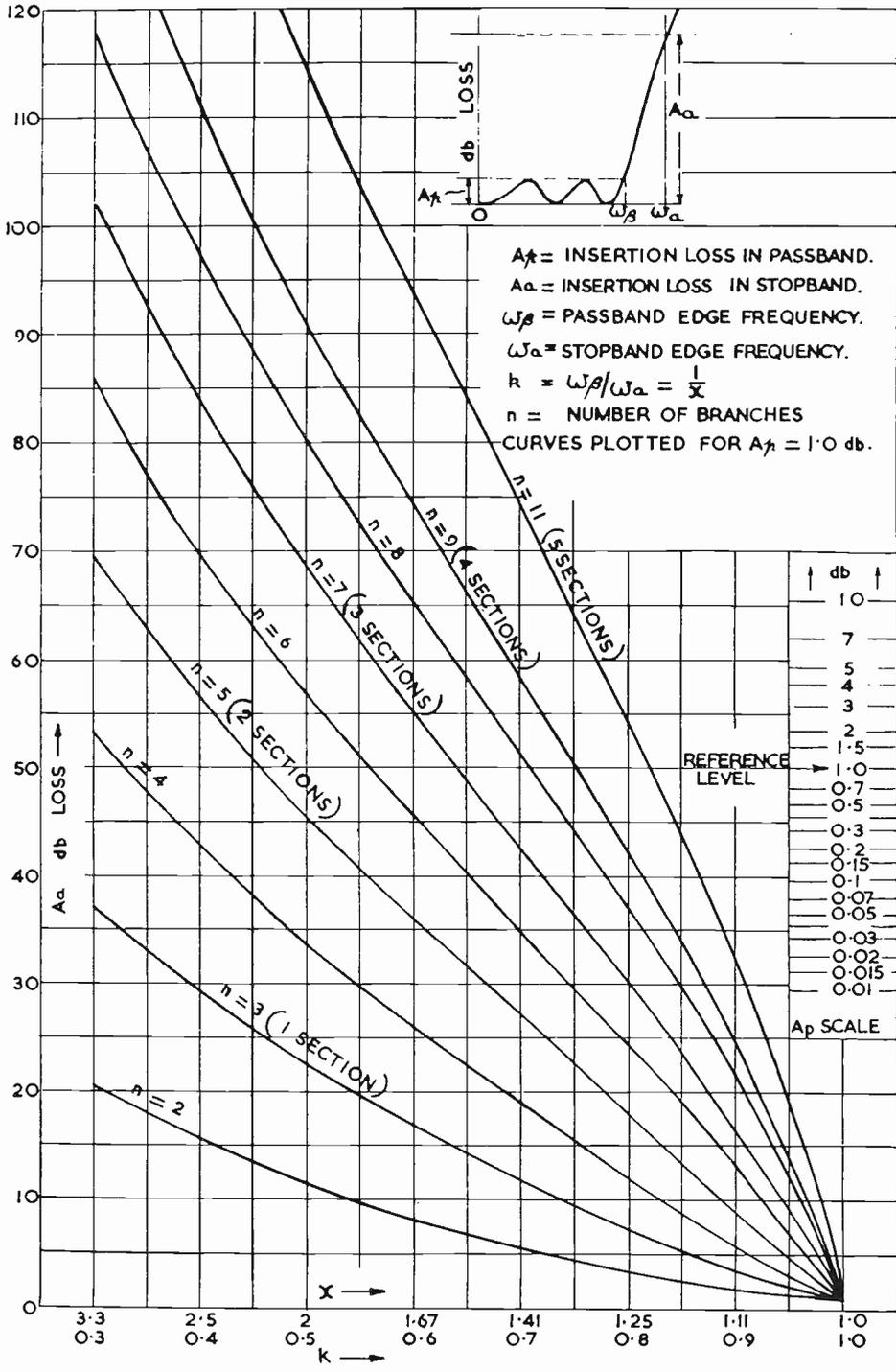


FIG. 7  
Attenuation Curves for Type C Response.

