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THE MARCONI REVIEW

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Editor : L. E. Q. WALKER, A.R.C.S.

A NOTE ON THE WORLD-WIDE DISTRIBUTION OF THE SPORADIC E IONISATION

BY L. J. PRECHNER, B.Sc.

The probable world-wide distribution of Sporadic E is described and reference is made to its variations in time and space. It is suggested that there are two limiting types of Sporadic E Layer, one in the high and one in the low latitudes, whilst in the intermediate latitudes both these types can occur. This work was initiated some years ago when the author was a member of the Marconi Company's Propagation Section. He is now with the Engineering Division of the B.B.C.

Introduction

THE results presented here are based on ionospheric data obtained for varying periods during 1944 and 1945, and it is possible that the conclusions then reached may need modification in detail in the light of more recent knowledge. However, it is considered that the broad outlines are still valid.

Records Available

The ionospheric data used below were published in Bulletins F, Numbers 1 to 18, issued by the Interservice Radio Propagation Laboratory in Washington, U.S.A. (now called the Central Radio Propagation Laboratory).* Records were available for varying periods during the years 1944–45 from thirty-one ionospheric sounding stations listed in Table I. Although a detailed analysis has been made, no thorough mapping of the world-wide distribution of Sporadic E ionisation has been attempted, as apart from North America and Australia, there were too few stations in the world. Also, observations taken at various stations may not have been always strictly comparable owing to possible differences in transmitting power and aerials; in recording techniques and in the interpretation of records.

Method of Analysis

The vertical incidence soundings of Sporadic E (Es) ionisation generally show much greater variability and irregularity than is the case for the other layers. It

* The results for Durbanville and Quassassin were obtained from other sources.

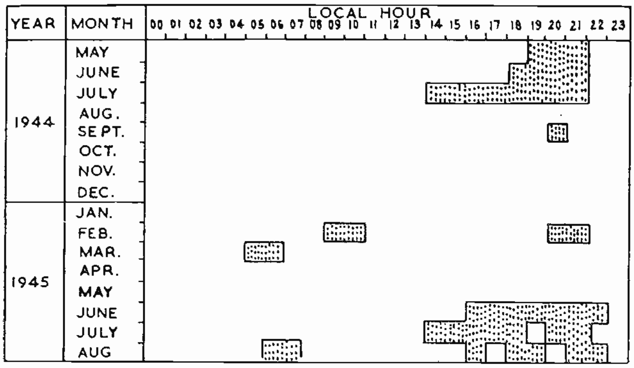
follows that in an analysis of such data the calculation of the averages of the widely scattered readings may be rather meaningless, and that the evaluation of the rate of incidence of Sporadic E may perhaps be of more practical use. This paper deals mainly with the rate of incidence of Sporadic E which has been measured here only

TABLE I

LIST OF IONOSPHERIC STATIONS

Station	Location	Approx. Co-ordinates	
		Latitude	Longitude
Baton Rouge ...	U.S.A.	30°N.	91°W.
Boston	U.S.A.	42°N.	71°W.
Brisbane	Australia	27°S.	153°E.
Canberra	Australia	35°S.	149°E.
Cape York ...	Australia	11°S.	142°E.
Christchurch ...	New Zealand	44°S.	173°E.
Christmas I. ...	Pacific Ocean (U.K. Colony)	2°N.	157°W.
Churchill	Canada	59°N.	94°W.
Clyde	Canada	70°N.	69°W.
Colombo	Ceylon	7°N.	80°W.
Durbanville ...	South Africa	34°S.	19°W.
Fairbanks ...	Alaska	65°N.	148°W.
Great Baddow ...	United Kingdom	52°N.	0°E.
Guam	Pacific Ocean (U.S.A. Colony)	14°N.	145°E.
Huancayo... ...	Peru	12°S.	75°W.
Kermadec ...	Pacific Ocean (N.Z. Colony)	29°S.	178°W.
Leyte	Philippine Is.	11°N.	125°E.
Maui	Hawaii	21°N.	156°W.
Oslo	Norway	60°N.	11°E.
Ottawa	Canada	45°N.	76°W.
Prince Rupert ...	Canada	54°N.	130°W.
Quassassin ...	Egypt	31°N.	32°E.
Reykjavik ...	Iceland	64°N.	22°W.
San Francisco ...	U.S.A.	37°N.	122°W.
San Juan ...	Puerto Rico	18°N.	66°W.
St. John's ...	Newfoundland	48°N.	53°W.
Sverdlovsk ...	U.S.S.R.	57°N.	61°E.
Tomsk	U.S.S.R.	56°N.	85°E.
Trinidad	British West Indies	11°N.	61°W.
Washington ...	U.S.A.	39°N.	77°W.
Watheroo	Australia	30°S.	116°E.

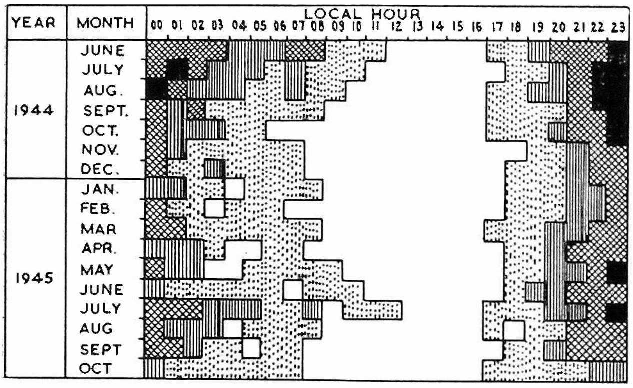
in one way, that is with the help of a symbol (c). (c) is defined as the percentage of days for which the penetration frequency of Es layer exceeded 5 Mc/s. at a given hour at a given month at a given place. This rather arbitrary lower frequency limit was chosen so as to avoid possible confusion with the normal E layer reflections, which practically never reach that high value. It is, of course, realised that reflections



CLYDE
ARCTIC ZONE N OF NORTHERN AURORAL BELT.

GEOMAGN. LAT. 82° N
GEOGRAPHIC COORDS. 70° N
69° W

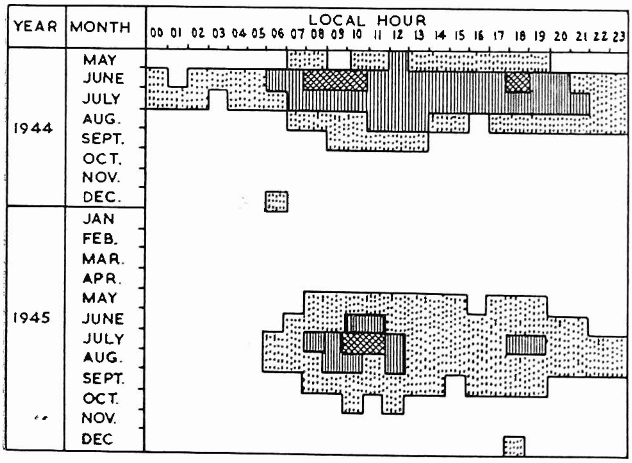
Fig. 1



CHURCHILL
NORTHERN ZONE OF MAX. AURORAL FREQUENCY

GEOMAGN. LAT. 69°
GEOGRAPHIC COORDS. 59° N
94° W

Fig. 2



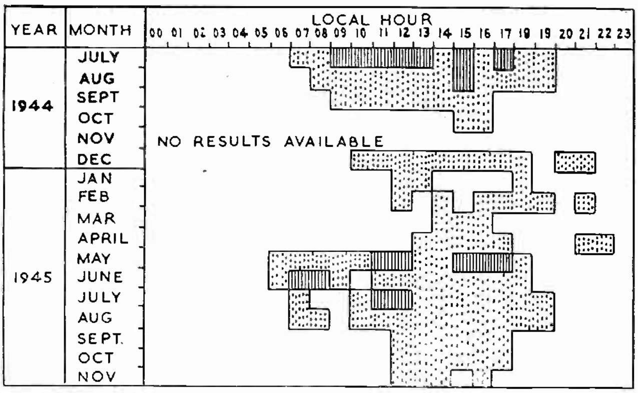
GT. BADDOW
NORTHERN MIDLATITUDE ZONE.

GEOMAGN. LAT. 54° N
GEOGRAPHIC COORDS. 52° N
0° E

Fig. 3

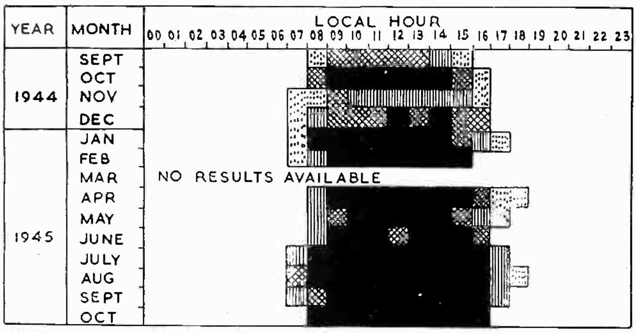
HOURLY AND SEASONAL VARIATIONS OF (C)

CODE
□ .... $C < 10\%$
▩ $50\% < C < 70\%$
▤ $10\% < C < 30\%$
■ .... $C > 70\%$
▥ $30\% < C < 50\%$



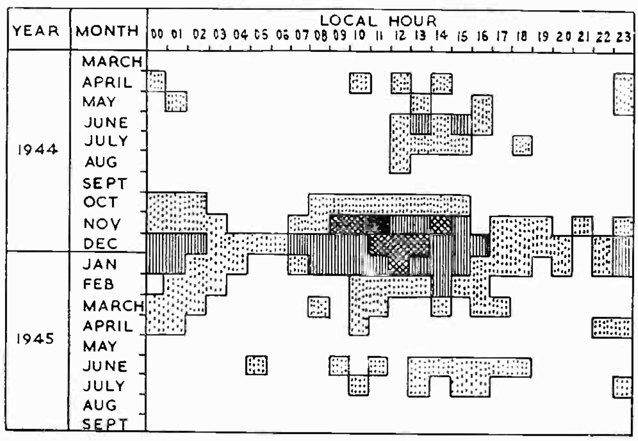
SAN JUAN
NORTHERN MIDLATITUDE ZONE (SUBTROPICAL)
GEOMAGN. LAT. 30°N
GEOGRAPHIC COORDS. 18°N 66°W

FIG. 4



HUANCAYO
EQUATORIAL ZONE
GEOMAGN. LAT. 1° S
GEOGRAPHIC COORDS. 12°S 75°W

FIG. 5



WATHEROO
SOUTHERN MIDLATITUDE ZONE (SUBTROPICAL)
GEOMAGN. LAT. 42°S
GEOGRAPHIC COORDS. 30°S 116°E

FIG. 6

HOURLY AND SEASONAL VARIATIONS OF (C)

CODE. □ .... C < 10%   50% < C < 70%
10% < C < 30%   .... C > 70%
30% < C < 50%

from Es layer can occur well below 5 Mc/s., but any references to the rate of incidence of Es in this paper refer specifically to the (*c*) values. The variations in (*c*) values are discussed below.

Latitude Variations

The world-wide distribution of Es ionisation is studied in this paper with reference to the geomagnetic and geographical co-ordinates as shown in Figs. 1 to 6. Six stations were selected going from North to South in order to indicate the diurnal and seasonal variations in the values of (*c*) with reference to geographical position. In general, it appears that in the geomagnetic mid-latitudes the prevalence of the Es ionisation was greatest in the local summer daytime, but that in the auroral and equatorial zones where the night time and daytime values respectively of (*c*) were consistently high throughout the year, the seasonal variations, if any, were not so obvious (see particularly Figs. 2 and 5). The results from Clyde indicate that even north of the auroral zone there is a local summer maximum (Fig. 1). A more detailed discussion follows.

The northernmost station, Clyde, was the only station operating north of the northern auroral zone for which Es records were then available. The values of (*c*) obtained there were very small, indicating a low degree of Es ionisation (Fig. 1). This ionisation occurred mostly in the local summer evenings (May to August) and, even then, the maximum value of (*c*) never exceeded 30%.

