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MATERIALS USED IN RADIO AND ELECTRONIC ENGINEERING

A Survey by the Technical Committee of the Institution

5. MAGNETIC MATERIALS*

1. Introduction

This review summarizes the properties of magnetic materials used in radio and electronic engineering which are available in Great Britain.

With the development of electronic circuit techniques covering the whole radio frequency spectrum there is a growing demand for magnetic materials with special characteristics, calling for close manufacturing tolerances. A wide range of magnetic materials is available today and the main characteristics of the majority are given in this review.

The descriptive text deals only with some of the more important points and is complementary to the tables and references.

2. General

Magnetic materials may be generally divided into two groups, the magnetically hard which retain their magnetization, and the bigger and possibly more important group of magnetically soft materials which do not retain their magnetization. The two groups may be subdivided according to their use as shown in Table 1, from which it will be seen that the relatively new class of materials known as ferrites appear in both categories. Because of their increasing importance in radio-frequency applications, information on ferrites is given in a separate section.

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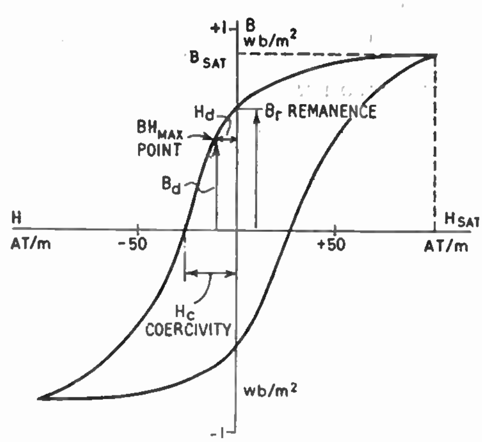
For magnetically hard materials the most important properties are usually (1), the ability of the magnet to retain its magnetization under the required operating conditions, and (2) the magnetic energy which can be stored in unit volume of the material.

In the magnetically soft group, the permeability saturation density, or specific resistance may be the most important parameter in a particular application.

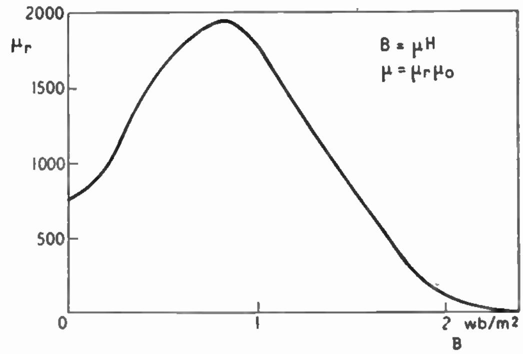
Table 1

The Categories of Magnetic Materials

Magnetically Hard Materials	Cast and sintered high grade magnets (Table 3)
	Rolled and forged low grade magnets (Table 4)
	Ferrite ("ceramic") magnets (Table 8)
Magnetically Soft Materials	Low-permeability laminated iron and iron-silicon alloys for power frequencies (Table 5)
	High-permeability laminated nickel iron alloys for l.f. applications, etc. (Table 6)
	Powdered metal materials for r.f. applications (Table 7)
	Ferrite materials for r.f. and microwave applications (Table 9)



(*a*)



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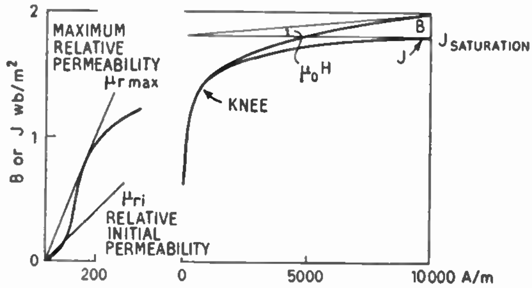

(*b*)

Fig. 1. Typical curves used in connection with magnetic materials.

(*a*) Hysteresis loop.

(*b*) Initial magnetization curve.

(*c*) Variation of relative permeability with *B*.

2.1. *Basic Parameters*

Terms used in this review are illustrated where possible in Fig. 1 (*a*), (*b*) and (*c*) and a summary of the units is given in Table 2. The terminology used is as defined in B.S.205 (see §9.1). It is important to remember the differences between British and American practice, e.g. the term "remanence" has entirely different interpretations. The units are given in the M.K.S. system with conversion factors to the C.G.S. e.m. practical system; the latter employs magnetic field quantities in C.G.S. e.m. units but current is given in amperes.

3. Magnetically Hard Materials (Tables 3, 4)

For a considerable period steel has been used for permanent magnets but many better materials have recently been developed. These are mainly alloys of iron with aluminium, nickel, cobalt, copper and other metals. The method of manufacture of these alloys influences their structure and is of extreme importance.

In general these alloys are too hard and brittle for forging or extensive machining. They are therefore cast or sintered into their approximate final shape, ground and finished, and then magnetized. Some of these alloys are *isotropic* and some *anisotropic*. The isotropic materials can be magnetized along any axis with similar results, whereas anisotropic materials have a desired axis of magnetization. The properties given in Table 3 are for that axis.

Where a high specific magnetic energy content is not absolutely essential, i.e. where volume and weight are not of prime importance, rolled or forged steel magnets are sometimes preferred. These are generally made of alloys of iron with additions of carbon and chromium, cobalt or tungsten. The properties of these comparatively machinable magnetic steels are given in Table 4 and can be readily compared with the properties of the cast or sintered high grade magnets given in Table 3.

4. Magnetically Soft Materials (Tables 5, 6, 7)

Iron used for electromagnets should be well annealed and to obtain the maximum magnetic performance, this should take place after machining has been finished.

Silicon-iron sheet material for small power transformers and chokes may be divided into two main groups, hot-rolled non-oriented silicon iron and cold-rolled grain-oriented iron. The former is normally obtained in the form of laminations insulated on one side to reduce eddy current loss; the usual thickness of commercial supplies is 0·014 in. for 4% silicon irons and 0·020 in. for lower grades. No heat treatment of the lamination is necessary but bending should be avoided.

With grain-oriented material the best properties are only obtained in the direction of rolling and for this reason it is not particularly suitable for laminations, although it is sometimes used in this form. The material is usually slit into long lengths in the direction of rolling. It is then wound on to a rectangular mandrel, annealed, varnish bonded, and cut to form two "C"s. These "C" cores are usually assembled in pairs in a shell-core construction in high quality transformers. They may be operated at 1·7 Wb/m² at 50 c/s compared with the usual 1·1–1·2 Wb/m² for hot-rolled material. The sheet is insulated with a very thin phosphate coating enabling a space factor of 95 per cent. to be obtained for 0·014 in. material.

The highest attainable permeabilities are obtained with nickel-iron alloys. Saturation flux densities range from 0·6–1·2 Wb/m² depending on the proportion of iron. These alloys are carefully manufactured from the purest available raw materials, either by vacuum melting or by powder metallurgy, and rolled or forged to the required thickness. They may be oriented or non-oriented, the former generally giving the very high permeabilities and square loop *BH* curves required for magnetic amplifiers and storage systems. Orientation of nickel-iron alloys gives two preferred directions of magnetization at 90 deg. to each other and so enables the material to be used in the form of laminations. With the high permeabilities obtainable with these alloys the magnetic

Table 2

Magnetic Quantities

Quantity	Symbol	Rationalized M.K.S System: Expression	Rationalized M.K.S System: Unit	C.G.S. E.M. Practical System: Expression	C.G.S. E.M. Practical System: Unit	Conversion Factor
Magneto-Motive Force	F	IN	Ampere turns (AT)	$4\pi/10IN$	Gilbert	1AT $=4\pi.10^{-1}$ Gilbert
Flux	Φ	BA	Webers(Wb)	BA	Maxwell	1Wb $=10^8$ Maxwells
Reluctance	S	F/Φ	AT/wb	F/Φ	Gilbert/ Maxwell	1AT/Wb $=4\pi.10^{-9}$c.g.s.
Flux Density	B	$\mu_r \mu_0 H$	Wb/m²	μH	Gauss	1Wb/m² $=10^4$ Gauss
Magnetizing Force	H	F/l	AT/m	F/l	Oersted	1AT/m $=4\pi.10^{-3}$ Oersted
Field Energy Density	w	$\frac{1}{2}BH$	Joules/m³	$\frac{1}{2}BH/4\pi$	Erg/cm³	1Joules/m³ $=$ $=10$ ergs/cm³
Intensity of Magnetization	J	$B-\mu_0 H$	Wb/m²	$(B-H)4\pi$	Gauss/4π	
Absolute Permeability of Free Space	μ_0	$4\pi.10^{-7}$	Henries/m	1	—	

circuit is very seriously affected by air-gaps and for this reason the best materials should be used in the form of toroidal cores or special stampings built up into cores in an overlapped manner to minimize the effective air-gap.

The alloys are extremely stress-sensitive and all fabrication must be done before final heat treatment. For these materials the makers generally recommend that after fabrication all oil and dirt should be removed and the parts heated in a pure dry hydrogen atmosphere to 1100°C for four hours followed by furnace cooling to room temperature. Manufacturers are generally willing to undertake the heat treatment of finished parts made of their alloys.

High saturation materials are available in both solid and laminated forms. Heat treatment after fabrication consists of heating in hydrogen to approximately 790° C for four hours then cooling slowly. They are expensive due to the high cobalt content (Table 6).

Dust-iron cores for use at radio frequencies are made up from minute particles of very pure magnetic material bound together with an insulating bond to reduce the eddy-current loss but the permeability is low and the effective iron cross-section is low due to the large number of "air" gaps. The cores may be obtained in pure iron or nickel-iron powders and have to be purchased as finished parts from the makers, who publish extensive tables of sizes and types available. (Table 7.)

Cores made with flaked instead of powdered iron have the advantage of a higher permeability due to the smaller effective air gaps in two preferred directions at 90 deg. to each other. These materials are intended mainly for use at power frequencies although modifications to the flake insulation processes have made them suitable for suppression inductors up to about 20 Mc/s. (Table 5.)

5. Ferrites—General

Ferrite is the name commonly given to a class of materials containing iron in a chemically-combined form and characterized by a crystalline structure; however, the term also refers to a form of metallic (α) iron containing a very small percentage of carbon.

The general formula for a ferrite is $MO.Fe_2O_3$ where M is a bivalent metal, e.g. Cu, Fe, Mg, Mn, Ni or Zn; $FeO.Fe_2O_3$ (or Fe_3O_4) is lodestone or magnetite.

A further group has the basic formula $MO.6(Fe_2O_3)$ where M is commonly Ba, Pb or Sn.

Both groups have a crystalline structure similar to the ceramic group of materials (hence the term "ceramic" magnet) and exhibit similar mechanical properties. Commercial materials are usually a mixture of various ferrites.

5.1. *Ferrites—Magnetically Hard* (Table 8)

Typical magnetically-hard ferrites are Fe_3O_4, $CoFe_2O_4$ and $BaFe_{12}O_{19}$. They have a high coercitivity and appreciably lower value of $(BH)_{max}$ compared with conventional magnets. They are very stable and difficult to demagnetize either by external fields or by mechanical shock. They do, however, have a negative temperature coefficient of remanence of 0·2% per °C (approximately 10 times that of metal magnets). The specific resistance is high which enables them to be used, for example, as the polarized core of an inductor carrying alternating current.

5.2. *Ferrites—Magnetically Soft* (Table 9)

The useful properties of these ferrites, which include $CuO.Fe_2O_3$, mixtures of $MnO.Fe_2O_3$, $ZnO.Fe_2O_3$, $NiO.Fe_2O_3$ and others, are the very high specific resistance, low coercive force and high initial permeability obtainable.

These properties make them particularly suited for the manufacture of small coil cores of all shapes and uses, Q's of up to 600 being obtained in a 1-in. cube. Disadvantages are that the Curie point can be low (120° C) and that the temperature coefficient of permeability can be as high as 1% per °C.

Apart from various types of cores for r.f. inductances, materials can be made having a ratio of B_R to B_{max} of greater than 0·9, and a flux density at $H_c/2 > 80\%$ of B_{sat}, or in other words with a nearly rectangular hysteresis loop. This non-linear characteristic can be utilized in computer switching and memory circuits, magnetic amplifiers and modulators.

The property of gyro-magnetism has enabled apparatus having non-reciprocal properties to be manufactured from magnetically-soft ferrites for use at microwave frequencies.

Table 3

CAST AND SINTERED HIGH GRADE MAGNETS

This range of permanent magnetic materials has been developed for particular purposes and their application depends on the conditions under which they are used. The properties have been expressed in M.K.S. units throughout. Their cost roughly corresponds to the magnetic performance except in the case of sintered magnetic materials.