With matched terminations  $D = 1.0$ , and from equation (3)

$$2d_1 = \frac{S_7}{\sin(\pi/14)} = \frac{.26}{.239} = 1.09$$

$$d_1 = 0.545$$

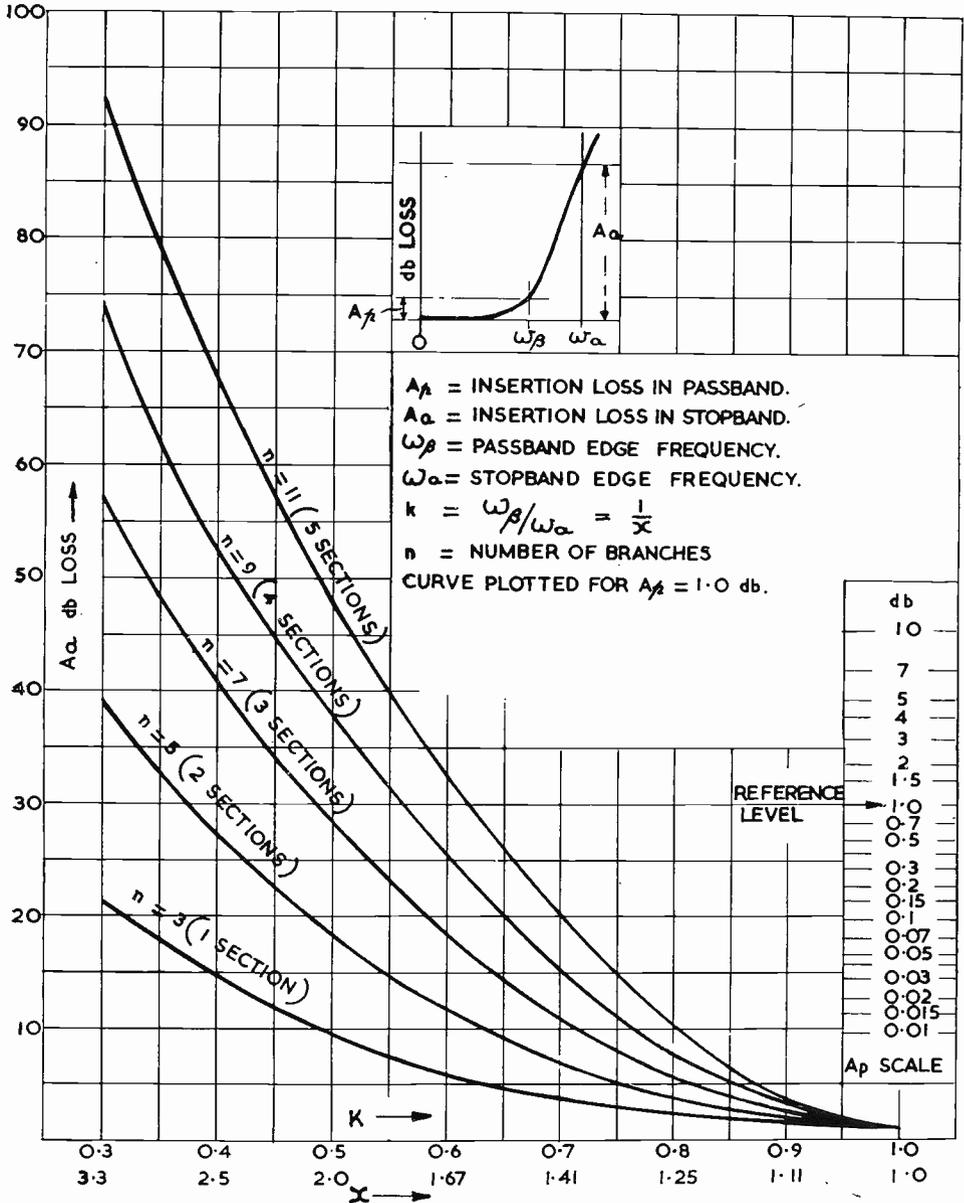


FIG. 8  
Attenuation Curves for Type A Response.