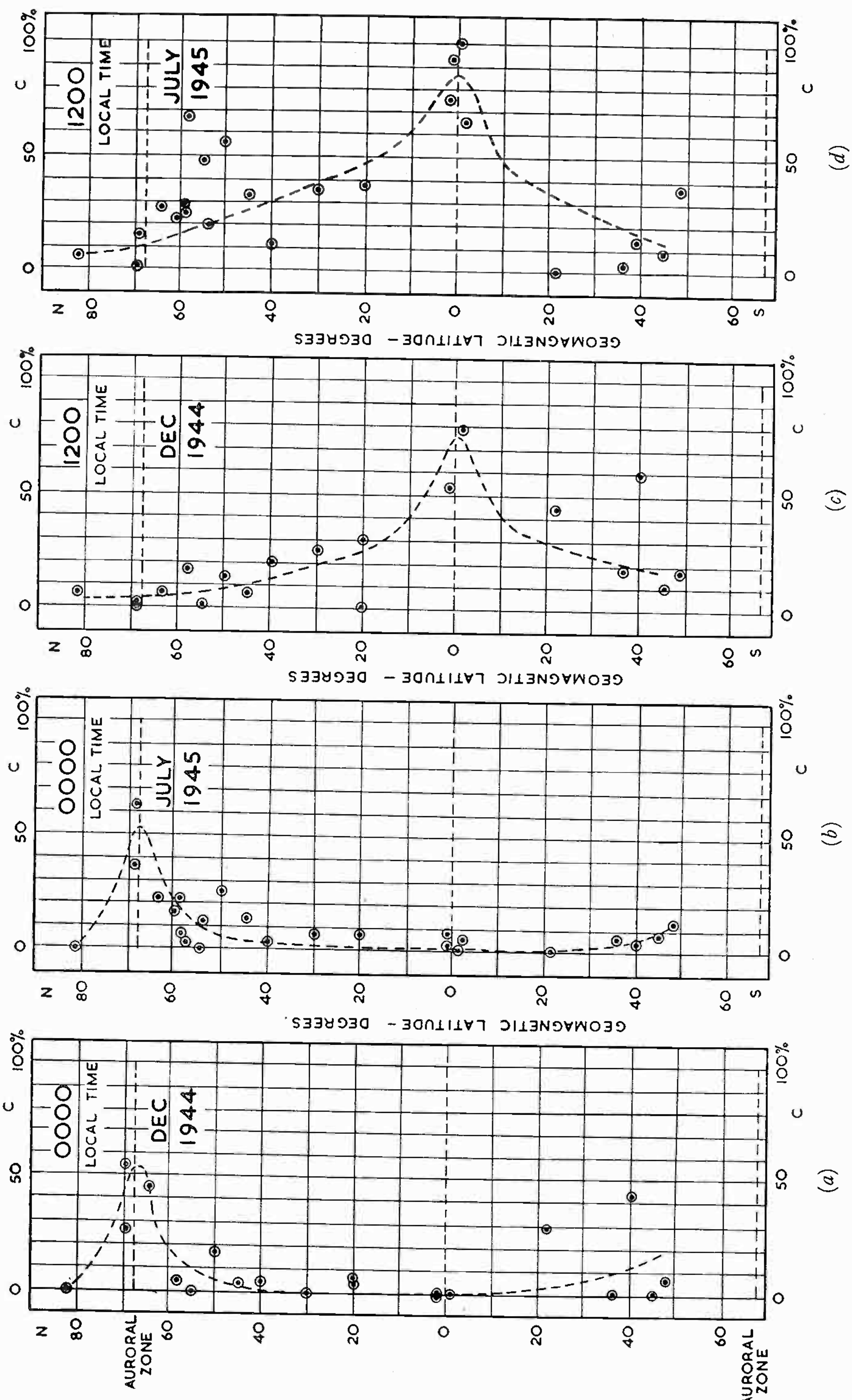
At Churchill, a station situated inside the zone of maximum auroral frequency, the characteristics of the Es ionisation were very different from those at Clyde. The (*c*) values reached at times very high values (83%), particularly during the local hours 2100 to 0200, but there was very little Es ionisation during the daylight hours, and no apparent seasonal variation from month to month (Fig. 2).

Two stations, Great Baddow (Fig. 3) and San Juan (Fig. 4) were selected from those situated in the geomagnetic mid-latitudes, and in both cases the values of (*c*) were very much lower than at Churchill. A definite maximum in the Es ionisation was observed during the summer days (June-August) with a suggestion of two daily peaks, one before, and one after the local noon. During the local night Es ionisation was very weak, and it was noticeable that at San Juan, which is much nearer to the equator than Great Baddow, there was practically no night-time Es ionisation.

This lack of night-time Es ionisation was even more marked at Huancayo (Fig. 5), just south of the geomagnetic equator. There the Es ionisation was confined to the local daytime, a very rapid rise taking place from the value of $(c) = 0$ at sunrise to the high value of (*c*) in the first hour of daylight, often remaining at 100% during the daytime. There appeared to be no obvious monthly variation in the values of (*c*). The highest frequencies of Es reached around noon were also very high and well above 7 Mc/s.

As most of the stations in the southern hemisphere were situated in similar geomagnetic latitudes, only one of them was selected. At this station, Watheroo (Fig. 6), Es ionisation was not confined to the local daytime. The daytime values of (*c*) were much lower than at Huancayo, and a definite maximum was reached in the local summer (November-January).

As there were no stations in the southern auroral zone and to the south of it, this north to south survey cannot be completed. Possibly, the values of (*c*) in these regions are comparable with those obtained for the similar zones in the northern latitudes, on the assumption that in both cases the ionising agency is the same.


0000
LOCAL TIME
DEC
1944
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GEOMAGNETIC LATITUDE – DEGREES
AURORAL ZONE
AURORAL ZONE
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0
20
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60
(a)
(b)
(c)
(d)

Fig. 7

Variations of (*c*) with geomagnetic latitude are shown in Figs. 7 (a—d) for local noon and midnight in December, 1944 and July, 1945, for a number of stations going from the northern to the southern latitudes. The midnight curves show quite clearly a maximum in the northern auroral zone, whilst the noon curves peak around the geomagnetic equator.

This distribution may have a physical significance indicating the different types of ionising agencies.

Not enough information was available to compare values of (*c*) for stations with the same geomagnetic latitude but on opposite sides of the geomagnetic equator.

Longitude Variation

Stations having almost the same geomagnetic latitude were often found to have quite different values of (*c*) for the same local time in any particular month. Sometimes the value of (*c*) differed considerably for relatively near stations and striking examples from Australia are given below:-

0000 Local time December, 1944.	*c* at Watheroo = 43.
	c at Canberra = 0.
1200 Local time March, 1945.	*c* at Watheroo = 6.
	c at Canberra = 45.

Marked differences have also been found between the (*c*) values at Churchill and Reykjavik, although these stations are situated at nearly the same geomagnetic latitude in the northern auroral zone.

Unfortunately, too few records were available for studying the longitude variation on a world-wide scale.

Conclusions

Two belts of intense Es ionisation have been established so far ; a daytime belt in the geomagnetic equatorial zone, and a night-time belt in the northern auroral zone. No information was available about the southern auroral zone.

The different local times at which the maxima in the values of (*c*) occur in these zones suggest different ionising agencies, and actually, it is probable that two limiting types of Es layer occur, one in the arctic zone, and one in the equatorial zone.

Thus, in the northern auroral zone, the Es layer is observed mostly at night. It usually has a definite cut-off frequency, and it is opaque in the sense that the frequency range over which reflections occur simultaneously from the Es and F layers is very small. It often occurs during magnetic storms.

In the equatorial zone the Es layer is observed only in the daytime. It is transparent to F echoes in that reflections of small amplitude occur over a large range of frequencies. Its variations are fairly regular and there is no apparent correlation with magnetic storms.

In the intermediate latitudes both these types can occur but the opaque type is also apparently independent of magnetic storms. The observations suggest that there is a seasonal variation, usually reaching maximum in daytime during the local summer. In a number of cases a longitude variation has been noted, which at times was very considerable, as for example, in the northern auroral zone.

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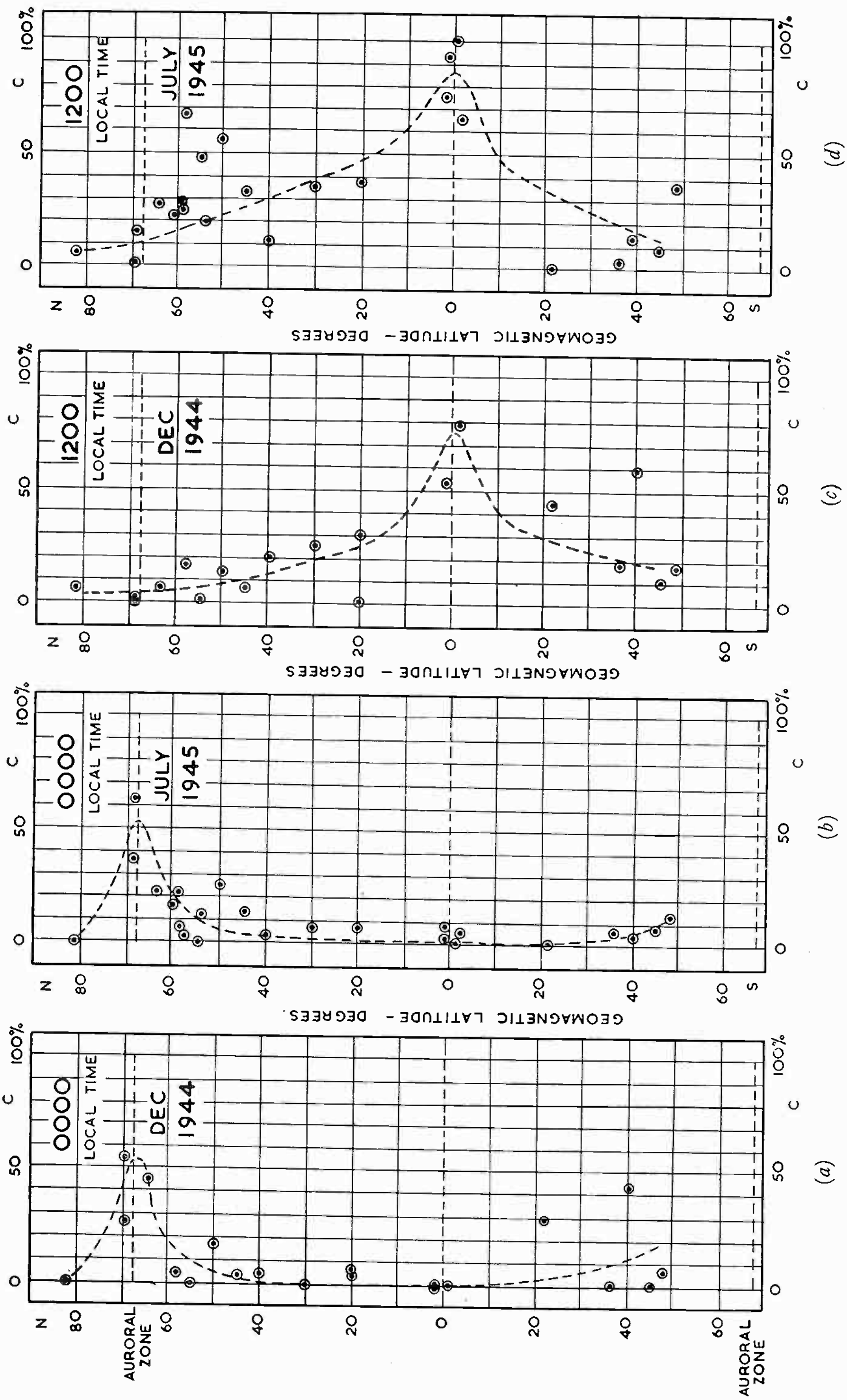
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(c)
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(a)
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N
S
80
60
40
20
0
20
40
60
GEOMAGNETIC LATITUDE – DEGREES
AURORAL ZONE
AURORAL ZONE

FIG. 7

Variations of (c) with geomagnetic latitude are shown in Figs. 7 (a—d) for local noon and midnight in December, 1944 and July, 1945, for a number of stations going from the northern to the southern latitudes. The midnight curves show quite clearly a maximum in the northern auroral zone, whilst the noon curves peak around the geomagnetic equator.

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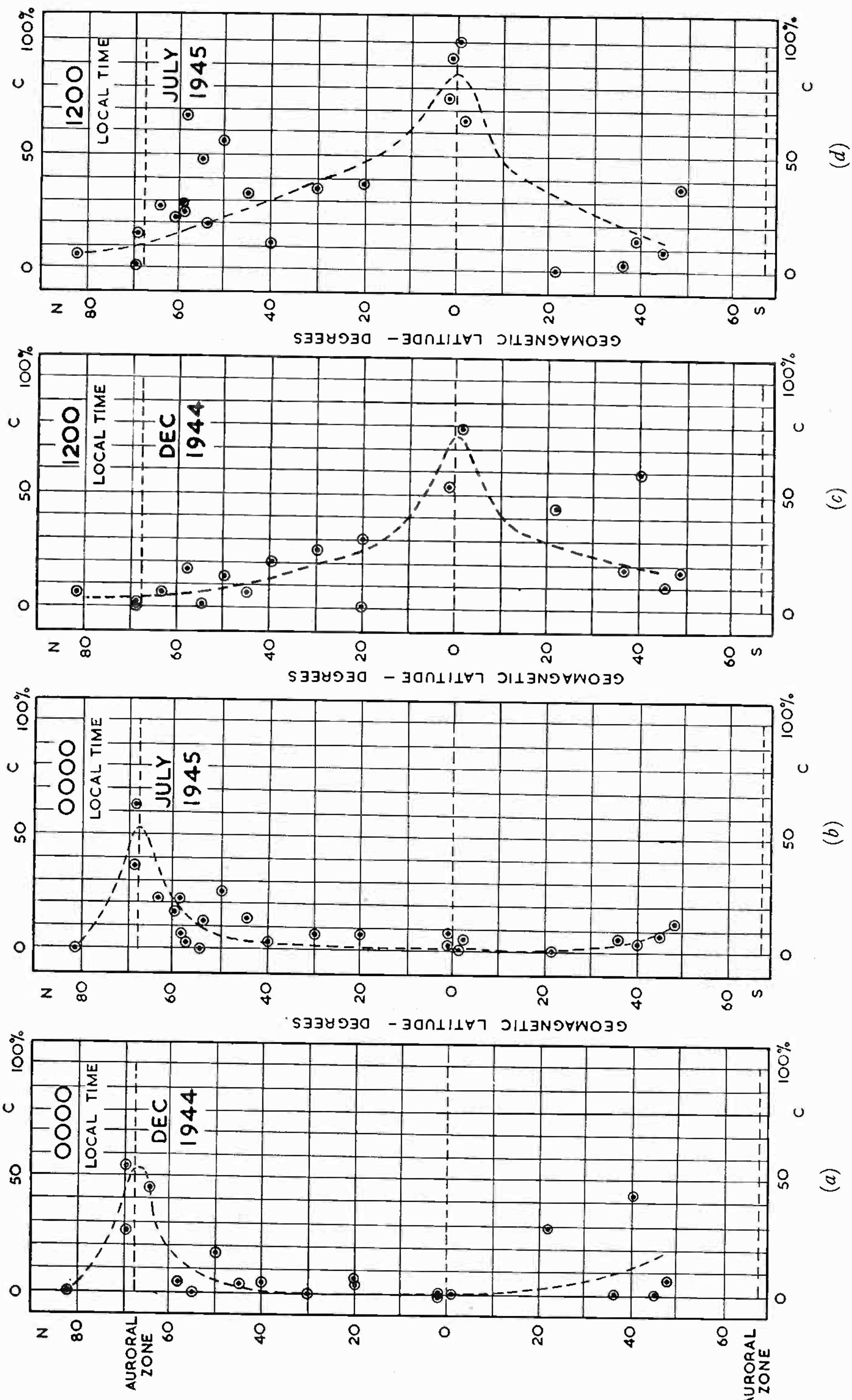
1200
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GEOMAGNETIC LATITUDE - DEGREES
AURORAL ZONE

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A VARIABLE FREQUENCY DRIVE OF HIGH STABILITY FOR AIRCRAFT USE

By Thomas T. Brown, A.M.I.E.E.