(*a*) Anisotropic Materials

Material Trade Name	$(BH)_{max}$ J/m³	B_r Wb/m²	H_c AT/m	B_d	H_d	B_{sat}	H_{sat}	Approximate Composition	Uses in Radio Industry	Machinability	Manufacturer
	$\times 10^4$		$\times 10^4$		$\times 10^4$		$\times 10^5$				
Columax	5–6·35	1·34	6·2–6·7	—	—	1·7	2·4	8Al 14Ni 24Co 3Cu	Used only where the highest magnetic performance is required for special purposes.	For practical purposes these can be ground. The materials are cast as closely as possible to the finished dimension and ground only where essential Approximate tolerances are 0·040 ins. on small castings and about 2% on large castings, i.e., over 4 ins.	P.M.A.
Ticonal GX	6·0	1·35	5·74	1·2	5·0	1·7	2·4				Mullard
Alcomax III SC	4·6	1·3	5·6	1·03	4·45	1·7	2·4	8Al 13Ni 24Co 3Cu 1Nb	High grade materials with performance slightly lower than that of Columax and Ticonal GX. They represent an excellent compromise between performance and cost.		P.M.A.
Ticonal L	4·3	1·35	4·6	1·2	3·6	1·7	2·4	7Al 14·5Ni 19·5Co 1·5Cu			Mullard
Ticonal G	4·18	1·29	5·1	1·03	4·06	1·7	2·4	8Al 14Ni 24Co 3Cu			Mullard
Alcomax II SC	4·1	1·28	4·78	1·05	3·9	1·7	2·4	8Al 11·7Ni 24Co 6Cu			P.M.A
Alcomax III	4·06	1·25	5·35	0·975	4·16	1·7	2·4	8Al 13Ni 24Co 3Cu 1Nb			P.M.A.
Ticonal C	4·0	1·267	5·4	0·97	4·12	1·65	2·4	8Al 14Ni 24Co 3Cu			Mullard
Alcomax II	3·74	1·24	4·65	1·00	3·74	1·7	2·4	8Al 11·7Ni 24Co 6Cu			P.M.A.
Ticonal S	3·34	1·107	4·94	0·883	3·75	1·7	2·4	8Al 14Ni 24Co 3Cu	High grade material but sintered for use where intricate shapes and accurate dimensions are required.		Mullard
Alcomax III (Sint.)	3·34	1·13	4·94	0·875	3·8	1·7	2·4	8Al 13Ni 24Co 3Cu 1Nb			P.M.A.
Alcomax IV SC	4·06	1·17	6·2	0·87	4·65	1·65	2·4	8Al 13Ni 24Co 3Cu 2Nb	High performance magnetic materials with a particularly high coercive force.		P.M.A.
Alcomax IV	3·43	1·12	5·98	0·8	4·30	1·65	2·4	8Al 13Ni 24Co 3Cu 2Nb			P.M.A.
Ticonal K	3·2	0·9	10·4	0·5	6·4	1·5	4·8	—			Mullard

(*b*) Isotropic Materials

Material Trade Name	$(BH)_{max}$ J/m³	B_r Wb/m²	H_c AT/m	B_d	H_d	B_{sat}	H_{sat}	Approximate Composition	Uses in Radio Industry	Machinability	Manufacturer
Reco 3A	1·44	0·745	5·6	0·428	3·36	1·35	2·4	—	Lower quality and priced material for applications where high performance is not essential.	As above	Mullard
Alnico (high B_r)	1·36	0·8	4·0	0·52	2·62	1·5	2·4	10Al 17Ni 12Co 6Cu			P.M.A.
Alnico	1·36	0·725	4·45	0·47	2·9	1·45	2·4	10Al 17Ni 12Co 6Cu			P.M.A.
Alnico (high H_c)	1·36	0·65	4·95	0·425	3·2	1·4	2·4	10Al 17Ni 12Co 6Cu			P.M.A.
Alnico (Sint.)	1·22	0·72	4·4	0·457	2·8	1·45	2·4	10Al 17Ni 12Co 6Cu			P.M.A.
Alnico (high B_r)	1·0	0·62	3·84	0·403	2·48	1·2	1·6	12Al 24Ni			P.M.A.
Alni	1·0	0·56	4·65	0·348	2·87	1·2	1·6	12Al 24Ni			P.M.A.
Alni (high H_c)	1·0	0·5	5·45	0·304	3·3	1·2	2·4	12Al 24Ni			P.M.A.

Table 4
ROLLED, FORGED AND MICROPOWDER LOW GRADE MAGNETS

Material Trade Name	$(BH)_{max}$ J/m³	B_r Wb/m²	H_c AT/m	B_d	H_d	B_{sat}	H_{sat}	Approximate Composition	Uses in Radio Industry	Machinability	Manufacturer
	$\times 10^4$		$\times 10^4$		$\times 10^4$		$\times 10^4$				
35% Cobalt steel	0·76	0·9	2·0	0·593	1·28	1·65	12·0	—	Steel magnets are still used for many applications but are being replaced by cast and sintered magnets given in Table 2. The chief reason for their continued existence is their mechanical properties	They can be machined drilled and tapped.	P.M.A.
15% Cobalt steel	0·495	0·82	1·44	0·525	0·94	1·55	12·0	—			
9% Cobalt steel	0·4	0·78	1·28	0·5	0·8	1·5	12·0	—			
6% Cobalt steel	0·353	0·75	1·16	0·468	0·75	1·45	12·0	—			
3% Cobalt steel	0·27	0·72	1·04	0·422	0·65	1·4	12·0	—			
6% Tungsten steel	0·24	1·05	0·52	0·698	0·345	1·7	4·8	—			
3% Chrome steel	0·225	0·98	0·56	0·62	0·366	1·6	4·8	—			
P.F.1. (HC)	0·514	0·4	0·398	0·23	0·00223	—	—	—	Micropowder magnets used where an extremely small sized magnet is needed.	—	Salford Electrical Instruments Ltd.
P.F.1. (HR)	0·796	0·7	0·263	0·46	0·0017	—	—	—			
P.F.2. (HC)	0·81	0·52	0·50	0·29	0·0028	—	—	—			
P.F.2. (HR)	1·2	0·90	0·279	0·61	0·00198	—	—	—			

Table 8
FERRITE MAGNETS

Material Trade Name	$(BH)_{max}$	B_r	H_c	B_d	H_d	B_{sat}	H_{sat}	Approximate Composition	Uses in the Radio Industry	Machinability	Manufacturer
	$\times 10^4$		$\times 10^5$		$\times 10^5$		$\times 10^6$				
Magnadur 1	0·76	0·2	1·4	0·095	0·8	1·78	1·12	$BaFe_{12}O_{19}$	Television focusing magnets where their high coercive force and non-conductive properties make them particularly suitable for use in the high stray fields. They are beginning to be used in loudspeakers and other applications.	Pressed and fired as closely as possible to finished dimensions. Ground only where essential.	Mullard Ltd.
Caslox II	0·76	0·2	1·4	0·095	0·8	1·78	1·12				Plessey Co. Ltd.
Feroba	0·72	0·21	1·35	0·105	0·675	1·80	1·12				P.M.A.
Gecolite B	0·69	0·20	1·20	0·108	0·636	1·78	1·12				G.E.C.
Magnadur 2	2·0	0·36	1·12	0·22	0·91	1·68	1·12				Mullard Ltd.
Magnadur 3	1·8	0·3	1·52	1·16	1·12	1·84	1·12				Mullard Ltd.
Caslox IV	2·0	0·37	1·27	0·21	0·95	1·68	1·12		Shift correction and centring magnets.	Pressed to required dimensions.	Plessey Co. Ltd.
Caslox III	0·32	0·14	0·88	0·065	0·48	1·70	1·12				Plessey Co. Ltd.

Table 5

PROPERTIES OF MODERATE PERMEABILITY SOFT MAGNETIC MATERIALS

Material Trade Name	Composition and Treatment	Permeability μ_{ri}	Permeability μ_r max	Saturation B sat Wb/m²	B_r Wb/m²	H_c AT/m	Resistivity microhm-cm	Density g/cm³	Remarks	Manufacturer
		$\times 10^3$	$\times 10^3$			$\times 10$				
Dynamo Magnet Iron	Low carbon and Alloy Cast Steel	2·0	3·0 H=480-560	2·13			10	7·88	D.C. dynamo and motor frames.	Edgar Allen & Co. Ltd., Sheffield.
Super Hyperm	Iron, 0·05% C, 0·05% Si, 0·05% Mn, 0·01% S, 0·015%P.	0·25	4·0 at H=136	2·15	0·85	8·0	10	7·88	D.C. electromagnets.	Lowmoor Alloy Steelworks Ltd.
	Annealed at 990°C. Aged for 4 wks.	8·0	10·0		1·15	5·2-6·4				
Unalloyed Sheet Steel	0·3% Si	0·25	5-6	2·12	1·2	5·4	14-16	7·82	Small machines; all applications where high flux density required at power frequency. Chokes for fluorescent lighting.	Joseph Sankey & Sons Ltd.
Lohys RTB No. 1										Richard Thomas & Baldwins Ltd.
Special Lohys RTB No. 2	0·85% Si	0·3	6·0	2·05		6·0	21	7·75	Meters, alternators.	Richard Thomas & Baldwins Ltd.
Stalloy RTB No. 4	4% Si	0·4-0·45	7-8	1·95	1·2	4·0	50-55	7·55	Power transformers.	Richard Thomas & Baldwins Ltd.
Crystalloy	3 - 3½% grain oriented Si-steel	1·5	60 H=12	2·0		1·2	45-47	7·65	High quality low loss transformers.	Joseph Sankey & Sons.
Alphasil										Richard Thomas & Baldwins Ltd.
Unisil										Steel Co. of Wales Ltd.
Caslam Grade 53	Laminated flake iron	0·18	0·9	2·1	0·6	28		7·0	Television line scan transformers and deflection yokes, also r.f. suppression.	Plessey Co. Ltd.
Caslam Grade 70	Laminated flake iron	0·25	1·2	2·1	0·7	32		7·2	Small machines, chokes for fluorescent lighting at frequencies up to 2400 c/s.	Plessey Co. Ltd.

Table 6

HIGH PERMEABILITY SHEET MATERIALS

Material Trade Name	Composition and Treatment	Permeability μ_{ri}	Permeability $\mu_{r\max}$	Saturation B_{sat} Wb/m²	B_r Wb/m²	H_c AT/m	Resistivity microhm-cm	Density g/cm³	Remarks	Manufacturer
		$\times 10^3$	$\times 10^3$							
Super Permalloy C Mumetal	77% Ni, rest iron and additions	20-30	90	0·78	0·35*	2·4*	62	8·8	Wide frequency band transformers, magnetic shielding, chokes.	S. T. & C. Telegraph Construction & Maintenance Co. Ltd. (Metals Division)
Super-Mumetal	Ditto	50	200	0·78	0·35*	0·8	60	8·8		Ditto
Permalloy C Sanbold NA76	75% Ni, rest iron and additions	15-40	50-150	0·8	0·35*	2·4*	60	8·8		S. T. & C. Sanderson Bros.
Radiometal	50% Ni, rest iron and small additions	2	25	1·56	0·43*	12·8*	45	8·0	Polarised relays, transformers and chokes requiring higher operating flux density than Mumetal and Permalloy C.	Telegraph Construction & Maintenance Co. Ltd. (Metals Division)
Permalloy B Sanbold NA47	Ditto Ditto	2·0-4·0	15-40	1·6	0·45*	15·0*	55	8·3		S. T. & C. Sanderson Bros.
HCR Alloy	Cold rolled grain oriented nickel iron alloy	0·5-1·0	100	1·56-1·6	1·5	10·4	40	8·25	Square hysteresis loop material; saturable reactors, memory storage cores, special components.	Telegraph Construction & Maintenance Co. Ltd. (Metals Division)
Permalloy F	Ditto	0·4-1·0	200-400	1·4	1·33†	4·0†	26	8·4	Used only in form of spiral cores.	S. T. & C.
Rhometal	36% nickel and iron alloy	1·8	7·0	0·9	0·36*	25·4*	85	8·1	Magnetic cores for radar transformers, pulse transformers; high resistivity allows higher frequency use.	Telegraph Construction & Maintenance Co. Ltd. (Metals Division)
Permalloy D	Ditto	1·8-3·0	12·0-20·0	1·3	0·35	15·0	90	8·15		S. T. & C.
Permendur	49% Fe, 49% Co, 2% V	0·8	5·0	2·36	1·5	160·0	28	8·05	Telephone diaphragms, pole pieces, electrical machines in aircraft, ultrasonic transducers.	Telegraph Construction & Maintenance Co. Ltd. (Metals Division)
V-Permendur	Ditto	0·7-1·0	3·0-5·0	2·4	1·6‡	200	26	8·2		S.T. & C.