and from the table

$$\begin{aligned} K_{12}^2 &= K_{67}^2 = 0.34 + 1.806 \times 0.0676 = 0.462 \\ K_{23}^2 &= K_{56}^2 = 0.272 + 0.444 \times 0.0676 = 0.302 \\ K_{34}^2 &= K_{45}^2 = 0.264 + 0.278 \times 0.0676 = 0.283. \end{aligned}$$

So that when  $R_1$  and  $\omega_\beta$  are given the elements can be found.

### Additional Notes

The attenuation curves can also be used for mismatched filters, stagger damped or stagger tuned amplifiers having Type B or Type C overall response.

#### (a) Mismatched Filters

The  $A_a$  curves show the variation in response at the output. To find the attenuation we must add a constant db loss at all frequencies equal to that at the peaks of response. If  $|\rho_n|_{\min}$  is the reflection coefficient at this point, the fraction of available power transmitted is  $(1 - |\rho_n|_{\min}^2)$ .

$$\text{Additional db loss} = -10 \log_{10} (1 - |\rho_n|_{\min}^2) \quad (4)$$

#### (i) Low Pass Networks

For low pass networks with response types A, B and C, the last for odd values of  $n$  only,  $|\rho_n|_{\min}$  occurs at zero frequency ( $x = 0$ ) so that

$$|\rho_n|_{\min} = \frac{R_1 - R_n}{R_1 + R_n} \quad (5)$$

$$1 - |\rho_n|_{\min}^2 = \frac{4 R_1 R_n}{(R_1 + R_n)^2} \quad (6)$$

For type C response and even values of  $n$  the transmission is a minimum at  $x = 0$  so that

$$1 - |\rho_n|_{\min}^2 = \left(\frac{V_p}{V_\beta}\right)^2 \cdot \frac{4 R_1 R_n}{(R_1 + R_n)^2} \quad (7)$$

The fact that this must be less than 1, limits the possible values of  $R_n/R_1$  when  $V_p/V_\beta$  is specified<sup>2</sup>.

#### (ii) Band Pass Networks

In this case  $D$ , the ratio of decrements of the terminal branches is a more convenient parameter.

For type B response from reference 2 equation (1)

$$|\rho_n|_{\min} = \left\{ \frac{1 - D}{1 + D} \right\}^n \quad (8)$$

$$1 - |\rho_n|_{\min}^2 = 1 - \left\{ \frac{1 - D}{1 + D} \right\}^{2n} \quad (9)$$

and for type C response

$$|\rho_n|_{\min} = \frac{\sinh na'}{\sinh na} \quad (10)$$

where

$$\sinh na = \left\{ \left| \frac{V_p}{V_\beta} \right|^2 - 1 \right\}^{-1/2}$$

$$\sinh a' = \frac{1-D}{1+D} \sinh a = \frac{1-D}{1+D} \cdot S_n \quad (11)$$

$S_n$  being given in Fig. 3.

$$1 - |\rho_n|_{\min}^2 = 1 - \sinh^2 na' \left\{ \left| \frac{V_p}{V_\beta} \right|^2 - 1 \right\} \quad (12)$$

**(b) Filters with Internal Losses**

In this case also we must add at all frequencies the db loss at the peaks of response.

(c) For stagger damped, or stagger tuned amplifiers, with type B or type C overall response the  $A_a$  curves show the frequency response.

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- (1) E. Green: "Exact Amplitude Frequency Characteristics of Ladder Networks," *Marconi Review*, Vol. 16, No. 108, pp. 25-68, 1st Quarter 1953.
- (2) E. Green: "Design of Low Pass Ladder Networks with Unequal Resistance Terminations," *Marconi Review*, Vol. 17, No. 116, pp. 21-28, 1st Quarter 1955.
- (3) E. Green: "Amplitude Frequency Characteristics of Ladder Networks." *Marconi Monograph*, 1954.