The problem of providing a continuously variable frequency drive for aircraft transmitters is examined. The inherent difficulties in the use of simple L.C. oscillators are described, together with a system for overcoming these disadvantages by mixing the output of an L.C. oscillator with that of a multiple crystal oscillator to stabilize the frequency. Reference is made to the production of spurious frequencies in such cases and their reduction by the choice of appropriate frequency ratios and other methods.

The design of such a unit, which includes several novel features, is detailed below. This unit provides frequencies between 2 and 12 Mc/s with a high degree of frequency stability. Microphony, moisture and altitude effects are small. Tuning scale resetting accuracy is high, as the system provides an equivalent scale length of 30 feet between 2 and 12 Mc/s.

First models had an average temperature coefficient of $\pm$ 2 parts in 10^6 per °C. over the temperature range -40°C. to $+55$°C.

Introduction

THE design of a high stability master oscillator for aircraft use involves somewhat different problems from such as would be applicable to one required for a ground transmitter. Changes of temperature, humidity, atmospheric pressure and vibration are far more extreme than are experienced in a normal installation, and size and weight limitations and the problem of tuning up under vibration add further difficulties.

In these circumstances it is very often not possible to follow out many of the recommendations of the recognized authorities on the subject.

The Atlantic City Convention fixed the requirement for frequency stability of aircraft transmitters at $\pm$ ·02% from all causes, so that taking a temperature range of -40 to $+50$°C., a mean temperature coefficient of about four parts per million per °C. is required. This stability is close to that obtainable from a crystal oscillator without oven control. Whilst it is possible to build up a simple L.C. oscillator to have such stability, the design of such an oscillator covering a frequency range of 2-12 Mc/s would probably involve a considerable effort in development and also would necessitate an amount of precision workmanship on the production units to make it a poor economic proposition to manufacture in quantity.

When a requirement for a continuously variable frequency drive unit for the AD107 transmitter occurred, it was decided to utilize a slightly more complicated electrical system if a commensurate saving of mechanical precision work could be obtained. The type AD.1611 Drive Unit was conceived along the lines of this policy.

Choice of System

After investigation several systems, most of which were considered to be too complicated and to have various disadvantages, the system described in detail below was chosen. Briefly, this is a frequency synthesis system, which makes use of the following principles.

The frequency stability improvement obtained from the combination of a low frequency oscillator of moderate stability and a higher frequency oscillator having good stability.

The high stability that can be obtained from a restricted frequency range oscillator in the ·5 Mc/s region.

The expanded frequency tuning scale obtained when a low frequency oscillator is bandspread against a high frequency crystal oscillator.

These three principles are realized by the combination in a balanced modulator

TABLE I

COMPOSITION OF FREQUENCY BANDS

Crystal Frequency mc/s.	L.C. Frequency mc/s.	Output Frequency (Difference) mc/s.	Stabilisation Improvement Factor	Output Frequency (Sum) mc/s.	Stabilisation Improvement Factor
2·5	·4—·5	2·1—2·0	5·25×—4·0×	2·9—3·0	7·25×—6 ×
2·6	,,	2·2—2·1	5·5 ×—4·2×	3·0—3·1	7·5 ×—6·2×
2·7	,,	2·3—2·2	5·75×—4·4×	3·1—3·2	7·75×—6·4×
2·8	,,	2·4—2·3	6·0 ×—4·6×	3·2—3·3	8·0 ×—6·6×
2·9	,,	2·5—2·4	6·27×—4·8×	3·3—3·4	8·25×—6·8×
3·0	,,	2·6—2·5	6·5 ×—5·0×	3·4—3·5	8·5 ×—7·0×
3·1	,,	2·7—2·6	6·75×—5·2×	3·5—3·6	8·75×—7·2×
3·2	,,	2·8—2·7	7·0 ×—5·4×	3·6—3·7	9·0 ×—7·4×
3·3	,,	2·9—2·8	7·25×—5·6×	3·7—3·8	9·25×—7·6×
3·4	,,	3·0—2·9	7·5 ×—5·8×	3·8—3·9	9·5 ×—7·8×
3·5	,,	3·1—3·0	7·75×—6·0×	3·9—4·0	9·75×—8·0×

mixer of the outputs of a crystal oscillator with eleven crystals switched in 100 kc/s steps between 2·5 and 3·5 Mc/s and a continuously variable oscillator of 400-500 kc/s. By using both sum and difference frequencies, the resulting output between 2 and 4 Mc/s is split into twenty sub-bands of 100 kc/s which are selected by the crystal switch with continuous tuning in these sub-bands controlled by the L.C. oscillator. A tunable amplifier following the mixer is used to select the sum and difference terms and also to multiply the frequency of the 2-4 Mc/s range.

The composition of each frequency sub-band in terms of crystal and L.C. oscillator frequencies, and also the factor by which the stability of the L.C. oscillator is improved, is given in table 1, and a simplified circuit in Fig. 1.

In practice, the following advantages are obtained from this system.

(a) A frequency stability of nearly the same order as that of a single frequency crystal controlled oscillator, namely about 2-5 parts per million per degree centigrade.

(b) The reduction of the effects of microphony, atmospheric pressure and humidity on the L.C. unit by the same factor that the temperature coefficient of frequency is reduced.

(c) The provision of a semi-decade system of tuning with scales reading directly in frequency to four-figure accuracy having an effective scale length of 30 feet between 2 and 12 Mc/s with little mechanical complication.

(d) The elimination of hand calibration of scales. Owing to the fact that the frequency range increases in steps of 0·1 Mc/s, the difference between each step depending solely on the accuracy of the crystals, the calibration of the L.C. oscillator scale is not critical and a simple standard engraved scale is sufficient to provide the necessary four-figure accuracy of calibration. For comparison the calibration of a conventional oscillator every 5 kc/s between 2-4 Mc/s would require 400 calibration points or about one division per 0·9° rotation of a circular scale. With the present system only twenty divisions are required, or about one division per 18 degrees rotation.

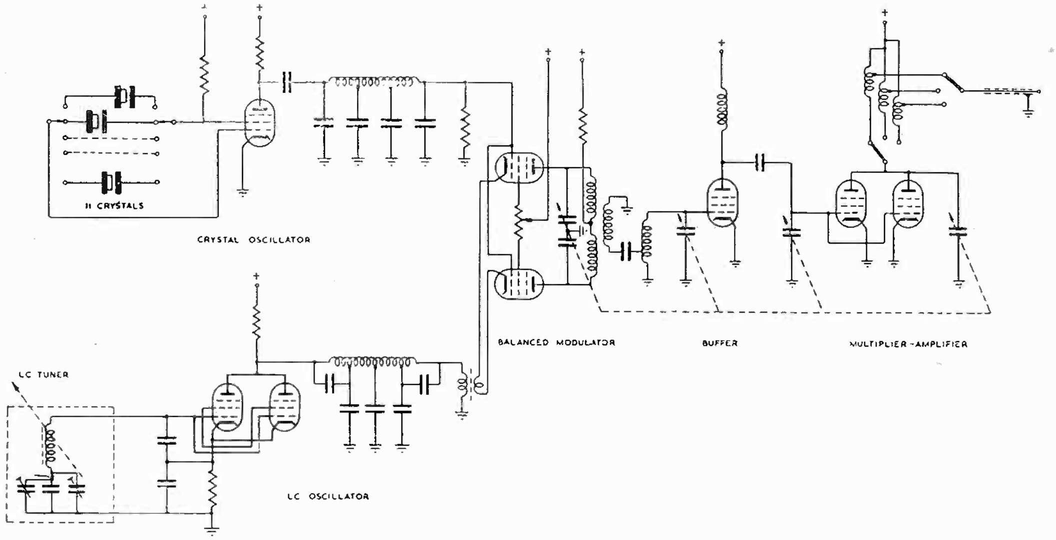

FIG. 1
Simplified circuit diagram.

(e) The reduction of extremely accurate machining and annealing processes in the manufacture of the mechanical assemblies of the L.C. oscillator to a lower order of accuracy and the provision of a design suitable for conventional assembly line practice.

Comparison of Inductance and Capacitance Tuning

Comparing the relative merits of variable capacitance and variable inductance tuning the effect of atmospheric variations on an air dielectric capacitor must be taken into account. The pressure changes from 14·7 lb./sq. in. at sea level to 2·7 lb./sq. in. at 40,000 feet causing a change of about 400 ppm in dielectric constant and capacitance.

The effect of humidity is of the same order, being greater at higher temperature. The dielectric constant of air may be expected to change about 800 ppm at + 50°C. when the humidity varies from 0 to 100 %. Condensation causes even more violent changes, variations on an experimental model of over 2000 ppm having been measured. Another factor requiring serious consideration is that of the microphony caused by mechanical vibration of tuning capacitor blades. According to British Standards Institution data, the amplitude of vibration in the central region of a typical aircraft can be taken as ·01 inch at 10 c/s and ·005 inch at 100 c/s. It is difficult to get a capacitor having a structure that does not exhibit some resonant vibrations under these conditions.

These factors all tend to favour a solid dielectric capacitor with some form of variable inductance tuning. Permeability tuning by moving an iron dust slug into a solenoid was chosen as being the most simple mechanical combination giving adequate rigidity and electrical stability. By a simple modification, described later, a straight line frequency tuning law can also be obtained, and the final structure can be hermetically sealed without a great deal of extra complication.

The Mixing Circuit

The mixing circuit for the production of the required sum and difference frequencies from the two oscillators should generate the wanted frequencies free from spurious carriers, and clear of audible modulation, reduce or eliminate the crystal frequency in the output circuit, and provide a reasonable output and conversion efficiency.

As a prerequisite, it is essential that the mixer be supplied by those frequencies chosen to produce minimum spurious terms, and that these frequencies be themselves free from harmonics. Adequate filtering of the oscillators must therefore be provided before the mixer to reduce the harmonics to at least 60 dB below the fundamental.

In the mixing process, the most important single factor governing the production of spurious terms is the ratio of the two oscillator frequencies. The calculation of all spurious cross modulation frequencies up to, say, the fifth or sixth harmonics when the oscillator frequencies are known, is straightforward, but conversely, the choice of oscillator frequencies giving wanted sum and difference frequencies at an optimum spacing from spurious terms must be carried out by a series of successive approximations, which is tedious when a large number of fixed ratios are required, and becomes virtually impossible when one of the oscillators is continuously variable.

It was not found possible to construct a frequency system with any degree of confidence until a graphical system was devised. This is described in the appendix. The chart (Fig. 7) gives all sum and difference components of harmonics up to the sixth order for any oscillator frequency ratios. When one oscillator frequency is fixed and the other varied, the convergence and divergence of the various spurious terms with respect to the wanted frequency are readily visualised, and it is possible to choose frequency ratios that give a spacing between wanted and spurious terms sufficient to enable the latter to be rejected by the selective circuits following the mixer. Even when such spacing is not possible, it is still possible to choose ratios giving only higher order harmonics having a relatively low level.

For the mixing stage, the choice of a balanced modulator is regarded as essential, for in a single ended mixer the crystal frequency component in the anode circuit may be as much as 20 dB above the wanted terms, and unless this is balanced out by the balanced modulator connection, the selectivity of the succeeding stage may be insufficient to prevent cross modulation in the grid of the amplifier.

Two CV.138 valves are used with cathode injection of the L.C. frequency and the balancing control is a preset potentiometer in the screen grid circuit.

Low Pass Filter Units

Since it is essential that the mixer only receives input signals of a pure waveform, both oscillator outputs are fed through low pass filters giving about 60 dB attentuation above their cut-off frequencies. These are mounted in hermetically sealed screening cans.

The crystal frequency filter is of four constant-K sections whilst the L.C. frequency filter is a composite of two constant K and two m-derived half sections.

In both cases the coils are wound on a common former of bakelite rod, this is permissible, as although the ensuing mutual inductance between coils affects the cut-off frequency slightly, it does not reduce the attenuation.

As the input impedance of the mixer is a balanced 30 Ω, a transformer is needed to match this to the 1000 Ω single-ended output of the L.C. filter. By using a " Ferroxcube " pot core it was possible to get a low enough leakage inductance combined with sufficient primary inductance not to interfere with the filter characteristics.

Crystal Oscillator

The twenty 100 kc/s frequency steps are obtained from eleven miniature crystals covering a range of 2·5 to 3·5 mc/s.

A CV.138 pentode is used as the oscillator in an untuned Pierce circuit. Unfortunately, the 1000 Ω characteristic impedance of the low pass filter is too low an anode load to ensure oscillation, so to obviate this, the crystals are operated between the screen grid and control grid, the output being electron-coupled to the anode. Besides reducing the danger of accidentally over-driving the crystals, this arrangement gives improved isolation from the mixer circuit and better frequency stability.

The L.C. Oscillator

This covers the range 400-500 kc/s and it is on this part of the system that the inherent stability chiefly depends. It is possible to start with several advantages here, as, for instance, this frequency region is one at which an L.C. combination tends to be most stable. The restricted frequency range is advantageous, as the temperature compensation varies somewhat with frequency, and if this range is restricted, exact compensation is easier to obtain.

The oscillator circuit chosen is the modified Colpitts described by Gouriet and Clapp. In this circuit the effect of the maintaining circuit on the frequency is small and the fact that a tapped coil is not required is convenient for production.

The circuit suffers from variation of output voltage with frequency with capacitative tuning, but this does not occur if inductive tuning is used, and in fact, on the prototype model, the output voltage remains constant to within 5% over the tuning range.