* $B_{max} = 0{\cdot}5$ Wb/m² † $B_{max} = 1{\cdot}4$ Wb/m² ‡ $B_{max} = 2{\cdot}0$ Wb/m²

Table 7

BONDED, SINTERED AND COMPRESSED IRON POWDER MATERIALS

Material Trade Mark	Type of Material	Initial Permeability μ_{ri}	Characteristics and Applications	Some Loss Data			Manufacturer
				Hysteresis Coeff. $\mu_i a$	Residual Coeff. $\mu_i c$	Eddy Current Loss Coeff. $\mu_i e$	
				$\times 10^3$	$\times 10^3$	$\times 10^6$	
Gecalloy PL	Reduced iron powder with ceramic insulant	100	Particle size and insulation can be adjusted to meet a variety of purposes in communication engng.	1·0	20	10	G.E.C.
Gecalloy III	Nickel-iron powder with ceramic insulant	12-130	Loading and filter coils. Low losses	0·04-0·045	1·5-6·5	0·03-2·1	G.E.C.
HiQOR	Molybdenum nickel-iron powder	26-125	Loading and filter coils	0·17-0·22	3-6	0·16-2·0	Telephone Manufacturing Co. Ltd.
HiQOR Carbonyl Iron Grade ME	Carbonyl iron	15	ditto	0·05	4	0·007	ditto
Mond Carbonyl Iron, Grade ME	Carbonyl iron with resin insulation	12	Cores for telecommunications, filters, etc. Low eddy current losses	0·1	2·5	0·002	Mond Nickel Co.
Mond Carbonyl Iron, Grade MC	Heat treated carbonyl iron with resin insulation.	20-45	Higher permeability than non-heat treated powder. Higher losses.	0·7	0·15	0·03	Mond Nickel Co.
Permalloy	Nickel-iron powder with insulant.	14-145	Loading and filter coils. Low losses.	0·15	1·5-5	0·1-3·7	S. T. & C.

Table 9

MAGNETICALLY SOFT FERRITES

Material Trade Name	Type of Material	Initial Permeability	Characteristics and Applications	Some loss data	Manufacturer
Ferroxcube A. Grades A1 to A4	Manganese-zinc ferrite	750–1200 $B_{sat} = 0{\cdot}34 - 0{\cdot}38\ \mathrm{Wb/m^2}$	High permeability and high resistivity. Low eddy current and hysteresis losses.	Loss factor $\frac{\tan\delta}{\mu_i}$ at 250 kc/s $18\text{–}35 \times 10^{-6}$ Hysteresis coefficient $C_h\ 4{\cdot}2 - 9{\cdot}2 \times 10^{-6}$ Resistivity $\rho > 20\ \Omega$-cm	Mullard Ltd.
Ferramic NW1 Ferramic NW8	Sintered manganese-zinc ferrite Sintered modified nickel-zinc ferrite	800 850	For use in high Q coils and transformers in the frequency range 1 kc/s to Mc/s 1, i.e., higher than power frequencies.	$\frac{\tan\delta}{\mu_i} = 10 \times 10^{-6}$ at 100 kc/s $\frac{\tan\delta}{\mu_i} = 50 \times 10^{-6}$ at 100 kc/s Resistivity $\rho \backsim 10^4\ \Omega$-cm	Plessey Co. Ltd.
Stanferite 1 Stanferite 2	Sintered manganese-zinc ferrite Sintered manganese-zinc ferrite	2000 to 3000 1100 to 1900 $B_{sat} = 0{\cdot}45\ \mathrm{Wb/m^2}$		$\frac{\tan\delta}{\mu_i} = 5 \times 10^{-6}$ at 100 kc/s and 13×10^{-6} at 250 kc/s $C_h = 1 \times 10^{-6}$ $\frac{\tan\delta}{\mu_i} = 2{\cdot}6 \times 10^{-6}$ at 100 kc/s and 6×10^{-6} at 250kc/s $C_h = 0{\cdot}7 \times 10^{-6}$	S.T. & C. Ltd.
Gecolite Grade P Gecolite Grade R	Manganese-zinc ferrite Manganese-zinc ferrite	1100–2400 790–1100		$\frac{\tan\delta}{\mu_i} = 63 \times 10^{-5}$ at 90 kc/s Hysteresis factor: $F_H = 0{\cdot}62 - 6{\cdot}5$ ohms/mH/kc/s for pot cores $\frac{\tan\delta}{\mu_i}$ = approx. 45×10^{-6}	Salford Electrical Instruments Ltd.
Ferroxcube B Grades B1 to B5	Nickel-zinc ferrite	18–650 $B_{sat} = 0{\cdot}3 - 0{\cdot}19\ \mathrm{Wb/m^2}$	Low eddy losses, curie point 130–550°C. Used in coils up to 50 Mc/s.	$\frac{\tan\delta}{\mu_i}$ varies with grade, from B1: 50×10^{-6} at 0·5 Mc/s to B5: 4200×10^{-6} at 50 Mc/s Resistivity: $10^5\ \Omega$-cm	Mullard Ltd.

Material Trade Name	Type of Material	Initial Permeability	Characteristics and Applications	Some loss data	Manufacturer
Ferramic NW6 Ferramic NW10	Sintered modified nickel-zinc ferrite	125	Used in coils operating at low flux density up to a frequency of 20 Mc/s.	$\frac{\tan\delta}{\mu_i} = 40\times10^{-6}$ at 1 Mc/s Resistivity: $10^7\Omega$-cm	Plessey Co. Ltd.
Stanferites 4, 5, 6, 7 & 8	Sintered nickel-zinc ferrite	16–650 $B_{sat}=0{\cdot}4$ to $0{\cdot}25$ Wb/m^2	Very high resistivity and low flux density. Extremely low eddy current losses. Used in coils over a very wide frequency range up to 150 Mc/s.	$\frac{\tan\delta}{\mu_i}$ varies with grade from Stanferite 4: 40×10^{-6} at 500 kc/s to Stanferite 8: 440×10^{-6} at 50 Mc/s	S.T. & C. Ltd.
Gecolite Grades K8, K6, K4, K2	Nickel-zinc ferrite	15–700	As for other nickel-zinc ferrites.	$\frac{\tan\delta}{\mu}$ varies with grade and frequency from 20×10^{-4} to 50×10^{-6}	Salford Electrical Instruments Ltd.
Ferramic R1	Sintered magnesium-manganese ferrite	50	Microwave ferrite facilitating the use of Faraday rotation and gyromagnetic resonance in gyrators, isolators, microwave switches, etc.		Plessey Co. Ltd.
Ferramic S1, S3, S4, S5	Sintered magnesium-manganese ferrite (different grades for different coercivities and speeds of operation)	$B_{max}=0{\cdot}15$–$0{\cdot}18$ Wb/m^2	Rectangular hysteresis loop, used in shift registers, logical and switching elements and matrix stores. Resistivity 10^5 Ω-cm	Squareness factor B_r/B_{max} approx. 0·95 Coercive force approx. 80 AT/m	Plessey Co. Ltd.
Ferroxcube D	Magnesium-manganese ferrite	$B_{max}=0{\cdot}15$–$0{\cdot}17$ Wb/m^2		Squareness factor B_r/B_{sat} 0·95	Mullard Ltd.
Stanferite 3	Mixed ferrite	$B_{sat}=0{\cdot}33$ Wb/m^2	Rectangular hysteresis loop used in switching	Squareness factor $B_r/B_{sat}>0{\cdot}9$ Switching time $\backsim 5\mu$sec. Coercive force=48 AT/m	S.T. & C. Ltd.
Stanferite 11		$B_{sat}=0{\cdot}21$ Wb/m^2	Rectangular hysteresis loop used in memory arrays	Squareness factor $B_r/B_{sat}>0{\cdot}9$ Switching time $\backsim 1{\cdot}5\mu$sec. Coercive force=136 AT/m	

6. Magnetic Recording Media

Magnetic materials for recording purposes should ideally have the following characteristics:

1. High coercivity.
2. High remanence.
3. Uniformity of output signal.
4. Minimum inherent noise and freedom from print-through (leakage of recorded signal from one layer to another).
5. High elastic limit of base material.
6. Freedom from curl—both inherent and during storage.
7. Flexibility.

There are many other factors which affect final performance and their importance will depend on the ultimate use of the material, e.g. whether it is for professional audio-frequency recording, instrument work, computers, or h.f. recording.

Steel wire will satisfy many of the criteria, but it has serious disadvantages, e.g. print-through, difficulty in handling and splicing. It has, however, one feature, namely its small volume for a given recording time, which makes it useful where economy in size is of major importance.

Most magnetic recording media are in a non-homogeneous tape form, which consists of a non-magnetic base, e.g. paper, cellulose acetate, cellulose nitrate, polyvinyl chloride or polyester film, coated with a basic oxide of iron Fe_2O_3.

The preparation and treatment of the oxide determines the magnetic characteristics, whereas the backing material determines the physical characteristics. For a specific application, both these properties must be correctly chosen.

In the case of instrument tapes, pulse modulation recording is normally employed and high tape-speeds are encountered. The main requirement therefore is for a high tenacity backing material and a magnetic coating which is free from inconsistencies which cause "drop outs" (loss of signal which prevents the intelligence being read).

To increase the life of the recording media, non-contact heads are desirable and in instrument work, there is wide use of the magnetic-coated drum. Magnetic paint of various types is available for the preparation of drums.

Television signal recording requires high tape speeds (of the order of 30 ft. per second) or high differential speed between tape and head, and a high coercivity tape.

When an application, other than for sound recording, is being considered, it is essential to approach a tape manufacturer, as tapes not already covered by the existing range can be produced to a specification. A table comparing the properties of different tapes is not published here because of the absence of an absolute standard of comparison of data.

7. Acknowledgments

Thanks are due to Mr. A. J. Tyrrell (Member), Dr. K. Hoselitz and the many manufacturers who have collaborated in this review by providing and checking the information given in the Tables.

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8.2.1. General

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9. Specifications

9.1. *British Standards**

B.S.205:1943 Glossary of terms used in electrical engineering. 186 pp. 15s.

B.S.406:1931 Apparatus for workshop testing of permanent magnets. 33 pp., 7 plates. 4s. 6d.

Apparatus designed for finding such values as remanence and coercive force is described and working drawings are included. Relevant terms and quantities are described and defined. Instructions are given for methods of test for moving coil and rotating disc apparatus, also adjustments, corrections, and other points to be observed. The apparatus applies to horseshoe-type magnets not greater in size than 11 cm across outside of limbs and not less than 5 cm between limbs for cross-sectional areas of about 2 to 10 cm^2 and also to straight or slightly curved magnets 6 to 10 cm long and 2 to 10 cm^2 section.

B.S.601:1935 Steel sheets for transformers for power and lighting. 16 pp. 3s. 6d.

Two sheet sizes are specified with tolerances on width, length and thickness, and tests are prescribed for space factor, brittleness, losses for 3 grades of sheet and permeability. Test procedure and formulae are included in the appendices.

B.S.933:1941 Magnetic materials for use under combined a.c. and d.c. magnetization. 43 pp. 6s.

Supplements existing magnetic testing specifications primarily by providing for the important case of incremental magnetization; that is, magnetization by combined a.c. and d.c. The materials dealt with are silicon steel sheets of varying silicon content, and nickel-iron. Definitions are given, and requirements are stated for the materials. Appendices describe the methods of test which include in particular the measurement of incremental permeability and loss extending into the audio-frequency range.

B.S.1568:1953 Magnetic tape sound-recording and reproduction for programme interchange. 16 pp. 4s.

Specifies the features of recording on magnetic tape and of associated recording and reproducing equipment necessary to ensure the successful interchange of recordings for broadcasting or similar purposes. It applies only to single-track full-width recordings. The recording and reproducing characteristics are those adopted by the Comité Consulatif International Radiophonique (C.C.I.R.).

* Obtainable by purchase from British Standards Institution, British Standards House, 2 Park Street, London, W.1.

B.S.1617:1950 Mild steel castings of high magnetic permeability. 10 pp.

Provides for two grades of steel castings possessing special magnetic properties for electrical applications. For each grade the chemical compositions and mechanical properties are specified. Recommended values for magnetic properties are included. The process of manufacture, fettling and dressing, freedom from defects, testing facilities, branding and repairs to castings are dealt with. Recommendations for welding are given in an appendix.