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**BOOK REVIEW**

*Journal of the Institution of Telecommunication Engineers, New Delhi*

For many years there have existed scientific journals which have been valuable sources of information on radio research carried out in Indian universities. Many papers on the ionosphere and on ionospheric propagation have been published in these Indian journals, but in the main they have been written for the physicist rather than the engineer interested in the application of radio to telecommunications. In their outlook they may be compared for instance to the Proceedings of the Physical Society, London, and there has been no publication in India corresponding to Part B of the Proceedings of the Institution of Electrical Engineers.

The publication of a journal by the Institution of Telecommunication Engineers, New Delhi, is aimed at supplying this need. The first issue was published in March, 1955, and contained three technical articles and a number of items on the Institution and its affairs. As the Institution is only for telecommunication engineers and not for electrical engineers in general, it is probably not desirable to publish the *Institution News* separately in a Journal and the technical papers in Proceedings as has been done by the I.E.E.

The three papers in this first issue deal respectively with line fault location by pulse technique, radio measurements at Jabalpur during the solar eclipse of June 30th, 1954, and power density diagrams of short-wave aerial arrays. They fulfil the criterion that they should be of interest to the radio engineer and written from his point of view.

The application of radio to communications in India is rapidly expanding and a great deal of able research and development is being done there. This journal should prove a useful medium for the publication of this work and in due course it should take its place among the well-known radio journals of the world as a valuable reference and source of information.

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where

$$\sinh na = \left\{ \left| \frac{V_p}{V_\beta} \right|^2 - 1 \right\}^{-1}$$

$$\sinh a' = \frac{1-D}{1+D} \sinh a = \frac{1-D}{1+D} \cdot S_n \quad (11)$$

$S_n$  being given in Fig. 3.

$$1 - |\rho_n|_{\min}^2 = 1 - \sinh^2 na' \left\{ \left| \frac{V_p}{V_\beta} \right|^2 - 1 \right\} \quad (12)$$

**(b) Filters with Internal Losses**

In this case also we must add at all frequencies the db loss at the peaks of response.

(c) For stagger damped, or stagger tuned amplifiers, with type B or type C overall response the  $A_a$  curves show the frequency response.

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- (1) E. Green: "Exact Amplitude Frequency Characteristics of Ladder Networks," *Marconi Review*, Vol. 16, No. 108, pp. 25-68, 1st Quarter 1953.
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**BOOK REVIEW**

*Journal of the Institution of Telecommunication Engineers, New Delhi*

For many years there have existed scientific journals which have been valuable sources of information on radio research carried out in Indian universities. Many papers on the ionosphere and on ionospheric propagation have been published in these Indian journals, but in the main they have been written for the physicist rather than the engineer interested in the application of radio to telecommunications. In their outlook they may be compared for instance to the Proceedings of the Physical Society, London, and there has been no publication in India corresponding to Part B of the Proceedings of the Institution of Electrical Engineers.

The publication of a journal by the Institution of Telecommunication Engineers, New Delhi, is aimed at supplying this need. The first issue was published in March, 1955, and contained three technical articles and a number of items on the Institution and its affairs. As the Institution is only for telecommunication engineers and not for electrical engineers in general, it is probably not desirable to publish the *Institution News* separately in a Journal and the technical papers in Proceedings as has been done by the I.E.E.

The three papers in this first issue deal respectively with line fault location by pulse technique, radio measurements at Jabalpur during the solar eclipse of June 30th, 1954, and power density diagrams of short-wave aerial arrays. They fulfil the criterion that they should be of interest to the radio engineer and written from his point of view.

The application of radio to communications in India is rapidly expanding and a great deal of able research and development is being done there. This journal should prove a useful medium for the publication of this work and in due course it should take its place among the well-known radio journals of the world as a valuable reference and source of information.

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