By using a pentode, the anode is effectively electron-coupled, and load variations of the anode circuit have little effect on the frequency, so that a buffer stage is not required between the oscillator and the mixer. In fact, the low-pass filter is inserted directly in the anode circuit with no apparent deleterious effects.

The conditions of oscillation are such that the total coupling capacitance

$$C_g = \tfrac{1}{2}\sqrt{\frac{Gm\, C_o\, Q}{2\pi f}}$$

where C_0 is the tuning capacitance.

From this it is seen that to obtain a large C_g/C_o ratio and hence maximum stability, the mutual conductance of the valve G_m should be as high as possible. With the pentode connection however, the screen acts as the effective anode and G_m is the control grid-screen conductance. This is rather low in the CV.138 valve, and so both to increase this factor and to ensure sufficient output across the 1000 Ω anode load, two CV.138's are used in parallel. The frequency stability with supply voltage variation of this circuit is very good; changing the H.T. of production models from 100 to 300 v. only changed the frequency by 2 c/s at ·5 mc/s.

L.C. Tuning Unit

The L.C. tuning unit is basically a solenoid coil traversed by an iron dust slug. An interesting feature is the use of pyrophillite for the coil assembly. This material, a form of soapstone, can be accurately machined by any of the conventional means

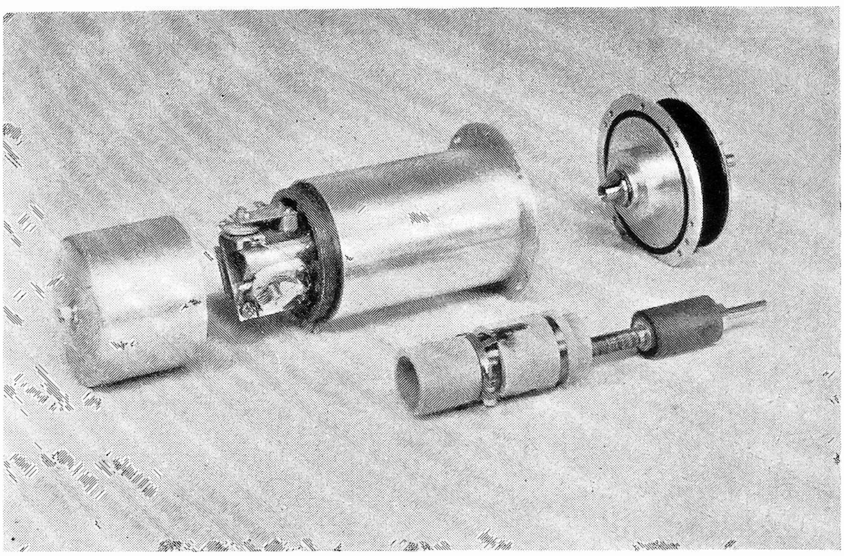

FIG. 2
L.C. Tuner unit assembly.

in the raw state; if fired to 1350°C. it is converted with a minimum dimensional change to a material having the mechanical and electrical properties of ceramic. The coil former, machined from pyrophillite, is clamped between the ends of a cylindrical spun aluminium can, the faceplate of which carries a leadscrew mechanism, a tongue and slot in a hollow control shaft driving a three-start thread to give 1 inch travel for 5 turns. The bearings incorporate two neoprene hermetic seals.

The silvered mica tuning condenser, trimmer and temperature compensator are mounted on an assembly at the other end of the can.

An end cap is fitted over this assembly and screws on to the end of the can. A ceramic bush cemented into this aluminium cap with " Araldite " provides a sealed output lead. The complete unit is sealed by neoprene gaskets. A 5 : 1 stepdown gear concentric with the control shaft drives a calibrated frequency dial mounted on the faceplate thus making the tuner a self-contained unit, which is interchangeable between oscillators without further calibration. A view of the component parts of the dismantled unit is given in Fig. 2.

Temperature and Humidity Compensation

The temperature coefficient of the silvered-mica tuning capacitor is about + 15 to + 20 ppM/°C. That of the coil is about + 40 to + 60 ppM/°C. according to the position of the slug. The position of the slug is also varied with temperature owing to the expansion of the brass leadscrew. The use of an invar leadscrew to obviate this was not successful owing to the large temperature coefficient of resistivity of invar, which when inductively coupled into the coil produced greater drift than that produced by the mechanical expansion of the brass leadscrew. Copper plating of the invar was not found practicable at this frequency.

By anchoring the coil to the supporting can at the opposite end to the leadscrew fixing, the expansion of the can can be made to move the coil in opposition to the movement of the slug. This method of fixing is employed to reduce the temperature coefficient of the coil. In addition to this, a temperature compensating capacitor, as used in the RG.44 and AD.94 receivers, which consists of an adjustable single plate capacitor carried on a bimetallic strip, is used to provide a preset amount of

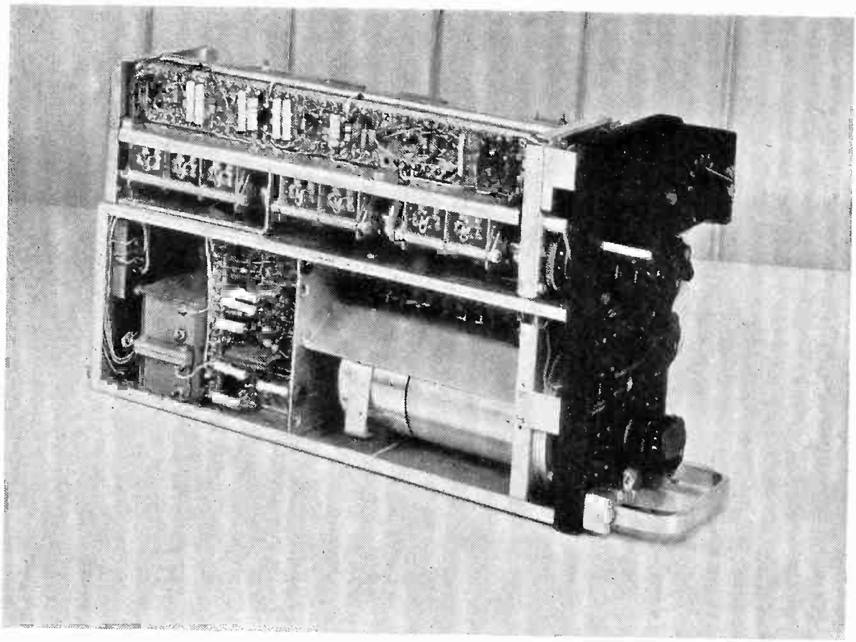

FIG. 3
Left-hand view of unit with covers removed.

compensation to take up batch variations in temperature coefficient of the silvered mica capacitors and coils.

Humidity effects are minimised by heating the complete assembly to 110°C. for an hour with the end cap off. This is then replaced with a silicone-greased gasket before the unit is quite cool. This process is also essential as an annealing operation to relieve stresses in the coil windings and mechanical structure.

Tuning Law

To ensure a straight line frequency response on the variable inductance, an auxiliary coil on an adjustable pyrophillite former mounted over the far end of the main solenoid is connected in series to prevent the rate of inductance variation from falling off when the slug approaches the end of the main coil. By choice of the number of turns, position and dimensions of the auxiliary coil, good linearity can be obtained.

In fact, the maximum departure from a straight line law is less than 1 kc/s at the far end of the 400-500 kc/s range. Adjustment of the position of this auxiliary coil is used to control the slope of the tuning law and hence as a fine trimmer to line up the calibration of the dial.

The Amplifier Multiplier

This section is a three valve, two stage tuned amplifier of conventional design. The tuning capacitors are ganged with those of the mixer circuit.

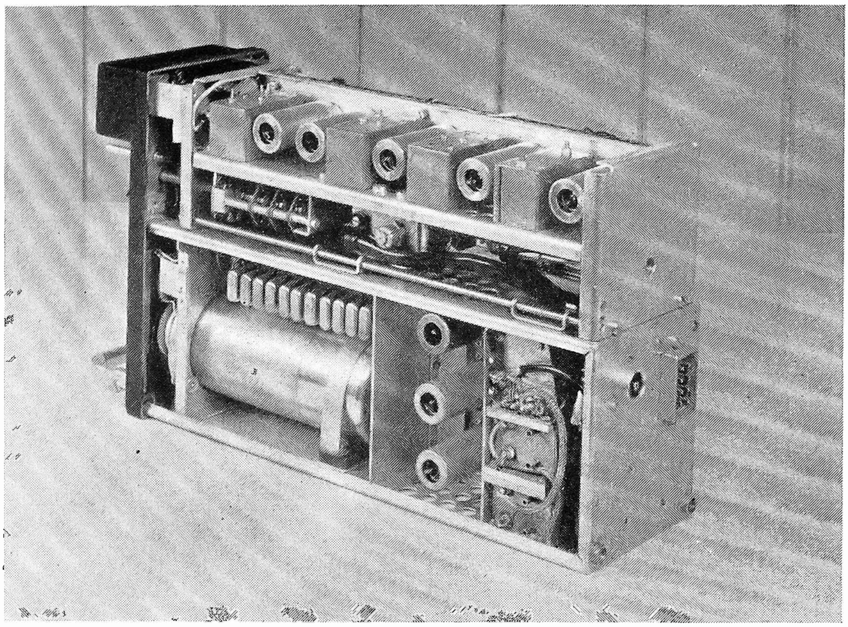

FIG. 4
Right-hand view of unit.

The amplifier performs the following functions. It selects the sum and difference frequencies and mechanically exposes the appropriate oscillator dials on the front panel. It provides selectivity to discriminate against spurious frequencies and amplifies the low level output of the mixer to provide sufficient power to drive the transmitter, finally providing frequency multiplication to extend the frequency ranges.

Since the latter two requirements have to be met regardless of whatever type of stable M.O. were used, much of the apparent complication of the crystal—L.C. system disappears.

When using sum frequencies an increase of L.C. oscillator frequency gives an increase of output frequency, but on a difference frequency will give a reduction of output frequency. This necessitates two sets of scales, one for each condition. To obviate ambiguity, a mechanical interlock is provided on the ganged capacitor which operates a shutter to reveal the right-hand scales between 2-3 Mc/s, and the left-hand ones between 3-4 Mc/s.

Frequency multiplication is provided by switching coils, trimmed to exact multiples of the fundamental frequency, so that the " X1 ", " X2 ", " X3 " switch multiplies the output frequency without retuning.

A d.c. potential proportional to the grid current of the output valves of the A.D.107 Amplifier Unit is applied as A.G.C. to the amplifier, and maintains the output level at a value to give optimum drive despite varying L/C ratios of proceeding circuits.

In the final stage, two CV.138 pentodes are used in parallel rather than one larger valve, thus keeping all the valves in the unit to one type, and so simplifying maintenance.

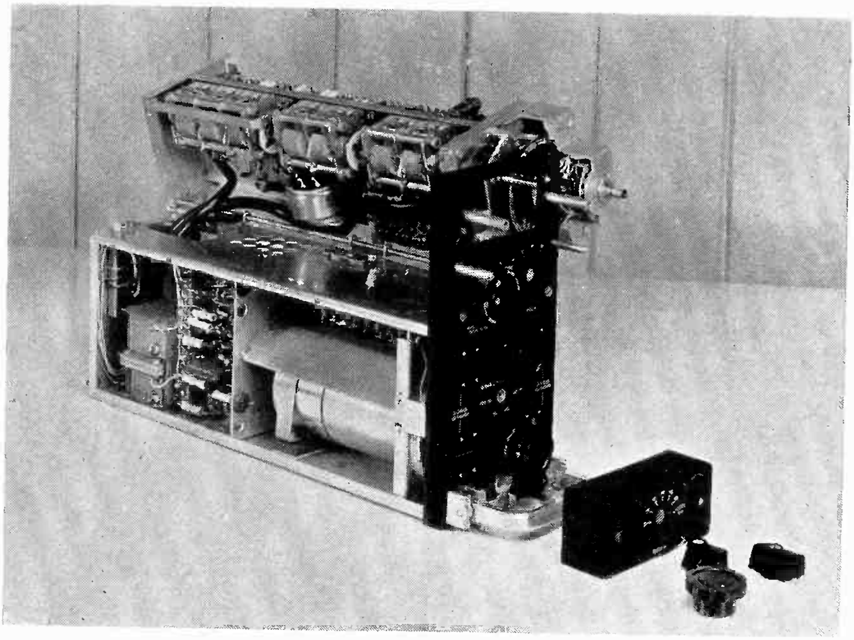

Fig. 5
Unit hinged open.

General Mechanical Assembly

The complete unit which fits standard aircraft racking, is shown in Figs. 3 and 4 and consists of two vertically mounted sub-units. The upper section containing the mixing and amplifying circuits and the lower the two oscillators and associated filters. The overall dimensions are 4 inches $\times$ 8 inches $\times$ $12\frac{1}{2}$ inches.

The upper section may be withdrawn from the front panel by removing the knob and fixing screws, whence sliding hinges enable it to be rotated sideways to gain access for servicing whilst the unit is still operating. A view of the unit with the upper section open is shown in Fig. 5.

Five of the valves are mounted at the top of the upper section, ensuring that the heat dissipated does not affect the frequency controlling components. The top section components are mounted on a long metal tagboard using ceramic bush terminals as first used in the AD.107 equipment.

The L.C. tuning assembly is self contained in a hermetically sealed unit, which may be withdrawn, complete with dials and gears, from the front of the lower section.

The sub-chassis containing the three oscillator valves may be withdrawn complete with valves by removing four screws. This forms its own tagboard by the use of ceramic bush terminals inserted along L bends in the chassis, and hence is readily wired as a complete sub-assembly. The mounting of this sub-chassis serves as structural bracing to prevent chassis wringing, and as a funnel to conduct heat from the oscillator valves away from the L.C. tuner.