(Note.—This standard is now only available as part of B.S.3100:1957. Steel castings for general engineering purposes. 40 pp. 7s. 6d.)

B.S.1637:1950 Memorandum on the M.K.S. system of electrical and magnetic units. 8 pp. 3s.

Explains the principles of the system of electrical and magnetic units founded on the metre, the kilogramme and the second, originally proposed by Professor Giorgi in 1901, and places on record the decisions of the International Electrotechnical Commission on the subject.

B.S.2454:1954 Methods for the determination of magnetic permeability of iron and steel bars, forgings and castings. 16 pp. 3s.

Deals with the determination of the normal magnetization curve connecting flux density and magnetizing field of iron and steel (other than permanent magnet materials) in form of bars, forgings and castings. Parts 1 to 4 give the basic requirements of the methods and an appendix gives full details of the recommended procedure for testing ring and bar specimens.

B.S.2857:1957 Nickel-iron transformer and choke laminations. 12 pp. 3s.

Lays down the requirements for non-oriented nickel-iron laminations either insulated or un-insulated, for use in transformers and chokes. It provides for four classes, A, B, C and D, each in three standard thicknesses, namely 0·015 in., 0·008 in., and 0·004 in. Limits are given for thicknesses, burrs, and space factor. Minimum magnetic permeability figures at 50 c/s are specified for each class and group, in each standard thickness, the laminations being divided into three groups according to the main magnetic path. Recommendation for methods of making audio-frequency tests also are given. Appendices give details of the standard method of determining space factor and a recommended method of determining permeability at 50 c/s.

B.S.——. "Silicon iron sheets, strips and laminations," under preparation.

The issue of this standard will follow after revisionary work has been completed on B.S.601 with which it is linked.

9.2. *D.T.D. Series**

DTD.330 Soft iron sheets, strips, bars and tubes (suitable for electrical purposes). 2s. 6d.

9.3. *Defence Specifications*†

DEF-5192 High permeability, magnetically soft alloys. August 1956. 2s. 6d.

This specification relates to magnetic alloys intended for use at audio and radio frequencies. The material consists generally of about 75 per cent. to 80 per cent. nickel, and 12 per cent. to 16 per cent. iron with minor amounts of modifying elements, which may be one or more of copper, chromium, manganese or molybdenum. The material is rolled into strip of thickness 0·008 in. or less, and is supplied without annealing. It does not apply to finished products, either laminations or spiral cores.

9.4. *R.C.S.S. List of Standards*‡

RCL.191 Transformer and choke laminations. October 1952.

RCL.193 Cores, magnetic, strip wound (rectangular and toroidal). March 1952.

10. List of Manufacturers of Magnetic Materials

Edgar Allen & Co. Ltd.,
Imperial Steel Works,
Sheffield.

The Mond Nickel Company Ltd.,
Thames House, Millbank.
London, S.W.1.

Mullard Limited,
Component Group,
Mullard House, Torrington Place,
London, W.C.1.

Lowmoor Alloy Steel Works Ltd.,
Low Moor,
Bradford.

Permanent Magnet Association,
301 Glossop Road,
Sheffield, 10.

The Plessey Company Ltd.,
Woodburcote Way,
Towcester, Northants.

Salford Electrical Instruments Ltd.,
(Components Group),
School Street, Hazel Grove,
Stockport, Cheshire.

Sanderson Bros. Ltd.,
Tyne Bridge Works,
Gateshead-on-Tyne.

Standard Telephones & Cables Ltd.,
Magnetic Materials Dept.,
North Woolwich, London, E.16.

The Steel Company of Wales Ltd.,
Margam House, St. James's Square,
London, S.W.1.

Joseph Sankey & Sons Ltd.,
Bankfield Works,
Bilston, Staffs.

The Telegraph Construction & Maintenance Company Ltd.,
Metals Division,
Telcon Works, Manor Royal,
Crawley, Sussex.

Telephone Manufacturing Company Ltd.,
Components Division,
Sevenoaks Way, St. Mary Cray,
Orpington, Kent.

Richard Thomas & Baldwins Ltd.,
Midland Section,
Wilden Works,
Stourport-on-Severn, Worcs.

* Obtainable by purchase from Her Majesty's Stationery Office, P.O. Box 569, London, S.E.1.

† Obtainable free of charge from Radio Components Standardization Committee, Ministry of Supply, Castlewood House, 77-91 New Oxford Street, London, W.C.1. (Restricted to Government Service Departments and contractors.)

ELECTRONIC SECTOR SCANNING*

The Rapid Swinging of an Acoustic Beam Across a Sector by Electronic Means and Its Application to Echo-Ranging Systems

by

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SUMMARY

A system of scanning, or swinging, the beam of an acoustic receiver by electronic means, while the transducer itself remains stationary, is described. The scanning may be very rapid, so that in a pulse echo-ranging system (e.g. asdic) the whole scanned sector may be covered within the duration of a pulse, giving effectively simultaneous reception over a sector which is shown to be approximately n times the 3-db width of the beam itself, n being the number of sections into which the array is divided. The more critical details of the design are discussed, and experimental results are given to confirm the theoretical analysis.

LIST OF SYMBOLS

n = number of sections into which the array is divided.

d = distance between centres of the sections of the array.

l = overall length of array.

θ = angle of any direction relative to the normal to the line or face of the array.

φ = phase difference between outputs of two adjacent sections of array.

φ_T = overall phase-shift in delay line.

t_1 = time delay between arrival of wave-front at adjacent sections of array.

t_2 = time delay per section of delay line.

k = number where $2k+1=n$.

$\varkappa$ = rate of change of frequency with time.

r = integer.

λ = wavelength.

q = angular frequency of signal.

p, s, are angular frequencies of local oscillators.

ω = difference frequency.

T_s = sweep period.

T_p = pulse duration.

$\Delta f = f_2 - f_1$ = extent of frequency sweep on delay line.

f_c = frequency of carrier of signal.

Part 1: DESCRIPTION OF THE SCANNING SYSTEM

1. Introduction

It is well-known that acoustic‡ beams are formed by electro-acoustic transducers of dimensions larger than the wavelength of the radiation, and that if such a transducer is divided into sections along a particular axis and the sections are connected together electrically through phase shift networks, then the beam may be deflected from its normal axis. This is fully explained elsewhere[1] and need not be discussed in detail here.

The word "beam" suggests a *transmission* of

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‡ The terms "acoustic" and "sound" are used even where the frequency is such that there is no question of audibility.

energy along directions restricted to a certain solid sector; but it is frequently used in connection with the *reception* of energy, in the sense that the electrical output of the transducer is large only for directions of reception restricted to a certain solid sector. "Beam" is used in the latter sense in this paper. "Directional pattern" (or "response") is the quantitative relationship between the electrical output amplitude of the transducer and the direction of received signals, and the "beamwidth" in any plane is usually taken as the angular width of this pattern in the plane concerned between the points at which the output amplitude is 3 db below that at the centre of the beam.

In principle there is no reason why a beam of any shape (produced by a transducer of any shape and/or non-uniform sensitivity) should not be deflected from its normal axis by electrical (as opposed to mechanical) means. In practice, however, it is generally only fan-beams—i.e. beams produced by a long, narrow transducer or "strip array"—which are so treated. The use of strip arrays is attractive theoretically for their simplicity; but in practice, too, they are useful, and when the strip has its long axis horizontal, so that the beam is wide vertically and narrow horizontally, the application to high-definition echo-ranging equipment is particularly attractive. A narrow horizontal beamwidth, coupled with the use of a short-pulse transmission, enables a target to be delineated in plan view with comparative accuracy and detail. Most radar systems do, in fact, operate like this, although underwater acoustic echo-ranging systems ("Asdic"), as used in fish-shoal detection, do not at present have fan beams. In the case of an echo-sounder, which is an acoustic echo-ranging system with the beam axis vertical, the use of a narrow beam to give high transverse resolution, coupled with beam swinging in the transverse direction, could lead to a three-dimensional picture of the sea-bottom configuration, showing the details of ridges, etc., which would be very much more useful than the present two-dimensional record. But if, to explore a sector of any size, the whole array has to be mechanically swung from one position to another, as at present, the process of search may become too slow to be really useful, and the mechanical difficulties may be great.

With strip arrays, beam deflection is easily carried out electrically by means of delay lines[1b] to which the sections of the array are connected at appropriate intervals. Then at each output of each delay-line there is obtained what is virtually the output from a beam whose axis is deflected from the normal position by an amount proportional to the phase-shift of the delay-line. Thus if a series of delay-lines is used with phase-shift in the rth line equal to r times that in the first line, a series of deflected beams is virtually obtained with a uniform progression of deflection. The negative deflections are obtained at one end of the lines, and positive deflections at the other. In this way a sector of many times the width of the beam may be searched without mechanical rotation; but the way in which the multiplicity of delay-line outputs is displayed on, say, a cathode-ray tube screen is very complex. A multiway sampling switch has to be devised unless each output is displayed on a separate screen or a multi-beam cathode-ray tube is used. In a pulse echo-ranging system this switch has to sample all outputs at least once during one pulse duration if no information is to be lost. Moreover, each delay-line output has to have its separate high-gain a.g.c. amplifier, and apart from the complexity of this, there is difficulty in keeping all outputs at the same gain-level at the display. But if a sector n times the beamwidth is scanned in this way, n times as many echo pulses are received from any target in a given time as are received with mechanical scanning of the beam. This gives an improved rate of receipt of information in the operational sense, and if the transmitted power per unit solid angle is maintained constant, the detection threshold will, ideally, correspond to a signal-to-noise ratio lower by $5 \log_{10} n$ db. This may represent an important improvement in performance.

The system[6, 7] to be described here provides all the facilities of the system described above, but with the use of only one delay-line, one single- or twin-beam cathode-ray tube, only one or two high-gain amplifiers, and with the great advantage of smooth and not discontinuous scanning of the search sector. The essential basic principle from which the system is developed is that if a signal is translated in frequency by means of a modulator, its phase-angle is preserved unaltered, except possibly in sign. Thus if a single delay-line is used with a strip array and its phase-shift increases linearly with frequency, then if the signals are translated to steadily increasing frequencies by modulation with a steadily changing frequency, the beam deflection produced also steadily increases. If the modulation frequency is swept through its cycle of change at least once in every pulse duration, then all directions in the search sector are sampled without serious loss of information. The display can be effected on standard cathode-ray tubes.

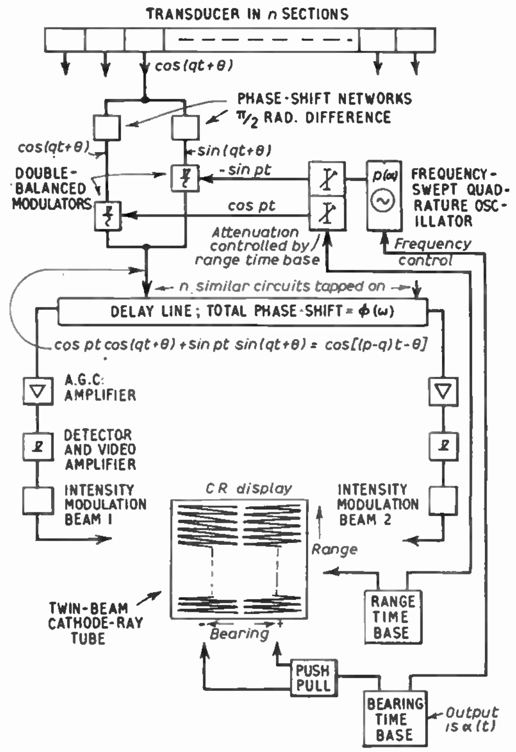

Fig. 1. Schematic of receiving system. (Low-pass single-modulation system.)

2. The Basic System

2.1. *Low-Pass Single-Modulation System*

2.1.1. Description

The complete receiving system in its primary form is shown in Fig. 1. But we must first of all discuss the deflection of the beam by a delay-line. If the number of sections, n, in the strip transducer is fairly large, and if the output signal is obtained from one end of a delay-line to which the transducer sections are connected directly (i.e. without modulation), the delay-line having a total phase-shift from one end to the other of φ_T rad, then the directional pattern or beam of the transducer has its main axis deflected from its normal direction by an angle $\varphi_T\lambda/2\pi l$ rad, where λ = wavelength, and l = length of transducer. This expression for the angle of deflection assumes that the angle is small; more strictly the deflection is arc $\sin(\varphi_T\lambda/2\pi l)$ rad. But when large deflections are produced with transducers divided into a relatively small number of sections, the beam becomes distorted and "diffraction secondaries" (or undesirable large secondary lobes in the directional pattern) are introduced; therefore, in practice, the angle of search over which the beam is deflected is limited. The determination of a suitable angle of search in relation to the number of transducer sections, or vice versa, is discussed later.