The amplifier tuning control is geared down by a 2 : 1 slow-motion which operates

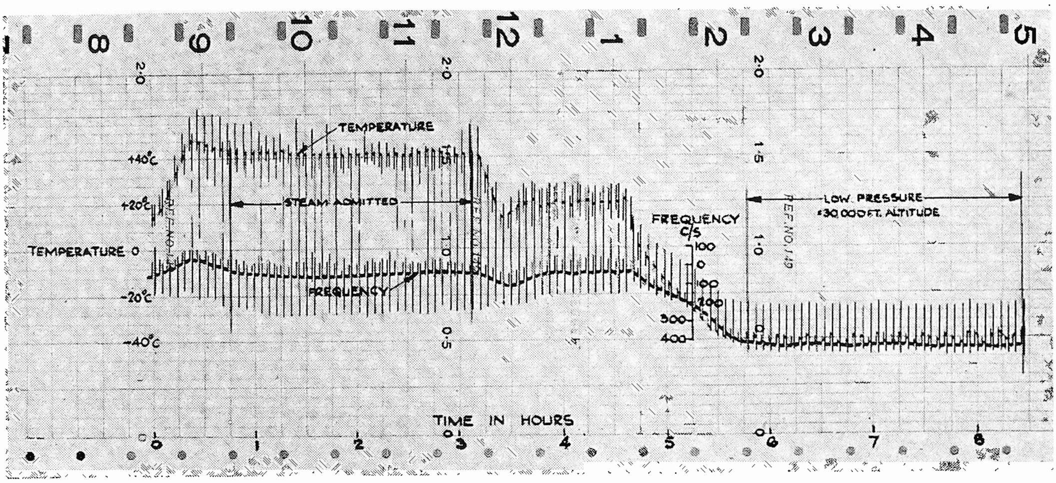

FIG. 6

Specimen frequency run.

a shutter by a cam mechanism arranged to automatically display the correct frequency scales when either the sum or difference frequency is tuned in. These perspex scales use edge-on illumination. A concentric knob locking device was developed to prevent accidental movement of the tuning controls after setting up.

Performance Tests

The first two models were subjected at various times to climatic cycles which included temperature ranges from —40°C. to +70°C., atmospheric pressure from 15 to 2·7 lbs./square inch, steam injection and also to vibration amplitudes of 0·5 thou. from zero to 100 c/s. The frequency drift was continuously measured by the automatic drift recorder developed in the Frequency Measurement Section at Baddow. A typical record of a day's run is shown in Fig. 6 which shows ambient temperature raised to +40°C., steam injected, temperature dropped to —40°C. and the atmospheric pressure reduced to an equivalent height of 30,000 feet.

The prototype model, when put through the standard K.114 seven-day cycle of climatic changes between these limits, gave a mean drift of 2·1 c/s per Mc/s. per degree C. at 3·251 Mc/s. The frequency returning to within 50 c/s of its original setting after nine days in the climatic tank and after a final vibration test.

The second model gave a mean drift of 1·8 c/s per Mc/s. per degree C. in its various tests.

Initial production models, whilst not having to undergo such comprehensive tests, are showing comparable or better stabilities on standard heating cycles to 60°C.

It was generally observed that all models show a better frequency stability on the hot cycles than on the cold. One model, for instance, only drifted about 50 c/s at 3·2 Mc/s on raising the temperature to 70°C., but drifted 400 c/s on reducing the temperature to —40°C. It is considered that the law of capacitance change with displacement of the bimetallic compensating condenser is a probable explanation of this effect.

Spurious radiation was measured in the output of the unit directly and also by driving an AD.107 transmitter and measuring field strength at a remote receiving station. Independent measurements were also carried out on the spurious radiation by means of a spectrum analyser.

Summary of performance of 1st Model

Frequency ranges. 2-4, 4-8, 6-12 Mc/s.

Frequency Stability.

Temperature	=mean drift, 2-4 cycles/Mc/s degree C. over range —40°C. to +55°C.
Humidity	=less than 10 cycles/Mc/s over range normal to 100% humidity at +50°C.
Altitude	=2 cycles/Mc/s/1,000 feet over range 0 to 40,000 feet.
Supply Variations	=2 cycles/Mc/s for 10% variation.
Vibration	=No audible microphony with ·0005 inch vibration amplitude between 0-100 c/s.

Spurious Frequencies. Spurious carriers: 60 dB below carrier level of transmitter. Spurious audible modulation: 40 dB below carrier level.

APPENDIX

The Frequency Mixing Chart

The problem is to predict all intermodulation frequencies occurring between any two mixed oscillator frequencies. In the specific case one frequency is fixed and the other is variable, and it is required to find a band of frequency ratios that give the main sum and difference frequency bands sufficiently clear of the intermodulation frequencies.

The expression for each component is given by:—

$$f_s = mf_0 \pm nf_x$$

where m and n are integers and correspond to the order of the harmonics of the two oscillator frequencies f_0 and f_x.

This expression is similar to that of the equation of a straight line and if by a simple transformation it could be converted into this form, a graphical display of all harmonic components would occur as a rectilinear chart.

This is done by dividing by f_x and utilising the frequency ratios

$$\frac{f_s}{f_x} = \frac{m f_0}{f_x} \pm n$$

which is now in the required form $y = a\,x + b$, the equation of a family of straight lines with axes $\frac{f_s}{f_x}$ and $\frac{f_0}{f_x}$ having slopes m and intercepts n.

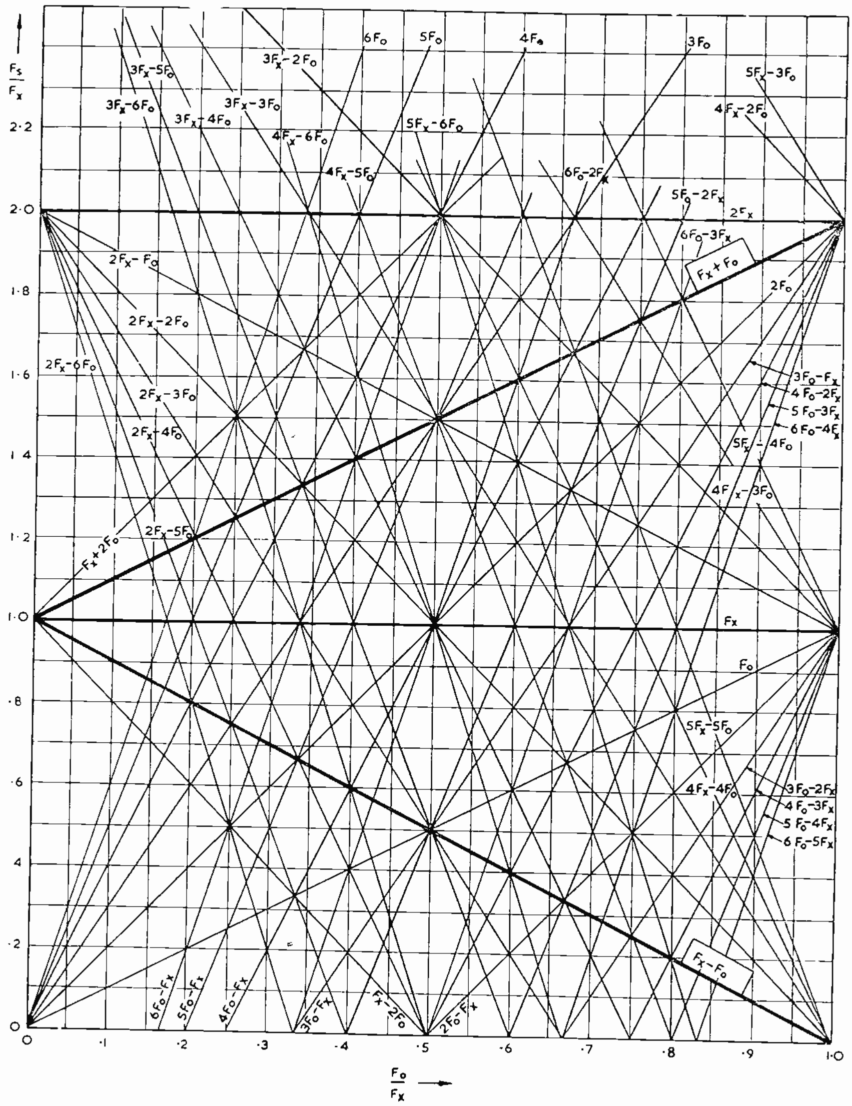

FIG. 7

Frequency mixing chart.

The chart is constructed for all harmonics up to the sixth order. For any one value of f_x (which case applies when a variable oscillator f_0 is mixed with a crystal frequency f_x) the axes can be labelled directly in frequency.

The two main sum and difference terms are read off the two appropriate thick lines, whilst any intermodulation frequency can be read off the correspondingly labelled line.

Consider a practical case when the crystal is 1·0 mc/s and the variable oscillator frequency f_0 goes from ·5 to ·7 mc/s. On the sum curve the output frequency f_s goes from 1·5 to 1·7 mc/s and on the difference curve from ·5 to ·3 mc/s. On the sum frequency when the variable oscillator is ·5 there is heavy intermodulation caused by the 3rd harmonic of f_0, and the difference frequency of f_0 and 2nd harmonic of the crystal. Other terms interfering are $3 f_x - 3 f_0$, $4 f_x - 5 f_0$ and $5 f_0 - f_x$.

When the $\frac{f_0}{f_x}$ ratio is not exactly integral these frequencies beat and produce audible modulation which cannot be filtered out with normal selective circuits. As the variable frequency rises, these terms diverge from the wanted carrier as upper and lower sidebands, and when f_0 is ·6 mc/s we have a wanted carrier of 1·6 mc/s and spurious carriers of 1·0, 1·2, 1·4, 1·8 and 2·0 mc/s. There is also some modulation on the wanted carrier.

In the crystal-L.C. system used in the variable frequency drive described, a choice of frequency ratios $\frac{f_0}{f_x}$ between ·1 and ·2 was made.

From the selectivity standpoint, this is about the best compromise between larger ratios giving heavy intermodulation and spurious frequencies, and smaller ratios requiring extremely selective circuits to separate out the sum, difference and crystal frequencies which are then very closely spaced. This value is also a reasonable compromise between the amount of frequency stabilisation obtainable and the number of crystals required to cover the band.

Acknowledgment

The writer is indebted, amongst many others, to Messrs. P. F. Turner and D. West who carried out the mechanical design, and Mr. E. J. Cole who performed much of the experimental work. Messrs. N. Lea and C. R. Ince for advice and details of the frequency drift recorder. Acknowledgment is also due to the McGraw Hill Publishing Co. for their kind permission to publish the mixing chart (Fig. 7) which originally appeared in " Electronics," April, 1951, p. 134. Mr. L. R. Mullin for the benefit of his experience with high stability receivers, and finally, Messrs. G. P. Parker and C. W. Rich for their guidance and helpful suggestions throughout the course of development.

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DESIGN OF LOW PASS LADDER NETWORKS TO WORK BETWEEN UNEQUAL RESISTANCES

By E. Green, M.Sc.

In a previous paper[1] formulae and curves were given for the design of ladder networks to give Butterworth (Type B) or Chebyshev (Type C) response in the pass band. These were in terms of the decrements (dissipation factors) of the terminations. This is convenient for band pass networks, as the results are independent of changes in impedance level, but for low pass networks it is often the terminating resistances that are given. For matched terminations and odd numbers of branches, no difficulty arises, since this condition requires equal terminating resistances, and equal decrements; but for all cases with unequal terminating resistances some conversion data are needed to enable the original formulae to be used, and these are discussed below.

Symbols

THE symbols used are those employed in the previous paper[1] on this subject but are repeated below for convenience. For the low pass prototype of Fig. 1,

ω_β = radian frequency defining the normal limit of the pass band. For Butterworth response this may be fixed at any level (often 3db). For Chebyshev response it is fixed at valley level.

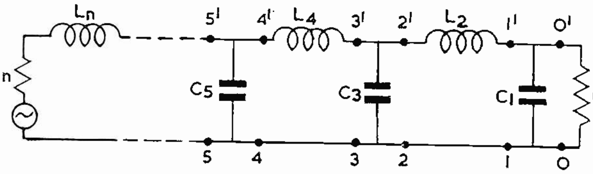

Fig. 1

General Low Pass Network.

ω = variable radian frequency.

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

Branches are numbered consecutively from the output terminals, and the elements making up the branch carry the number of the branch as suffix.

n = total number of branches in the network.

r = identification number of any branch.

$G_1 = \dfrac{1}{R_1}$ = conductance of output branch. (For a series branch we should use R_1).

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

$d_1 = \dfrac{G_1}{C_1\,\omega_\beta}$ = decrement at bandwidth ω_β for branch 1, or $\dfrac{R_1}{L_1\,\omega_\beta}$ in the case of a series branch.

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

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

If $\omega_r = \dfrac{1}{\sqrt{L_2 C_1}}$ = resonant frequency of L_2 and C_1

$K_{12} = \dfrac{1}{\sqrt{L_2 C_1}\,\omega_\beta} = \dfrac{\omega_r}{\omega_\beta}$ = coupling factor at ω_β between branches 1 and 2.

$K_{23} = \dfrac{1}{\sqrt{L_2 C_3}\,\omega_\beta}$, etc.

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

$K^2_{r(r+1)} = \dfrac{\text{reactance of shunt element at } \omega_\beta}{\text{reactance of series element at } \omega_\beta}$.

With constant input voltage or current at all frequencies

$V \quad$ = Output voltage at any frequency ω.

V_p = Output voltage at peaks of response curve.

V_β = Output voltage at edge of pass band and valley level.

$\gamma_n = \left\{\left(\dfrac{V_p}{V_\beta}\right)^2 - 1\right\}^{-\frac{1}{2n}}$

S_n = sinh a where $\sinh^2 na = \left\{\left(\dfrac{V_p}{V_\beta}\right)^2 - 1\right\}^{-1}$

For the conversion formulae either end of the network can be taken as the input. For odd numbers of branches and series terminations we work in terms of R_1 and R_n and arrange that $R_1 \geqslant R_n$, whilst for shunt terminations we use G_1 and G_n and $G_1 \geqslant G_n$. For an even number of branches $R_1 \geqslant R_n$ for series input, and $G_1 \geqslant G_n$ for shunt input. We shall work in terms of series input and R_1 and R_n, but by the principle of duality the results hold for shunt input using G_1 and G_n.

Type B Response (Butterworth)

It can be shown by equating alternative expressions for the reflection coefficient[2] at $x = 0$, that

$$\rho_n = \left\{\frac{1-D}{1+D}\right\}^n = \frac{R_1 - R_n}{R_1 + R_n} \quad \text{or} \quad \left(\frac{G_1 - G_n}{G_1 + G_n}\right)_{\text{shunt}} \tag{1}$$

so that

$$D = \frac{1 - \left\{\dfrac{1 - R_n/R_1}{1 + R_n/R_1}\right\}^{\frac{1}{n}}}{1 + \left\{\dfrac{1 - R_n/R_1}{1 + R_n/R_1}\right\}^{\frac{1}{n}}} \tag{2}$$

The relation between D and R_n/R_1 has been plotted in Fig. 2 for $n = 2, 3, 4$ and 5. For $n = 4$, as D varies from 0·5 to 2·0, (R_n/R_1) only varies by about $\pm$ 3%. Thus when R_n/R_1 (or G_n/G_1) is known D can be found, and to complete the solution we have from reference 1, equations (2) and (3).

$$d_1 = \frac{\gamma_n}{(1+D)\sin\theta} = \frac{\left\{\left(\frac{V_p}{V_\beta}\right)^2 - 1\right\}^{-\frac{1}{2n}}}{(1+D)\sin\theta}, \quad \theta = \frac{\pi}{2n} \tag{3}$$

$$K^2_{r(r+1)} = \frac{(\cos^2 r\theta + D^2 \sin^2 r\theta)\, d_1{}^2 \sin^2\theta}{\sin(2r-1)\theta \sin(2r+1)\theta} \tag{4}$$

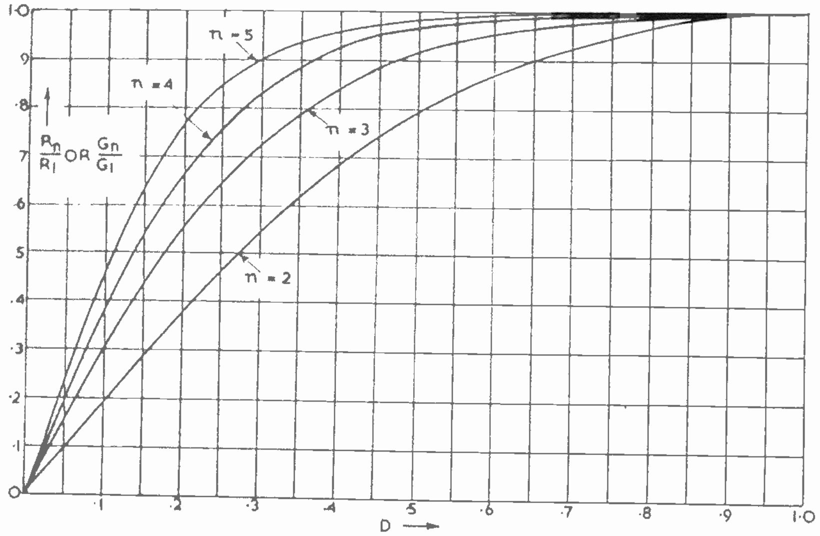

FIG. 2

Type B Response:—Variation of R_n/R_1 *or* G_n/G_1 *with* D, *for* n = 2, 3, 4 *and* 5. *For series input use* R_n/R_1, *and for shunt input* G_n/G_1.

Type C Response (Chebyshev)

Odd number of branches

Again by equating alternative expressions for the reflection coefficient[2] at $x = 0$ we have for series terminations

$$\frac{\sinh na'}{\sinh na} = \frac{R_1 - R_n}{R_1 + R_n} \tag{5}$$

with

$$\sinh na = \left\{\left(\frac{V_p}{V_\beta}\right)^2 - 1\right\}^{-\frac{1}{2}} \tag{6}$$

so that

$$\sinh na' = \frac{1 - R_n/R_1}{1 + R_n/R_1} \cdot \left\{\left(\frac{V_p}{V_\beta}\right)^2 - 1\right\}^{-\frac{1}{2}} \tag{7}$$

When n, (V_p/V_β) and (R_n/R_1) are known, the values of a and a' can be found from tables. Then

$$\frac{1-D}{1+D} = \frac{\sinh a'}{\sinh a}$$

so that

$$D = \frac{\sinh a - \sinh a'}{\sinh a + \sinh a'} \tag{8}$$

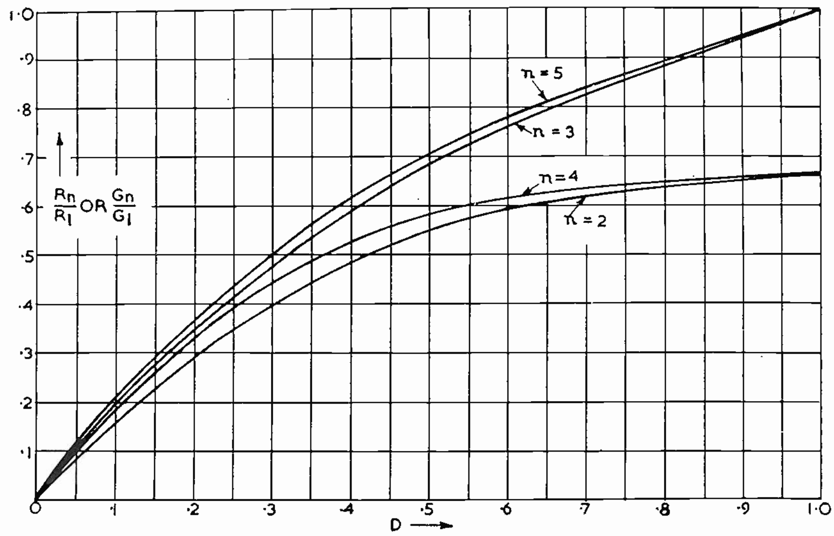

FIG. 3

Type C Response:—Variation of R_n/R_1 *or* G_n/G_1 *with* D *for* n = 2, 3, 4 *and* 5. *For series input use* R_n/R_1 *and for shunt input* G_n/G_1.

The relation between (R_n/R_1) and D for $(V_p/V_\beta) = 1{\cdot}02$ has been plotted in Fig. 3 for $n = 3$ and 5.

Even number of branches

When n is even, the terminations are unlike; i.e. one is series and one shunt. At $x = 0$ there is a minimum of transmission and the reflection coefficient is a maximum. Even in the matched condition the terminal resistances cannot be equal. The limiting ratio which occurs in the matched condition is fixed by the permissible ripple in the pass band. If E is the internal EMF of the generator, and the network is matched at the peaks of response, the power then delivered to R_1 is $E^2/4R_n$. At $x = 0$, R_1 and R_n are in series and the power delivered to R_1 is $E_1{}^2R_1/(R_1 + R_n)^2$. The ratio of output powers for these two conditions is $(V_p/V_\beta)^2$. Therefore

$$\left(\frac{V_p}{V_\beta}\right)^2 = \frac{E^2/4R_n}{E^2\,R_1/(R_1+R_n)^2} = \frac{(R_1+R_n)^2}{4R_1\,R_n} \tag{9}$$

$$\left(\frac{V_p}{V_\beta}\right) = \frac{1+\dfrac{R_n}{R_1}}{2\left(\dfrac{R_n}{R_1}\right)^{\frac{1}{2}}} \tag{10}$$

This limiting relation between R_n/R_1 and $(V_p/V_\beta) - 1$ is independent of n and is plotted in Fig. 4. Only values of R_n/R_1 (or G_n/G_1) lying on or below the curve are permissible.

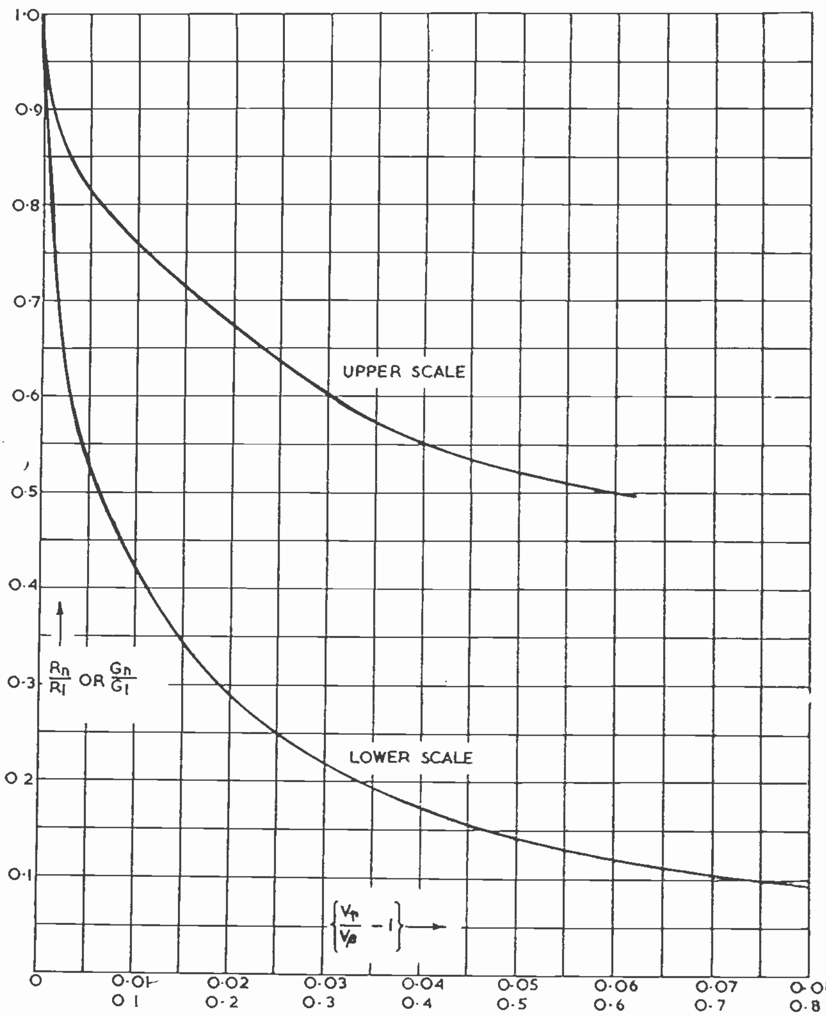

FIG. 4

Type C Response with n *even. Curves show the required relation between* R_n/R_1 *(or* G_n/G_1*) and the ripple* $(V_p/V_\beta - 1)$ *for matched terminations* (D = 1.0). *Permissible values for mismatched terminations lie below the curves.*

For the relation between (R_n/R_1) and D we find from the reflection coefficient at $x = 0$

$$\rho_n = \frac{\cosh na'}{\cosh na} = \frac{R_1 - R_n}{R_1 + R_n} \tag{11}$$

where

$$\cosh na = \left\{1 - \left(\frac{V_\beta}{V_p}\right)^2\right\}^{-\frac{1}{2}} \tag{12}$$

so that

$$\cosh na' = \frac{R_1 - R_n}{R_1 + R_n} \left\{1 - \left(\frac{V_\beta}{V_p}\right)^2\right\}^{-\frac{1}{2}} \tag{13}$$

Given (V_p/V_β) and R_n/R_1 and n, a and a' can be found from tables. Then as for n odd

$$D = \frac{\sinh a - \sinh a'}{\sinh a + \sinh a'} \tag{14}$$

The relation between R_n/R_1 for $V_p/V_\beta = 1{\cdot}02$ and $n = 2$ and 4 are shown in Fig. 3

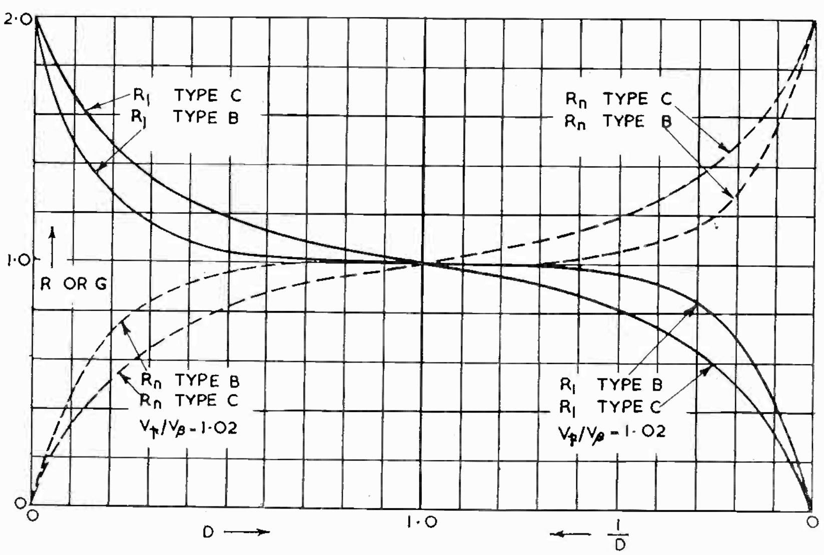

FIG. 5

Networks with odd numbers of branches. Curves show typical relation between R_1 *and* R_n *(series terminations). or* G_1 *and* G_n *(shunt terminations) and* D. $R_1 + R_n$ *(or* $G_1 + G_n$*) assumed constant.*

Solution for any value of n

When we have found D and a as shown above then from Ref. 1, equations (5) and (6)

$$d_1 = \frac{\sinh a}{(1 + D) \sin \theta}, \qquad \theta = \frac{\pi}{2n} \tag{15}$$

$$K^2_{r(r+1)} = \frac{\sin^2 r\theta \cos^2 r\theta + (\cos^2 r\theta + D^2 \sin^2 r\theta)\, d_1{}^2 \sin^2 \theta}{\sin (2r - 1)\, \theta \sin (2r + 1)\, \theta} \tag{16}$$

which completes the solution.