When $\varphi_T = \pi$, the angle of deflection is $\lambda/2l$. If the transducer is untapered, i.e. all sections have the same sensitivity, then the directional pattern at an angle $\lambda/2l$ from the direction of peak response has an amplitude of approximately two-thirds that of the peak, and it is very convenient, therefore, to think of a value of $\varphi_T = \pi$ as deflecting the beam by one half-beamwidth, although it is only an approximation to the 3 db beamwidth that it used.

In the system shown in Fig. 1 the outputs of the transducer sections are not taken directly to the delay line, but are taken via a frequency changer. This comprises a quadrature modulator system[2] which when correctly set up gives an output only of the difference frequency between the carrier (i.e. the frequency-swept quadrature oscillator) and the signal. In view of the frequency variation of the carrier, it is not feasible to separate the sum and difference frequencies by filtration in a single modulator stage, and thus the adoption of the quadrature system is necessary. Its action should be seen clearly from the diagram. While there is flexibility in the choice of frequencies, a suitable arrangement of frequencies is shown in Fig. 2. The frequency sweep from $3q/2$ to $(1+k)q$ is

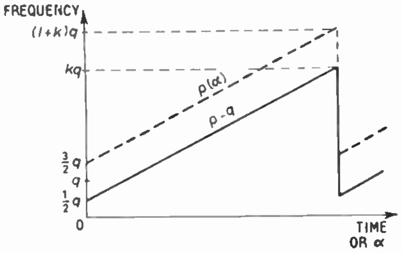

Fig. 2. Variation of oscillator frequency, $p(\alpha)$, under the control of the bearing time-base.

completed in the time of sweep* of the bearing time-base, which, in an echo-ranging system, should not be greater than the duration of the signal pulse if every echo is to be recorded.

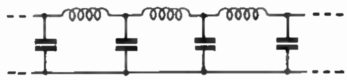

Fig. 3. Form of "low-pass" delay line.

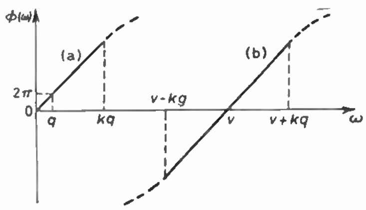

Fig. 4. Phase-shift/frequency response of delay line; curve (*a*) shows the response required with the frequencies as in Figs. 2 and 10, and with the system of Fig. 1; curve (*b*) shows the response required with the "band-pass" system, in which the frequency $p(\alpha)$ is replaced by $v+p(\alpha)$, and only one output of the delay line is used (see Section 2.3).

The frequency applied to the delay-line is $\frac{1}{2}q$ at the beginning of the sweep and kq at the end. If the phase-shift/frequency characteristic of the delay-line (which might in principle have the circuit arrangement of Fig. 3) is linear as shown in Fig. 4(*a*), with a phase-shift of 2π at the frequency q, then it is clear that the beam formed by the transducer is deflected by half a beamwidth at the beginning of the sweep and the deflection increases smoothly to a maximum of k beamwidths at the end of the sweep.

* A saw-tooth sweep as shown is not necessarily most convenient. A V-sweep is easier to produce but may lose targets if they are of negligible dimensions in the range axis. A sinusoidal sweep can also be used if proper allowances are made for its non-linearity with respect to bearing; it is certainly the simplest to obtain in practice, and minimizes transient effects due to sudden changes.

Everything then returns to the beginning and the cycle is repeated and so on.

At the right-hand end of the delay-line, the deflections obtained are to the right, and at the left-hand end they are to the left. If the intensity-modulated cathode-ray display is arranged, as shown, on a sector-scan basis (i.e. rectangular co-ordinates of range and bearing), with separate electron beams dealing with left- and right-hand deflections, then a complete sector-scan over $(2k+1)$ transducer beamwidths is obtained. Care must be taken at the central position to ensure a correct juxtaposition of the left- and right-hand scans; since the minimum deflection is one-half beamwidth, and no central position is ever formed, echoes on the centre bearing are indicated by the coincident

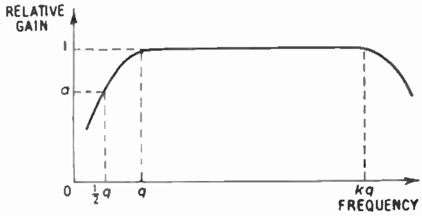

Fig. 5. Frequency response of the gain of the a.g.c. amplifiers. The fraction a is chosen to make the sensitivity of the display on centre bearing equal to that at other bearings.

presentation of signals of two-thirds amplitude from both left- and right-hand scans. Since this method gives a somewhat excessive response to centre-bearing signals, equalization must be provided by reducing the gain of the main amplifiers between the frequencies q and $\frac{1}{2}q$, as shown in Fig. 5. If the display intensity is a linear function of the voltage applied to the grid, then this reduction of gain at frequency $\frac{1}{2}q$ should be about 2·5 db.

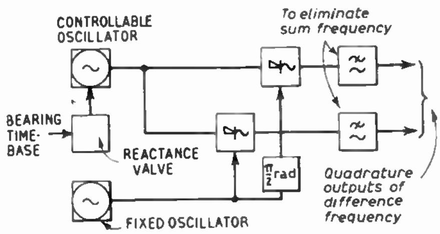

Fig. 6. Frequency-swept quadrature oscillator.

A suitable arrangement for the frequency-swept quadrature oscillator is shown in Fig. 6.

2.1.2 The Modulator Problem

The modulators used in the circuits linking the transducer to the delay-line are preferably devoid of active elements, and the well-known ring rectifier modulator[3] is most suitable. This modulator is, in principle, double-balanced so that none of the input signal and none of the carrier appears in the output circuit. In practice, however, this result cannot be obtained, since four identical rectifiers cannot be obtained, and some leak of both signal and carrier has to be tolerated.

Now the amount of these unwanted components which can be tolerated in the delay-line and subsequent circuits is not easily estimated exactly, because of (*a*) the phase-shifts in the delay-line (which vary with frequency), (*b*) lack of knowledge of the dynamic characteristics of the display, and (*c*) the uncertainty of the observer's response to an irrelevant background. It can be taken, however, that the unwanted components should be at least 20 db below the wanted signals. There is no difficulty in obtaining this amount of suppresion of the input signal by modulator balance, but the carrier leak is another matter. As a rule, the carrier level is kept constant at the longitudinal terminals of the modulator, and thus the carrier leak level is unrelated to signal level, and usually greatly exceeds it except on the strongest signals. This is because the carrier level has to be high enough to prevent serious overloading of the modulator on the strongest signals.

In an echo-ranging system, the signal strength to be expected from a target diminishes rapidly as the range of the target from the equipment increases. Since the range is indicated by the time after the transmission of the pulse, it is clear that immediately after the pulse is transmitted the received signal level will be high, but as time elapses the expected signal level diminishes. If, then, the carrier supply to the modulators is made large immediately after transmission, but then is made to diminish under the control of the range time-base, an adequate, but not more than adequate, carrier level can be maintained all the time. In this way, the carrier leak level can be made to diminish as the signal level diminishes, and an acceptable suppression of carrier leak can be maintained.

The process can be carried out by the use of attenuators, as shown in Fig. 1, which have an attenuation controlled by the sweep voltage of the range time-base, as indicated in Fig. 7. As the carrier level at the modulators diminishes, of course, the modulator conversion loss will increase somewhat; therefore the final design must be a matter of compromise.

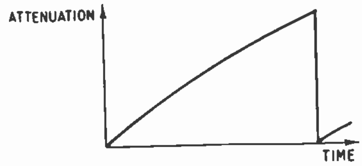

Fig. 7. Characteristic of attenuators controlled by the range time-base.

Another trouble which arises in the modulator stages is that due to third-order modulation. Normally the modulating function of a modulator such as those discussed here contains a fairly large proportion of third harmonic, which means that modulation products $3p \pm q$ are formed. At the beginning of the bearing sweep, the product $3p - q$ can fall within the frequency band of the delay-line plus a.g.c. amplifier circuit, and thus produces spurious signals on the display at a wrong bearing. It is important, therefore, that $3p - q$ products should be kept at least 20 db below the wanted signal. If this cannot be achieved by the modulator circuits themselves, it can certainly be made practicable by using a square-wave carrier signal of the form shown in Fig. 8. This waveform contains no third harmonic, and produces a modulating function which also contains no third harmonic.

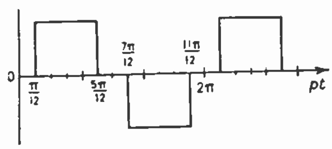

Fig. 8. Waveform of frequency-swept oscillator required for elimination of third-order modulation.

2.2. *The Double-Modulation Alternative System*

It was clear from the description above that the modulation arrangements of the scheme of Fig. 1 led to a number of difficulties in suppressing unwanted products. Although none of these difficulties appears insuperable, yet by the use of two stages of modulation in cascade it becomes possible to remove all the unwanted products by filtration. The double-modulation arrangement, as it may be called, is shown in schematic form in Fig. 9, and the way in which the frequencies are varied by the bearing time-base sweep is shown in Fig. 10.

In the first stage of modulation, the constant carrier frequency s is several times greater than kq, and all the unwanted products, including carrier leak, from this stage of modulation can be filtered out by a band-pass filter as shown. In the second stage of modulation, the carrier frequency is swept by the bearing time-base over a range of $(k-\frac{1}{2})q$ in such a way that the difference frequency is swept from $\frac{1}{2}q$ to kq as in the first system. All unwanted products of this stage of modulation are easily removed by a low-pass filter. Thus the remainder of the system is exactly as before.

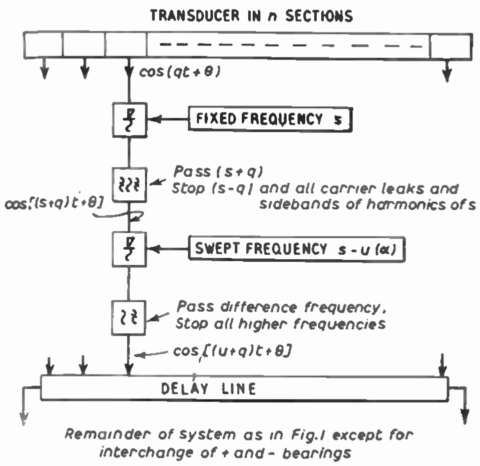

Fig. 9. Schematic of double-modulation system. (N.B. α represents the output of the bearing time-base.)

This double-modulation system has the important advantages of eliminating the need for quadrature paths and the attendant difficulties and expense of accurate phase-relationships, of eliminating the need for the carrier voltage to be varied by the range time-base sweep, and of making special carrier wave-forms unnecessary. Moreover, cruder and cheaper modulators may be used. Disadvantages are that the band-pass filters which have to be provided in every one of the n sectional circuits may not be particularly cheap, and they require to be identical within close limits in order to preserve accurately the phase relations between the sections of the transducer.

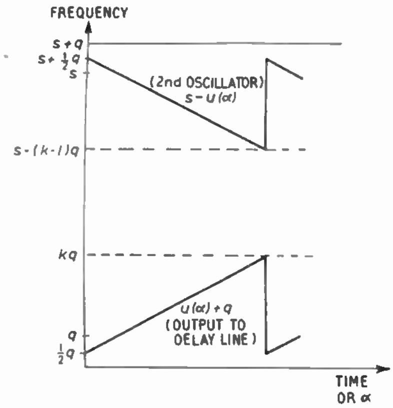

Fig. 10. Double-modulation system: Variation of frequency under the control of the bearing time-base.

2.3. *The "Band-Pass" System*

In either the quadrature or double-modulation schemes described above a modification may be made which removes the need for two channels from the delay-line to the display.

The delay-line previously referred to, and shown in basic form in Fig. 3, is of the low-pass type; that is to say it is of a form which gives a positive phase-shift increasing with frequency, and an attenuation which is nominally zero (if the line is correctly terminated), up to a cut-off frequency, above which the phase-shift becomes constant but the attenuation starts to rise. The term "delay-line" infers a network of this type, but for beam deflection purposes a true delay-line is not necessary as it is only phase relationships which matter in the steady state; of course, at the beginning and end of a pulse the transient response may be affected by the type of network used, and this is discussed in Part 2 of the paper.

If in Fig. 1 the frequency $p(\alpha)$ were raised by a constant amount of several times q, then the frequency applied to the delay line might have maximum and minimum values with a ratio of the order of 2 or less. If, in these circumstances, a network of "band-pass" type, as shown in Fig. 11, is used instead of the low-pass type, then some special advantages are obtained. Such a network, if correctly terminated, has zero attenuation and a phase-shift varying from negative through zero to positive angles over a pass-band, below and above which the attenuation rises and the phase-shift becomes constant. The type of phase response is shown in Fig. 4(*b*). It is clear from this that both negative and positive beam deflections are obtained at any one end of the "delay-line"' and that with suitable arrangement of frequencies and suitable network design, the whole scanned sector can be handled at one end of the network only, with only one main a.g.c. amplifier and only a single-beam cathode-ray tube. Although the bandwidth in c/s required in the amplifiers would be greater than in the low-pass arrangement, yet expressed as a number of octaves it is much smaller, and design might be easier.

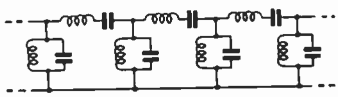

Fig. 11. Form of "band-pass" delay line.

All the problems associated with the modulation stages remain unaltered by the change from low-pass to band-pass working.

Having seen how the band-pass delay-line gives the advantage of having both positive and negative beam-deflection angles represented at one end of the delay-line, it is important to appreciate that the same advantage can be obtained if the band-pass line is replaced, without any other changes, by a low-pass delay-line whose phase-shift is 2π radians per section (or a multiple of 2π) at the frequency corresponding to zero deflection of the beam. In practice, this alternative may well be simpler to design and construct, and it is, in fact, used in the experimental equipment to be described in Part 3 and analysed in Part 2 of the paper.

It should also be observed that in the band-pass system (whether using a band-pass or low-pass delay-line) the basic dependence of swept-frequency range on scanned-sector width has been removed, and the frequency-range required in the delay-line is now free to be chosen in relation to quality of scanning performance, as discussed in Part 2. To reduce the noise level in the delay-line output circuit, evidently as narrow a frequency range as possible is desirable; but it is shown in Part 2 that to reduce distortion and consequent loss of information in echo-ranging systems using short pulses, a relatively wide band is required. Design is therefore, as usual, a matter for compromise.

3. Width of Scanned Sector

Some reference was made earlier to distortion of the beam on large deflection. This matter, and the consequent limitation on the width of the section which can be scanned, will now be dealt with in more detail.

We assume that the transducer is made up of n contiguous sections of length d, so that $nd=l$. It is shown in Reference 1(b) (and elsewhere) that the directivity pattern of this transducer, when all the sections are joined together in the same phase, is the product of two factors:

(*a*) the directivity pattern of the individual sections,

$$\frac{\sin\left(\frac{\pi d}{\lambda}\sin\theta\right)}{\frac{\pi d}{\lambda}\sin\theta}$$

(The angle θ is the direction of an incident acoustic wave relative to the normal to the length of the transducer),

(*b*) the diffraction pattern of n point receivers spaced at a distance d apart,

$$\frac{\sin\left(\frac{\pi nd}{\lambda}\sin\theta\right)}{n\sin\left(\frac{\pi d}{\lambda}\sin\theta\right)}$$

The geometry of the system is shown in Fig. 12, and Fig. 13 shows the above two patterns.

It is clear from Fig. 13 that when there is no phasing of the transducer sections—i.e. no deflection of the beam—there is no "diffraction secondary" at $\sin\theta=\lambda/d$ because the first pattern has a zero there.

When a delay-line is used to deflect the beam, it is only the diffraction pattern (*b*) which is deflected; clearly pattern (*a*) is not affected by

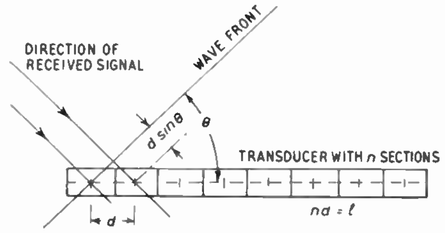

Fig. 12. The geometry of the scanning problem.

the delay-line. Thus as the deflection is increased, two things happen:—

(i) the peak of the resultant directional pattern diminishes,

(ii) the diffraction secondary at an angular distance of $\sin^{-1}(\lambda/d)$ from the main peak is no longer zero, and increases rapidly.

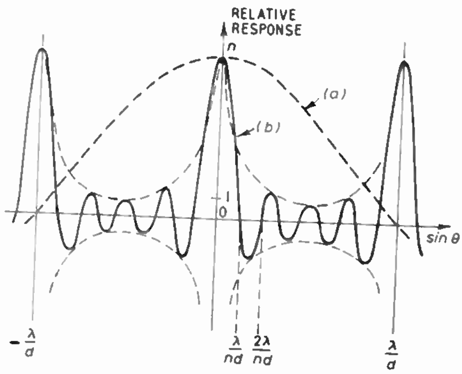

Fig. 13. (*a*) The directional pattern of one section of the transducer. (*b*) Diffraction pattern ($n=9$ in this graph).

When the main peak has been deflected by an angle $\sin^{-1}(\lambda/2d)$, the main and diffraction-secondary peaks are equal. At further deflection, the secondary becomes, in effect, the main peak, and the previous main peak becomes a diffraction secondary. Thus no effective deflection of the beam beyond an angle $\sin^{-1}(\lambda/2d)$ is possible*. The limit of the scanned sector is thus seen to be $\pm\sin^{-1}(\lambda/2d)$ or $\pm\sin^{-1}(\lambda n/2l)$. Since n is the only variable (assuming λ and l are fixed by other design considerations), an increase of scanned sector necessitates an increase in the amount of sub-division of the transducer and a consequent increase in the amount of electronic equipment needed. Since the 3 db beamwidth of the transducer approximates to $\sin^{-1}(\lambda/l)$, it is clear that the maximum scanned sector can be conveniently expressed as approximately n times the beamwidth provided the angles concerned do not exceed, say, about 45 deg.

* Referring to Section 2, it is seen that to obtain this limiting deflection, the phase-shift per section of delay-line is π rad, as would be expected.

4. Design of the Arrays for Transmitting and Receiving

It has so far been assumed that the receiving array is made up of n contiguous sections each of length l/n, and no mention has been made of the important matter in echo-ranging applications of the transmitting side of the system. However, the sections of the receiving array need not be contiguous, and its design is closely linked to that of the transmitting array.

It may be assumed that uniform echo-response over the scanned sector is desirable, and the simplest way to achieve this would be to have a point-source giving a uniform transmitted field over the whole plane containing the length of the receiving array, and to make the latter of n point receivers. The receiving directional pattern would then be the diffraction pattern of Fig. 13 (*b*), and its main lobe would have the same height whatever the angle of deflection. Unfortunately, the diffraction secondary lobes would be of the same height as the main lobe, and displayed information would be ambiguous as to bearing unless the transmitted field were confined to a sector of n times the beamwidth. In practice it is difficult to do this with precision, although a transmitting array with an excitation related to the distance x from the centre of the array by the expression

$$1+2\cos x\pi$$

where x is zero at the centre and unity at each end of the array, has a directional pattern[1b] uniform to ± 3 per cent. of amplitude over an angle of $2\sin^{-1}(\lambda/l)$, i.e. from $-\sin^{-1}(\lambda/l)$ to $+\sin^{-1}(\lambda/l)$. Nevertheless, even this good beam shape has sloping edges, not reaching zero until the angle $\pm\sin^{-1}(2\lambda/l)$; and considerable irrelevant and ambiguous information could thus be displayed.

In practice it may prove most satisfactory to make the receiving array of form intermediate between n point receivers and n contiguous sections. Since the use of sections of finite length means a reduction of receiving sensitivity at angles of deflection other than zero, it would probably be advantageous to make the transmitting array give slight peaking of its directional pattern at angles approaching $\sin^{-1}(\lambda/l)$ with as rapid a fall of response as possible at angles greater than this.

It should be pointed out that a receiving array of n contiguous sections has the best signal/noise performance, so that where performance is expected to be limited by noise arising in the sections of the array, or by uniformly-distributed noise in the medium itself, this design of receiving array should be chosen, and the transmitter made to match as well as possible.

5. Use of a P.P.I. Display with the Scanning System

The description of the scanning system given above has assumed that a B-scan display will be used. This is in many ways very suitable for underwater acoustic equipments, but a p.p.i. display has many advantages. The use of electronic sector-scanning with a p.p.i. display is quite feasible, and means that the array is rotated mechanically in the usual way while the beam is simultaneously scanned electronically. The p.p.i. problem is therefore to ensure that a target on a given bearing is presented on that bearing in the display all the time the scanned sector moves across the target. The method of doing this is shown in Figs. 14 and 15.

The cathode-ray tube of the display has magnetic deflection, with two coils on quadrature axes as shown in Fig. 14. The whole coil assembly rotates round the neck of the tube in synchronism with the rotation of the array. One coil is fed with the range time-base, signal A of Fig. 15, and the other with the bearing time-base, signal B. The scanned sector at any instant is then as shown in Fig. 14, and will clearly be an improvement on the ordinary p.p.i. with only a mechanical scan.

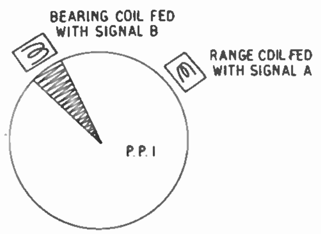

Fig. 14. Schematic diagram showing the use of a p.p.i. display with the scanning system. Both coils rotate, as an assembly, round the neck of the c.r. tube in synchronism with the rotation of the array.

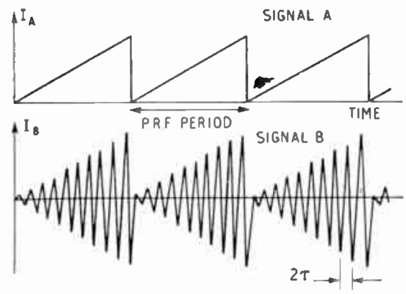

Fig. 15. Time-base arrangements for the p.p.i. display.

Part 2: DISTORTION AND LOSS OF INFORMATION DUE TO SCANNING IN SHORT-PULSE SYSTEMS

6. Introduction

In Part 1 of this paper a description of the electronic scanning system has been given without discussion of the rate of scanning. In general the system will be used for the reception of pulses; its most obvious application, indeed, is to echo-ranging systems. It is therefore important to examine its behaviour on short pulses, and to determine the limitations on the rate of scan. It becomes apparent, as a result of the analysis to be given in this part of the paper, that there are potentially serious effects due to the rapid scanning which cause distortion of the signal presented to the display unit, and thus cause a loss of information. These effects can often be reduced to unimportant proportions by proper design, or can be eliminated by the use of special devices.

Some transient effects are due to the fact that, during scanning, the beam is deflected from the normal axis, without any mechanical movement of the array. This means that a wavefront which is not parallel to the line of the array does not arrive simultaneously at all sections of the array, and thus there is an initial period during which the beam is not properly formed, and scanning is not effective. A corresponding period occurs at the end of a pulse. Distortion due to this cause is reasonably regarded as fundamental to electronic scanning systems, although it can be corrected. Other effects are due to the particular arrangements of the system described in Part 1, and so are not fundamental in the same way.

It will become clear from the analysis that the system should be designed to scan the whole sector at least once per nominal pulse duration. It is thus convenient to refer to the system as a "within-pulse" scanning system.

Since the experimental equipment, which is described in Part 3, uses the double-modulation "band-pass" arrangement discussed in Section 2.3 of Part 1, with a low pass delay-line having a phase-shift of 2π radians per section at the frequency corresponding to zero deflection, the detailed analysis given here is in terms of this particular arrangement. But first we must discuss some general considerations.

The directivity of the receiving system is obtained by the use of a row of transducers ("strip" array) with their outputs added together at the receiver. For an incoming wavefront which is parallel to the line of transducers, all the received signals will be in phase and their sum will be a maximum. For any other direction of approach the individual signal components will generally differ in phase and their resultant will therefore be less than the arithmetic sum of their amplitudes. In order to restore the total output level to its maximum value without changing the angle of approach of the waves it is necessary to introduce phase-shifts, by electronic means, between the outputs of successive elements of the receiving array. As far as the electrical detector circuit is concerned, the additional phase-shifts have roughly the same effect as if the transducer array had been rotated physically relative to the wave system. In other words, the desired scanning of the receiver "beam" can be achieved by purely electronic means if some method can be devised of making the added phase-shifts vary in accordance with a suitable function of time. Two ways in which this might be done have been suggested. One involves the use of a delay-line whose transmission characteristics can be varied (for example, a delay-line involving a ferrite magnetic material might have the permeability of the latter varied by the application of a polarizing magnetic field). Another method, and, in fact, the one which has been chosen for development and which is fully described in Part 1, makes use of a delay-line whose time delay is independent of frequency over the significant frequency range and whose phase-shift is therefore proportional to frequency. If the received signals are first frequency-modulated with a suitable time function and then fed to the delay-line, the desired phase modulation with respect to time can be obtained.