The apparent consistency of the curves of Fig. 2, for type B response for all values of n, conceal underlying differences between odd and even values of n. These begin to show in Fig. 3 for type C response where equal values of R_1 and R_n are not possible. These differences appear more clearly when we plot actual values of R_1 and R_n against D. For odd values of n, and assuming $R_1 + R_n = \text{constant}$ we get the curves of Fig. 5.

No such assumption is possible for even values of n, since the terminations are unlike. When $D = 0$, $R_n = 0$. As R_n and d_n increase, we should expect d_1 to decrease, but R_1 to increase. The general trend is shown in Fig. 6, with both R_1 and R_n increasing continuously as D increases. For Type B response the two curves touch at $D = 1{\cdot}0$ but for Type C response they are separate through the whole range.

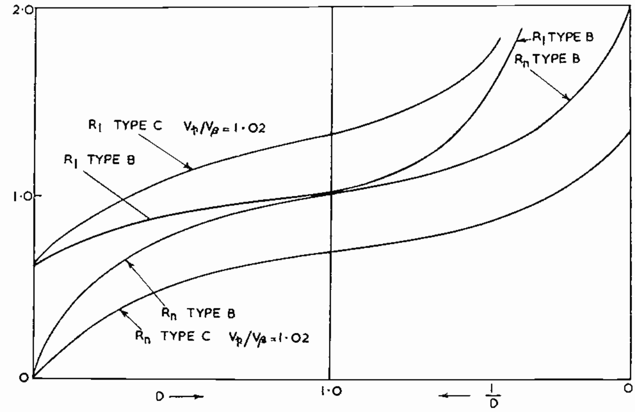

Fig. 6

Networks with even numbers of branches. Curves show typical relation between R_1 *and* R_n *(series terminations) or* G_1 *and* G_n *(shunt terminations) and* D*; not to scale.*

Example.—Type C Response

$$R_n/R_1 = 0{\cdot}6, \qquad V_p/V_\beta = 1{\cdot}02, \qquad n = 4.$$

The network will therefore be as Fig. 7 with a series input. Then from Fig. 3.

$$D = 0{\cdot}54$$

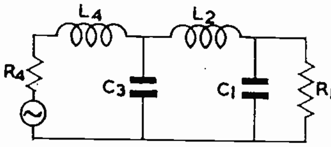

Fig. 7

Four Branch network.

From Ref. 1, Fig. 4b, or by calculation

$$\sinh a = 0{\cdot}61$$

and

$$\frac{1}{\sin\theta} = 2{\cdot}61$$

$$\therefore \quad d_1 = \frac{\sinh a}{(1 + D)\sin\theta} = \frac{{\cdot}61 \times 2{\cdot}61}{1{\cdot}54} = \underline{1{\cdot}03}$$

$$d_4 = Dd_1 = \underline{0{\cdot}56}$$

From Ref. 1, Fig. 33, or by calculation from equation (16)

$$K_{12}^2 = 0{\cdot}73, \qquad K_{23}^2 = 0{\cdot}4, \qquad K_{34}^2 = 0{\cdot}52$$

Then

$$\frac{1}{C_1\,\omega_\beta\,R_1} = d_1 = 1{\cdot}03, \qquad \frac{1}{C_1\,\omega_\beta} = \underline{1{\cdot}03\,R_1}$$

$$\frac{1}{L_2\,C_1\,\omega_\beta^{\,2}} = K_{12}^{\,2} = 0{\cdot}73, \qquad L_2\,\omega_\beta = \frac{1{\cdot}03\,R_1}{0{\cdot}73} = \underline{1{\cdot}41\,R_1}$$

$$\frac{1}{C_3\,L_2\,\omega_\beta^{\,2}} = K_{23}^{\,2} = 0{\cdot}4, \qquad \frac{1}{C_3\,\omega_\beta} = 0{\cdot}4\,L_2\,\omega_\beta = \underline{0{\cdot}564\,R_1}$$

$$\frac{1}{L_4\,C_3\,\omega_\beta^{\,2}} = K_{34}^{\,2} = 0{\cdot}52, \qquad L_4\,\omega_\beta = \frac{0{\cdot}564\,R_1}{0{\cdot}52} = \underline{1{\cdot}08\,R_1}$$

$$\frac{R_4}{L_4\,\omega_\beta} = d_4 = 0{\cdot}56, \qquad R_4 = {\cdot}56 \times 1{\cdot}08\,R_1 = \underline{0{\cdot}605\,R_1}$$

The last line is a check on the accuracy of the calculation which is well within that of the slide rule and curves used.

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(1) E. Green. Exact Amplitude Frequency Characteristics of Ladder Networks—Marconi Review, 1st Quarter 1953, No. 108, p. 25.

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BOOK REVIEW

The modern domestic television receiver, designed for optimum performance and reliability at the minimum first cost, is yet such a complicated assemblage of components, that an experienced radio service engineer whose locality suddenly becomes a television service area, might well be apprehensive of his new responsibilities. It is to such a one that this book,* and a second volume now in preparation, are addressed.

The book is logically divided into Chapters—each dealing with a section of the television receiver (" The Synchronised Time Base ", " Tube Circuits ", etc.) or, with a process (" Applying a signal ", " D.C. restoration ", etc.). Each subject is explained fully, with frequent reference to typical circuits in post-war receivers, and the differences and similarities between like circuits in receivers of different makes, are well and clearly explained.

In the reviewer's opinion, the book succeeds where others have failed, by not trying to be a service manual; the chief concern of the author is to teach his reader the basic principles underlying the circuits with which he will be confronted and to help him to use the fault-finding skill he has already acquired as a radio service engineer.

The author's wide practical experience is written deeply into his book, which, as a result, will undoubtedly be of considerable value to those who wish to know something of the techniques employed in the modern television receiver. The typescript is clear, and errors are few, and, for the most part, negligible. The style is eminently readable, with a tendency to repetition which may be attributed to the nature of the author's task—to teach a superficial familiarity with a brand of electronics without the conciseness of a mathematical treatment which is available to the writer of a textbook.

The publication of Volume 2 (Receiver and Power Supply Circuits) is awaited with interest.

* *Television Receiver Servicing.* Volume 1.: Time Base Circuits. by E. A. W. Spreadbury, M.Brit.I.R.E. Published by Iliffe & Sons, Ltd., London. Price 21/-.

A POLARIZED MIRROR DUPLEXER FOR USE WITH A CIRCULARLY POLARIZED LENS AERIAL

By J. F. Ramsay, M.A., M.I.E.E., and W. F. Gunn, B.Sc.

A polarized mirror or closely spaced grating provides a simple means of separating two waves whose polarization are at right angles. The wave of one polarization is reflected off the mirror, the other transmitted through it. The waves are thereby separated.

A simple " primary feed " duplexer embodying this principle is described. It is used with a circularly polarized lens aerial in the 8–9 *mm. band. A crosstalk factor for the duplexer alone of* −75 *db is obtainable over a narrow-band. Associated with a dielectric lens aerial and metal quarter-wave plate each of* 2 *feet dia., a crosstalk factor of* −55 *db was obtained over a* 2% *bandwidth while* −75 *db could still be realized over a moveable narrow-band. Other performance features of the arrangements are discussed.*

Introduction

In a previous paper[1] it was shown how the use of radiated circular polarization allowed loss free common transmission and reception on a single aerial. The techniques described included systems using aperture circularizers for producing radiated circular polarization, the same devices " decircularizing " the received circular polarization, and waveguide " duplexers " for separating the two orthogonally polarized waves associated with transmission and reception. Waveguide duplexers however tend to have a limiting performance on crosstalk between transmitter and receiver[2]: they require accurate construction, careful setting to work and freedom from temperature variations. In particular their realization is not a short term development.

With a view to finding a quicker means of obtaining an effective duplexer, consideration was given to semi-optical methods, i.e., to " free-space " as contrasted with " waveguide " systems. It was observed that in waveguide duplexers an inherent feature is the open or disguised use of polarization filtering, for separating the paths of the two orthogonal T and R waves. One of the simplest semi-optical components is the polarized mirror, i.e., a grating or grid of closely spaced wires; such a device would be a polarization filter for two suitably oriented orthogonally polarized waves incident upon it. The wave having its polarization parallel to the wires would be reflected from the grating, that having the orthogonal polarization would be transmitted through the grating.

In the modern constrained waveguide polarized mirrors have been used in the form of transverse grids, being replaced sometimes by a single resonant slot. A duplexing system was therefore devised using unguided radiation, i.e., a " free-space " arrangement, and Fig. 1 is a diagram showing how a polarized mirror provides duplexing in a circularly polarized radar system.

Schematic Arrangement

A radar transmitter, Tx, is connected to a horizontally polarized primary feed horn, H, the aperture of which is at the focus of the lens, L. Between the horn and lens is inserted, at 45°, a polarized mirror, M, consisting of a closely spaced grating of vertical wires. Since the feed horn is horizontally polarized, the transmitted energy passes through the vertically polarized mirror (save for some slight reflection arising from the finite thickness of the wires). This " primary " radiation illuminates the lens, L, which converts the incident spherical wave into a transmitted plane wave at the aperture of the lens. The polarization is still horizontal.

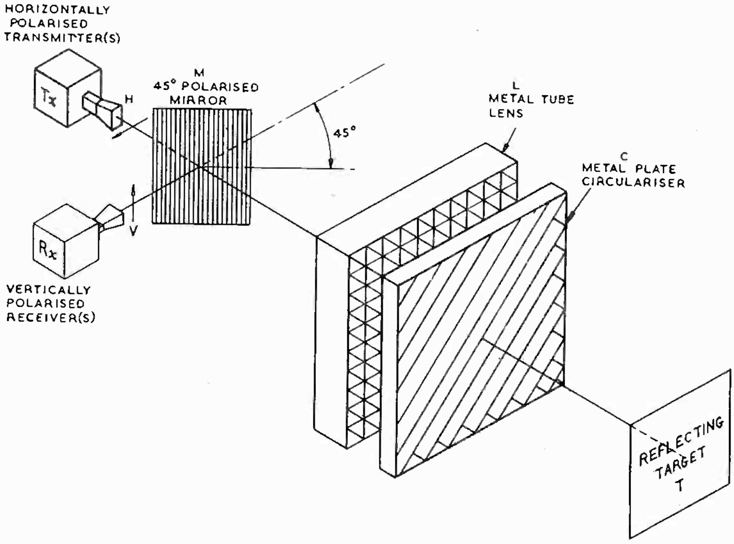

FIG. 1

Schematic arrangement of the C.W. Duplexer.

At the aperture of the lens is a metal plate circularizer, C, or " quarter-wave " plate, which converts the horizontally polarized wave from the lens into a circularly polarized wave radiated to space. The action and design of the quarter-wave plate is described in the paper on circular polarization.(1)

The circularly polarized wave falling on a plane symmetrical radar target, T, is reflected with the screw sense of the circularity reversed. On passing back through the quarter-wave plate, C, the echo wave is converted into a vertically polarized wave, since the circularity had been reversed by the target.

Between the circularizer, C, and the lens, L, the received wave is plane and vertically polarized. The lens is designed to transmit any polarization. Hence the echo, if allowed to, would focus at the aperture of the horn, H. Due to the presence of the 45° vertically polarized mirror, however, the echo is completely reflected from this mirror, and comes to a focus at the vertically polarized primary horn, V, whence it proceeds into the receiver, Rx.

Thus the polarized mirror acts as a " duplexer," in that it separates the received echo from the transmitted signal when used in conjunction with the circularly polarizing quarter-wave plate.

Since the lens and feed horns are standard components, the only additional components required to effect the duplexing are the circularizer and polarized mirror. In order to secure low crosstalk from the transmitter to the receiver it would be expected that some screening is also necessary. The practical construction of a suitable arrangement is, therefore, described in the next section.

Practical Construction of the Duplexer

The make-up of the duplexer is shown in Fig. 2, which is a photograph of a polarized mirror duplexer for a lens having a focal-length-to-aperture ratio or F-number of 1·5. This factor decides the size of the T and R primary feed horns, which are standard pyramidal horns designed to give a 10 db illumination taper over the lens aperture—a circular lens being assumed. The horns are, of course, substantially linearly polarized.

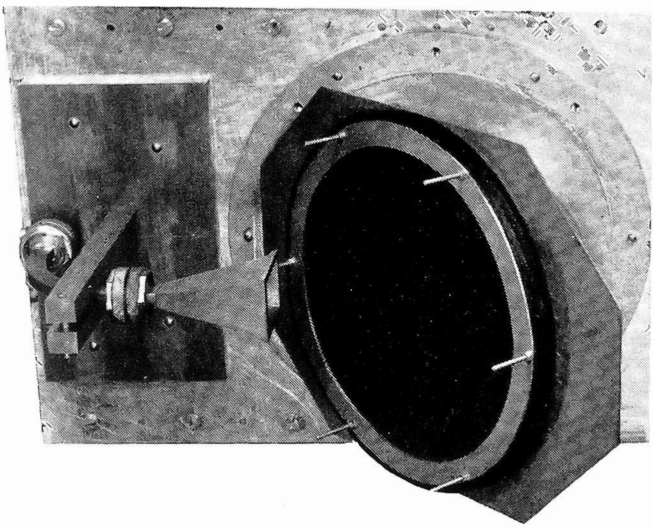

Fig. 2

Polarized Mirror Duplexer showing Transmitter Feed Horn below mirror, Receiver Feed Horn is enclosed behind mirror.

The polarized mirror is a grating of wires embedded in expanded ebonite (Fig. 3) and is set at 45° for convenience. The wires are No. 34 S.W.G. tinned copper (·0092 inch dia.) and are spaced ·040 inch, i.e., c.$\lambda/8 \cdot 5$. The expanded dielectric is marketed under the trade name of "Onazote," by the Expanded Rubber Co., Ltd. Slots of the same order as the wire diameter are milled in the sheet and the wires carefully pulled into them, so that they lie uniformly in a plane. The sheet thickness is $\frac{3}{8}$ inch and the slot depth $\frac{1}{16}$ inch.

The mirror is held on a 45° brass plate by means of a ring flange, the periphery of the mirror being surrounded by a ring or "washer" of loaded gutta-percha to prevent leakage at the edge of the mirror.

The 45° brass plate is brazed to a truncated screening cylinder of $6\frac{1}{2}$ inches dia., the latter being flanged to a supporting backplate containing one of the feed horns—in this case the receiver horn, hidden on the photograph but similar to the transmitting horn shown.

The same back-plate carried a waveguide run and associated brackets for the transmitter feed horn, which sits in free space facing the outside of the mirror.

The transmitter horn is jig assembled in order to place its aperture in exactly the right place, viz., at the image of the receiver horn.

The distances of the feed horns from the mirror have been made somewhat arbitrary in the model. If the horns are too close there is a risk of crosstalk both from multiple reflections in the cylinder and from cross-polarization in the diffracted field. If the horns are widely spaced there is a risk of mirror edge effects and reflections inside the truncated cylinder, apart from clumsiness of size.

A compromise distance of 2 inches was found suitable.

Performance of the Duplexer Unit

The most important performance figure for the duplexer is the leakage from the transmitter to the receiver, i.e., the "crosstalk." Using the twin horn and mirror unit alone, i.e., the "primary feed duplexer," a crosstalk level of -75 db was obtained at λ 8·80 mm. The performance, however, was limited to a relatively narrow band and was secured by the following adjustments:—

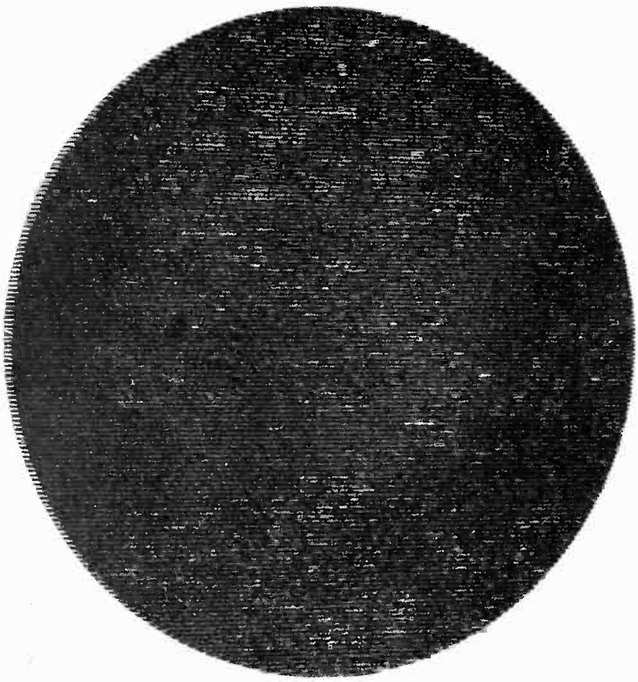

FIG. 3
Polarized Mirror of Wires embedded in Expanded Ebonite.

(1) The polarization of the transmitter feed horn was geometrically aligned with the polarization of the mirror.

(2) The receiver feed horn was then slightly rotated about its axis to obtain the level of minimum crosstalk.

(3) The transmitter feed horn was mounted on a sliding brass plate (see Fig. 2) so that the distance of this horn could be varied over two wavelengths, relative to the mirror, without any other change of co-ordinates.

In these adjustments the polarization and position of the mirror were taken as fixed references.

The low levels attained were measured by a high gain selective superheterodyne receiver utilizing an A.F.C. loop. The crosstalk level of the primary feed duplexer never went below about 65 db for 8·65-8·8 mm. and was due to radiation leaking through the mirror into the cylinder.

With the primary feed duplexer adjusted for minimum crosstalk, the match for the receiving horn looking through the polarized mirror was V.S.W.R.$=$0·85, while for the transmitter horn looking at the 45° mirror the V.S.W.R.$=$0·87, including a short waveguide run with one H-bend. In the latter case the duplexing mirror has no effect on the match. In the receiving case, however, reflections from the polarized mirror and shield affect the match slightly.

Application of the Polarized Mirror Duplexer to a Circularly Polarized Lens Aerial

The primary feed duplexer was built for use with any lens having a focal length/aperture ratio of 1·5, and has been associated with a dielectric (polyethylene) lens of 2 feet diameter. At the aperture of this lens is a "circularizer" or quarter-wave plate made of metal strips embedded in expanded ebonite, to effect the circular polarization.[1]

A photograph of the aerial is shown in Fig. 4 where the feed duplexer will be seen mounted on a rear plate attached to a lens ring. Tie-rods connect this back ring to the front ring which holds the lens, and to which the circularizer, also in a ring, is attached. The feed and objective arrangements are thus rigidly inter-connected.

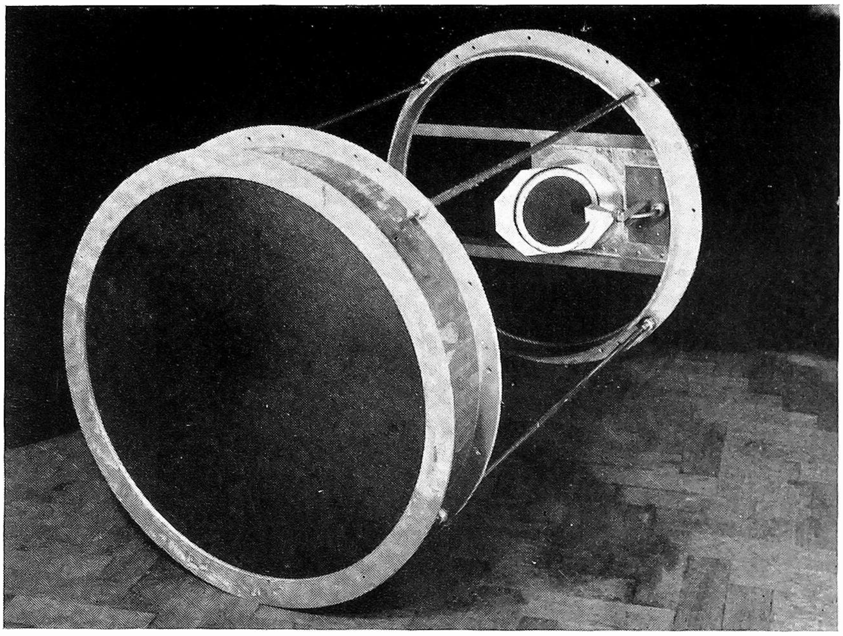

FIG. 4
Common T and R Aerial Assembly. Lens is behind Quarter-Wave Plate at Aperture.

No additional screening is provided or has been found necessary other than a substantial wooden crate which at once houses and screens the aerial, and is suitable for transporting it.

The lens (Fig. 5) is an experimental wide-angle dielectric lens with the following features:—

Illuminated with a 10 db taper a pencil beam of 1·35° with − 23 db sidelobes is obtainable. The gain is 42 db above an isotropic source. The profile is sphero-ellipsoidal, with steps on the front (ellipsoidal) surface.

The quarter-wave plate (2 feet dia.) consists of metal strips ·0092 inch thick, ·35 inch wide, spaced $\frac{9}{32}$ inch in expanded ebonite. The strips were positioned in the foamed ebonite (nominally 2 inches thick) by milling ·35 inch deep slots in one face to take the metal strips. One edge of each strip is then flush with the face of one side of the ebonite.

This large circularizer was based on the design of a 6-inch diameter model which provided a circularity of better than 0·95 (Fig. 6).

The circularity of the 2 feet circularizer used with the lens was also better than 0·95. This was confirmed by the gain measurements on the aerial.

In order to reduce any crosstalk resulting from reflections from the lens and circularizer the latter components were tilted by c. 5°. The aerial is thus used in a scanning position of the feed, the beam being elevated 5° above the (horizontal) axis of the lens. There is negligible scan in azimuth.

Fig. 5

Polyethylene Lens (2 ft. dia.) viewed from the front.

The setting up of the aerial involved the following adjustments:—

(1) The objective system and duplexer were set up for minimum crosstalk; this required a slight trim in the angle of tilt of the duplexer.

(2) The radiation patterns of the complete assembly as set up in (1) were then measured (see below).

(3) The gains of the "transmitting aerial" and of the "receiving aerial" were also measured thus providing the radiation efficiency of the complete system.

The crosstalk level of —75 db. of the duplexer was attained for the complete assembly by careful adjustment. The level obtained over a very narrow band at the specified frequency. The highest level over a 2% frequency variation was —55 db.

Radiation Patterns and Gain

The radiation patterns for both transmitter and receiver aerials were obtained using vertically polarized incident radiation from a distant transmitter (i.e., the duplexer was used as a receiver in both cases); the assembly being turned through 90° to obtain the elevation patterns.

RADIATION PATTERN TABLE

		Beam Width (—6 db.)	Sidelobes (db.)
T	Elevation	1·35°	—19
R	Elevation	1·35°	—19·5
T	Azimuth	1·38°	—21
R	Azimuth	1·40°	—21·5

The gain of the receiver aerial was 42·0 db. and the gain of the transmitter aerial was 41·6 db.; the small difference in gain being due to a slight misalignment of the circularizer with respect to the polarized mirror.

Possible Methods of Improving the Bandwidth of the Crosstalk

The Duplexer Alone

An improved polarization filter, instead of the simple grating mirror would be likely to lead to lower crosstalk. A cut-off waveguide stack of plates arranged like a Venetian blind with a spacing of 0·080 inch was constructed later. Apart from causing less reflection of the transmitted polarization this mirror was more accurately made and very robust. Considerably less leakage of the correct polarization occurred with this mirror but any cross-polarized component present in the incident radiation was transmitted as with the wire grating. A further improvement in the duplexer could be made by preventing reflection of the leakage radiation through the mirror from the inside surface of the screening cylinder, e.g. by "darkflex" absorber.

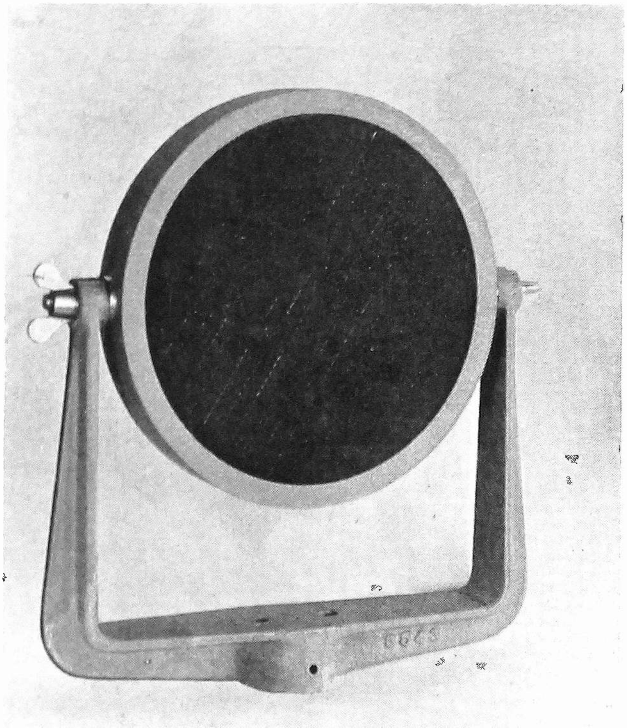

FIG. 6

Typical Circularizer, consisting of Metal Strips embedded on Expanded Ebonite.

The Duplexer and Objective

Since it was found that the presence of the objective narrowed the bandwidth of the system as compared with the duplexer alone, it was evident that reflections from the objective were the limiting factor. Reflectionless surfaces could always be achieved but with considerable difficulty. It should be noted that in reducing a reflection it is sufficient to minimize the cross-polarized component in the reflection.

Considering the sphero-ellipsoidal lens used, reflections from the back face will be focused to the conjugate focus and there will be very little cross-polarization present if there is none in the primary illumination.

Reflections from both faces of the circularizer will also be focused back to the same point. The component of cross-polarization in this reflection will be 6 db. down on the level of a similar specular reflection polarized and received parallel to the vanes of the circularizer.

Reflections from the front face of the lens will contain cross-polarization because it is ellipsoidal and stepped. Furthermore, these reflections will not be clearly focused and will spread into the region of the receiver. From these considerations it appears that the front surface reflections impose the limitations on the crosstalk bandwidth of this aerial.

Reasoning on these lines suggests that the sphero-ellipsoidal dielectric lens is not necessarily the best profiling to use. For example, in a sphero-ellipsoidal metal-tube lens the reflections from the front face would be divergent. With such a medium, however, there would be danger of a cross-polarization component existing in the back-surface reflections. An additional improvement could be achieved by reducing the number of steps on the front-face of the lens. However, cancellation of a reflection from any one surface can always be achieved by means of an equal path length loop from the transmitter to the receiver with the amplitude and phase in this loop adjusted accordingly.[3]

Conclusions

In a common aerial arrangement using circular polarization duplexing, the orthogonal T and R waves can be efficiently separated using a polarized mirror. The simplicity of the construction and the readily obtained crosstalk figure, as good as, if not better than, constrained duplexers taking much longer to develop, make the arrangement an attractive practical proposition. With further development the polarised mirror duplexer appears likely to be superior to waveguide duplexers in respect of both match and crosstalk.

Although not examined in this paper, the value of the single duplexer unit in catering for a plurality of transmitters and receivers operating with a single aerial is noteworthy. Such a requirement would necessitate with waveguide components a plurality of individual duplexers, producing both electrical and mechanical complication easily avoided by the use of the single polarized mirror.

The aerial described in this paper was built for the 8—9 mm. band. For experimental purposes a 3·2 cm. model can be visualized, but would be somewhat large. On the shorter millimetric waves, however, this semi-optical duplexer will be superior to a waveguide duplexer owing to the increasing difficulties in waveguide technique as the wavelength is reduced.

Acknowledgments

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