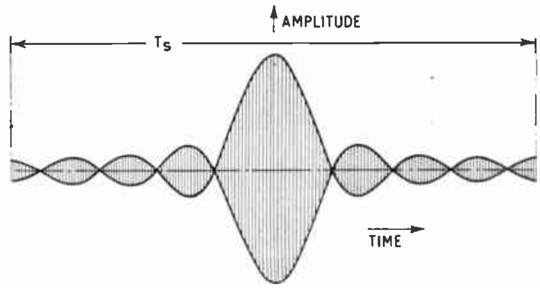

Fig. 16. Form of output waveform for continuous input signal.

If the channel outputs from the delay-line are identical except in phase, they will add together to form a diffraction pattern of the form

$$\frac{\sin n\,(\frac{1}{2}\varphi)}{n \sin\,(\frac{1}{2}\varphi)}$$

where φ is the phase difference between the outputs from any two adjacent channels. If φ is a linear function of time, the total output will have an amplitude-modulation envelope of the form shown in Fig. 16, the peak value occurring at the instant at which all the channel outputs are momentarily in phase.

The system would be arranged so that the peak output occurs at the middle of the sweep when the incoming wavefront is parallel to the line of transducers. This corresponds to a "target" on a zero bearing. For signals received on any other bearing, the peak will be displaced by an appropriate amount to one side or the other of the centre position.

Note that it has been assumed that there is no limitation of frequency bandwidth and that it is possible to devise circuits which will carry out the required functions. The extent to which this ideal can be realized in practice will be discussed later.

7. Signal Distortion

The within-pulse scanning system is liable to distortion due to the following causes:

(*a*) waveform distortion due to the delay-line

(*b*) pulse envelope delay in the medium and in the delay-line

(*c*) bandwidth limitations and non-linearity of phase characteristics.

These effects will be discussed in turn.

7.1. *Waveform Distortion due to the Delay-Line*

Consider first of all the "steady-state" case in which the signal pulses from the transducers are replaced by continuous carriers which are distinguished only by the phase-shifts between them corresponding to the direction from which they are supposed to have been received. These signals will be of the form

$$\sin(q\,t + \varphi_c)$$

Suppose now that one such signal is fed into a multiplicative modulator with a carrier frequency denoted by

$$\sin(p_1 t + \tfrac{1}{2}\varkappa t^2 + \varphi_\omega)$$

i.e. a carrier whose instantaneous frequency increases linearly with time from an initial value of p_1, the constant $\varkappa$ measuring the rate of change of frequency. One of the output components from the modulator will be proportional to

$$\cos(p_1 - q)\,t + \tfrac{1}{2}\varkappa t^2 - (\varphi_c - \varphi_\omega)$$

This will be a waveform whose instantaneous frequency increases linearly with time from an initial value of $(p_1 - q)$. Furthermore, the phase angle of the original signal has also appeared in the output so that, assuming that the same carrier supply is fed to all the channel modulators so that φ_ω is a constant, any phase differences between the original signals from the transducers will be preserved in the frequency-modulated outputs.

The next step is to feed the frequency-modulated channel signals into a network (the delay-line) whose primary function is to produce an additional instantaneous phase-shift which varies with time. Assume to start with that the delay-line is ideal[8] in the sense that its phase-frequency characteristic is a straight line whose intercept on the phase axis at zero frequency is either zero or an integral multiple of 2π. Any waveform fed into it is then merely delayed in time without any change of shape. The analysis can be simplified without loss of generality by assuming that $\varphi_c = 0$ (i.e. the target is on zero bearing) and that $\varphi_{\overline{\omega}}$ is also made zero by a suitable adjustment of the phase of the common carrier supply to all the channels. The inputs to the delay-line will all be of the form

$$\sin(\omega_1 t + \tfrac{1}{2}\varkappa t^2)$$

where $\omega_1 = p_1 - q$,

and the outputs will be

$$\sin(\omega_1 t + \tfrac{1}{2}\varkappa t^2);\ \sin[\omega_1(t - t_2) + \tfrac{1}{2}\varkappa(t - t_2)^2]\ \ldots$$
$$\ldots\ \sin\left\{\omega_1[t - (n-1)t_2] + \tfrac{1}{2}\varkappa[t - (n-1)t_2]^2\right\}$$

where t_2 is the time delay per section of the delay-line. The rth term of this series is

$$\sin\left\{\omega_1[t - (r-1)t_2] + \tfrac{1}{2}\varkappa[t - (r-1)t_2]^2\right\}$$
$$= \sin\left\{(\omega_1 t + \tfrac{1}{2}\varkappa t^2) - \varkappa(r-1)t_2 t - (r-1)\,\omega_1 t_2 + \tfrac{1}{2}\varkappa(r-1)^2 t_2^2\right\}$$

which is of the form

$$\sin[\omega_1 t + \tfrac{1}{2}\varkappa t^2 + \varphi_1 + \varphi_2]$$

where

$$\varphi_1 = -(r-1)t_2(\omega_1 + \varkappa t)$$
$$\varphi_2 = \tfrac{1}{2}\varkappa(r-1)^2 t_2^2$$

Thus the delay-line has had the effect of introducing two phase-shift terms, one of which is proportional to time and the other independent of time.

Furthermore, although φ_1 is proportional to $(r-1)$, φ_2 is proportional to $(r-1)^2$. The former fulfils the required condition for a phase-shift which varies linearly with time and is proportional to $(r-1)$. The latter term, however, clearly represents some kind of distortion because it infers the presence of phase differences which do not increase in equal steps from channel to channel. If the system is to work properly this term must be made negligible. The condition for this to be so throughout the sweep is $\varphi_2 \ll \varphi_1$ for all values of t in the range, i.e.

$$\tfrac{1}{2}\varkappa (r-1)\, t_2 \ll \omega_1$$

for all values of r.

Therefore

$$\tfrac{1}{2}\varkappa (n-1)\, t_2 \ll \omega_1$$

or

$$t_2 \ll \frac{2\omega_1}{\varkappa\,(n-1)} = \frac{2\omega_1 T_s}{(\omega_2-\omega_1)(n-1)} = \frac{2f_1 T_s}{(n-1)\,\Delta f}$$

where ω_1 and ω_2 are the limits of the frequency sweep, $\Delta f = (f_2 - f_1)$ and T_s is the sweep period. Note that for a low-pass delay-line having the required increase in phase-shift of 2π rad* between f_1 and f_2, $t_2 = 1/(f_2 - f_1)$. The inequality above thus leads to the condition for time of sweep:

$$T_s \gg \frac{n-1}{2f_1} \quad \text{[Condition A]}$$

Assuming the above condition to be met, we can now put $\varphi_2 = 0$. For all the channel outputs to be in phase, $\varphi_1 + \varphi_2$ must be an integral multiple of 2π.

Thus, if this is to occur at the middle of the sweep, i.e. when $t = T_s/2$,

$$t_2\left(\omega_1 + \frac{\varkappa T_s}{2}\right) = 2m\pi, \text{ where } m \text{ is any integer,}$$

so that

$$\omega_1 + \left(\frac{\omega_2-\omega_1}{2}\right) = 2m\pi\,(f_2 - f_1)$$

Therefore $$\frac{\omega_1+\omega_2}{2} = m\,(\omega_2-\omega_1)$$

or $$\frac{f_0}{\Delta f} = m \quad \text{[Condition B i]}$$

* This, as shown in Part 1, is the phase variation required to obtain the maximum width of scanned sector.

where f_0 is the mean frequency of the sweep. The mean frequency must thus be an integral multiple of the frequency sweep.

This conclusion has however been based on the assumption that the zero intercept of the phase characteristic of the delay-line is an integral multiple of 2π (including zero). Suppose now that the intercept is actually $(2r\pi + \beta)$ where $\beta \neq 2\pi$. In addition to the time delay t_2, the delay-line will now introduce a constant phase-shift of β at all frequencies. The condition for all the outputs of the channels to be in phase at the middle of the sweep becomes

$$t_2\left(\omega_1 + \frac{\varkappa T_s}{2}\right) = 2m\pi - \beta$$

so $$\omega_1 + \left(\frac{\omega_2-\omega_1}{2}\right) = (2m\pi - \beta)(f_2 - f_1)$$

Therefore $$\left(\frac{\omega_1+\omega_2}{2\pi}\right) = \left(m - \frac{\beta}{2\pi}\right)(f_2 - f_1)$$

or $$\frac{f_0}{\Delta f} = \left(m - \frac{\beta}{2\pi}\right) \text{[Condition B ii]}$$

If the delay-line takes the form of a low-pass filter, β will be zero. (In the first experimental 5-channel system, m has been made unity.) An alternative form for the delay-line is a band-pass filter. In this case there would be some freedom of choice of β so that $f_0/\Delta f$ would not have to be an integral number.

Once the system has been lined up to give a maximum output at the middle of the sweep when all the input signals are in phase (i.e. target on zero bearing), additional phase differences between the inputs (i.e. information about actual target bearing) will displace the instant of maximum output with respect to the middle of the sweep.

7.2. *Distortion due to Pulse Delay Times*

Having established the conditions under which the scanning system would work with continuous input signals, it is now necessary to take into account the fact that the signal received from the target is not actually continuous but is in the form of pulses of length T_p, where T_p is the length of the original pulses sent out by the transmitter.

Consider a wavefront approaching an array of receivers at an angle θ. The elements of the array are spaced a distance d apart. The pulse fronts will not reach the receivers simultaneously but will each lag behind that at the adjacent receiver by an amount t_1, corresponding to the extra distance $d \sin \theta$ which must be covered in the medium.

$$t_1 = \frac{d \sin \theta}{\lambda} \cdot \frac{1}{f_c}$$

where f_c is the carrier frequency in the medium and λ is the wavelength. Clearly, at the limits of the scan, where $\sin \theta = \lambda/2d$, t_1 has the maximum value of $1/2f_c$. In addition there will be the envelope delay t_2 due to the delay-line. Taking the elements in order from one end of the array, the delay-line delay will increase progressively as the signals are taken through longer sections of the delay-line. Thus, although the delay time t_1 is effectively reversed in sign as the sign of θ changes, t_2 does not change sign. This means that the total effect of the two delays is $t_1 + t_2$ for one direction of deflection and $t_1 - t_2$ for the other direction. In an acoustic system, the delay t_1 will generally predominate owing to the relatively low velocity of propagation in the medium.

The envelope curve shown in Fig. 16 applies only to the case where the input signals are continuous. In fact, there will be many scans with no received signal at all. Then, during the particular scan corresponding to the range of the target, the echo pulse front will reach the array. Unless the target bearing is zero, however, the transducers will not be energized simultaneously and the output signal will build up progressively as one, two, three, etc., elements of the array are reached by the wavefront.

Figure 17 shows diagrammatically what is likely to happen. The interval t_b between the peak output and the centre of the scanning period contains the bearing information. Note that the useful information contained by the pulse is

(*a*) position in time of the maximum output (bearing).

(*b*) height of peak (strength of echo possibly giving information about type of target when taken in conjunction with the range).

The range is determined merely by *which* scanning period contains the echo pulse.

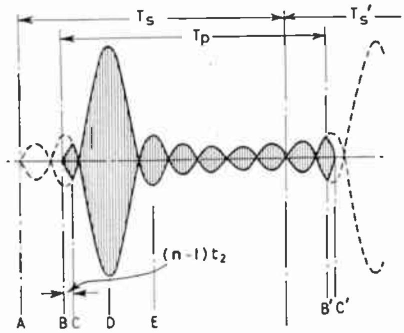

Fig. 17. Distortion due to envelope delays.

B = beginning of build-up. C = end of build-up. D = maximum output. E = centre of scan.

Thus it does not matter when the pulse front arrives provided that it does not coincide with the peak of the diffraction pattern. In simple terms, trouble is bound to occur if the pulse front reaches the array just at the instant when the scanning system is "looking" in that direction.

Figure 18 shows the worst case where the wavefront coincides with the scanning peak. It has been assumed that the pulse length is exactly equal to the scanning period. In this case, instead of a single indication, there will apparently be two targets on the same bearing separated in range by a distance corresponding to one complete scan.

The delay distortion at the beginning and end of the pulse becomes worse as the angle between the incident wavefront and the array is increased, particularly in the direction of deflection for which the two delays t_1 and t_2 are additive. In the worst case the distortion extends over a time $(t_1 + t_2)$. If $(n-1)(t_1+t_2)$ exceeds the effective time-width of the main peak of the pattern, there is a chance of information being lost if the pulse build-up happens to occur at the most inconvenient time.

Another point to be considered is whether or not the pulse *interval* should be an exact multiple of the scanning period. For example, a very small difference between the actual interval and the nearest multiple of the scanning time would introduce a "stroboscopic" effect which might have some advantage in viewing targets at a fixed range because it would ensure that unfavourable conditions could not exist on two successive pulse cycles. It might be an advantage to allow the operator a small range of adjustment of either scanning time or pulse interval, thus enabling him to choose the setting which gives the best results for a given set of conditions.

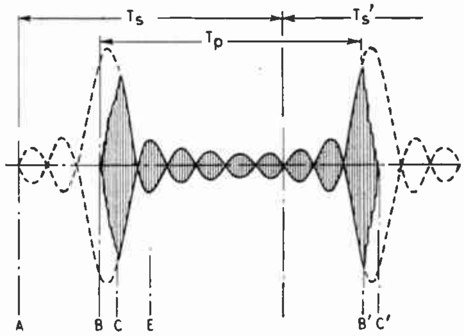

Fig. 18. False range indication due to distortion.

The importance of the effects described is shown by a practical example relating to underwater acoustic echo-ranging. Suppose the signal frequency in water is 50 kc/s, $n=9$ and the frequency modulation sweep is 500 kc/s. Then max. value of $t_1=10$ µsec; $t_2=2$ µsec; $(n-1)$ $(t_1+t_2)=96$ µsec. Taking the beam-width as $1/n$ th of the scanning sector, the time for one beam-width is T_S/n. The delay distortion will extend over one lobe-width of the scanning pattern if $T_S/n=96$ µsec, i.e. if $T_S=860$ µsec. This suggests that a pulse duration as short as 100 µsec would probably be unsatisfactory in this particular example, since the distortion would then extend over a whole scan. It might, of course, be argued that if two scans were made per pulse, then the second scan would give all the information required, and so distortion (which is a loss of information rather than an interference) could indeed be permitted to extend over the whole of the first scan. However, the assessment of such an argument involves some complex considerations regarding signal/background performance which have not yet been worked out; and until further research has been done it seems safer to adopt a conservative criterion of minimum scanning time (and therefore minimum pulse duration) such that the delay distortion is not allowed to extend over more than $1/m$ th of the scan, where m may well be equal to n, but could alternatively be regarded as a constant, independent of n, which defines the probability of losing an echo signal on a particular scan.

Thus, minimum scanning time

$$=m(n-1)\left[\frac{1}{2f_c}+\frac{1}{\Delta f}\right] \quad \text{[Condition C]}$$

It is clear that the design should be made to keep $1/\Delta f$ small compared with $1/2f_c$ so far as possible. It also appears that if condition C is met, then condition A will probably also be met in all practical cases.

8. Choice of Frequency Sweep and Design of Delay Line

A brief discussion follows of the problem of choosing the range over which the signal frequency is to be swept by the frequency modulation which precedes the delay-line. Two parameters have to be fixed; the ratio f_2/f_1 and the sweep $\Delta f=(f_2-f_1)$. One important consideration is that, for a phase sweep of $\pm\pi$ radians (relative to the mid-frequency phase), the time delay is $1/\Delta f$.

It is clear that Δf must be large compared with the acoustic signal frequency if unnecessary delay distortion is to be avoided so that the swept frequency range must lie above the acoustic signal frequency. Unless quadrature modulation is used, an initial stage of fixed-frequency modulation is necessary before the frequency modulation can take place. It can readily be shown that the signal frequency f_3 after the first modulation must be greater than the upper limit f_2 of the frequency sweep.

The first modulator must be followed by a filter whose function is to suppress the first carrier and its upper side-band. The design of this filter becomes progressively more difficult as the first carrier frequency is raised, thus bringing the side-bands relatively nearer to the carrier.

The design of the delay-line requires that between successive tapping-points there must be a maximum change of phase-shift of 2π radians for a frequency shift of Δf. If x denotes the ratio f/f_0 where f_0 is the cut-off frequency of the filter forming the delay line, the change of phase shift of 2π must be obtained for a change of x from x_1 to x_2

where
$$x_1 = \frac{f_1}{f_0}$$
$$x_2 = \frac{f_2}{f_0}$$
$$(x_2 - x_1) = \frac{f_2 - f_1}{f_0} = \frac{\Delta f}{f_0}$$

For a constant-k low-pass filter the phase-shift per T or π "cell" is given by $\sin \beta/2 = x$ and the slope of the phase characteristic $d\beta/dx = 2/\sqrt{(1-x^2)}$. At this stage it is necessary to choose the values of x_1 and x_2 somewhat arbitrarily. The following conditions must be met.

(*a*) x_1 must be large compared with f_c/f_0.

(*b*) x_2 must be three times x_1.*

(*c*) x_2 must be small enough for the phase curve to be reasonably linear over the swept range.

(*d*) The phase change $\Delta\beta$ must be as large as possible to avoid having an unnecessarily large number of cells in the filter.

A numerical example, based on the experimental equipment described in Part 3, will illustrate the design procedure. In this equipment x_1 was chosen to make $x_1 = 5(f_c/f_0)$; i.e. $f_1 = 5\,f_c$.

Since $f_c = 50$ kc/s

therefore $f_1 = 250$ kc/s

$f_2 = 750$ kc/s

$f_0 = 1{\cdot}25$ Mc/s

$x_1 = 0{\cdot}2$

$x_2 = 0{\cdot}6$

To improve the linearity of the phase curve it would be necessary to reduce the range between x_1 and x_2 by increasing the cut-off frequency. For example, if x_1 is made 0·15 instead of 0·2, the cut-off frequency would be 1·67 Mc/s and x_2 would be 0·45.

* $\frac{1}{2}(x_1 + x_2)$ must be an integral multiple of $x_2 - x_1$, when a low-pass delay-line is used. In this case $\frac{1}{2}(x_1 + x_2) = x_2 - x_1$.

In the present case,
$$\Delta\beta = 2\,[\sin^{-1} 0{\cdot}6 - \sin^{-1} 0{\cdot}2] = 0{\cdot}89,$$
therefore the number of cells required is
$$\frac{2\,\pi}{0{\cdot}89} = 7.$$
If x_1 is reduced to 0·15,
$$\Delta\beta = 2\,[\sin^{-1} 0{\cdot}45 - \sin^{-1} 0{\cdot}15] = 0{\cdot}59,$$
number of cells is $\dfrac{2\pi}{0{\cdot}59} = 10{\cdot}6$, say 11.

In the first case $d\beta/dx$ would vary from 2·04 to 2·5, i.e. a variation of about ± 12 per cent. on the mean. The decrease of x_1 to 0·15 would reduce this to about ± 5 per cent. but at the expense of a 50 per cent. increase in the number of cells in the delay-line.

Further theoretical and experimental work will be required to enable to optimum design to be chosen.

9. Bandwidth of Delay Line

So far very little has been said about the question of band-width restriction and its effect on the shape of the signal pulse. It has been tacitly assumed in fact that no such restrictions occur. On the other hand, in discussing the delay-line and the choice of frequency sweep, the signal has been treated as though it were represented by a single frequency. Actually, of course, this could only approach the truth if the scanning speed were very low. The greater the scanning speed the greater the bandwidth which must be allowed for to accommodate the significant side-bands produced by the frequency modulation.

Some idea of the bandwidth needed for the frequency modulation can be obtained by assuming a sinusoidal modulation with time (instead of the sawtooth waveform which would probably be used in practice). Then, roughly,[9] the band should exceed the swept range by eight times the scanning frequency. For 1 msec pulses this would mean an additional 8 kc/s on the 500 kc/s sweep, i.e. 1·6 per cent. increase.

Although this is only an estimate (based on 1 per cent. distortion for sinusoidal modulation) the result is not likely to be very different for the sawtooth case and it shows that, for pulses as long as 1msec there should be no difficulty in meeting the requirement that the phase and attenuation characteristics must remain correct, not only for the sweep Δf, but for enough outside the sweep to include the side-bands. In any case, however, even if distortion does occur from this cause, it will only apply at the limits of the scan.

The next step must be to check the bandwidth necessary to give a sufficiently rapid rise time to the envelope of a scanning pulse. Here the problem may be more serious because further pulse distortion is likely to occur if the rise time exceeds the total delay time difference (t_1+t_2). In the experimental case under consideration this delay has a maximum value of 12 μsec. This suggests that the effective bandwidth should exceed the sweep by something like 80 kc/s (i.e. 16 per cent. of the sweep). The bandwidth requirement is thus likely to be dictated, not by the presence of the frequency modulation, but by the required build-up time of the signal pulses which must not be unduly increased by the delay line. We see in fact that the delay line has two disadvantages from the pulse transmission aspect:

(*a*) it puts in differential delays between successive channels,

(*b*) it causes further pulse distortion due to the fact that its linear frequency range is restricted. (This trouble will be most marked at the end of the scan corresponding to the upper swept frequency because it is above the swept band that the non-linearity of the phase characteristic rapidly becomes worse.)

An important fact appears to be emerging so far, however: the limitation to the scanning speed (and thus the information rate) of the basic system is not likely to be set by the electronic circuitry. The limitation is probably set simply by the inevitable differential wave-front delays which occur when the sensitivity pattern of an array is deflected from its geometrically normal position. As pointed out previously, this appears to be the basic price which has to be paid for the convenience of electronic deflection as opposed to mechanical rotation of the array. Methods of removing the effect, however[7], are under investigation, and if these are successful, the limitations will then be set by the circuit design problems.

Part 3: THE EXPERIMENTAL SYSTEM AND RESULTS

10. Introduction

An experimental equipment was designed and built according to the block schematic of Fig. 19. This uses the double-modulation system with low-pass delay line. The object was to prove the principle of the scanning system, and to confirm that the theory has not overlooked any important source of scanning distortion as far as this is due to the delay line. It was therefore unnecessary to simulate the variation of received echo strength with range, and a constant signal level was used, together with constant gain throughout the system. Moreover, it was the waveform of the signal at the output of the delay line which was of greatest importance, and it was this which was observed on the cathode-ray oscilloscope; it was unnecessary to provide a B-scan or p.p.i. display. Deflection of the "target direction" was effected simply by inserting suitable phase-shifts in the signal input channels, representing the phase relations at the various sections of the array. Five input channels were provided, representing a five-section array, and the signal frequency was taken as 50 kc/s. This is, in fact, a suitable frequency for fish-finding under-water acoustic echo-ranging equipment. The signal could be continuous or pulsed; in the latter case the gating circuit, driven by a slow time-base, was used; for the former it was disconnected. The gating circuit was used to pulse the fixed local oscillator instead of the signal source, merely as a matter of convenience; this makes no difference to the results.

11. Outline of Design

11.1. *The Delay Line*

The way in which the delay line is designed was described in Part 2. The type of line used is a low-pass filter giving 2π radians phase-shift at the centre frequency of the swept band applied to it. The circuit diagram is shown in Fig. 20, from which it is seen that each section of the line (i.e. between tapping-points) comprises seven filter cells. The nominal cut-off frequency is 1·25 Mc/s, and the centre frequency of the applied band is 500 kc/s; the sweep is from 250 to 750 kc/s, approximately.

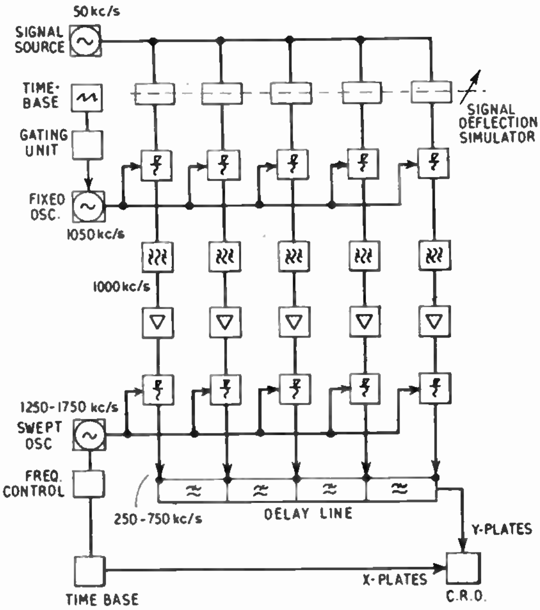

Fig. 19. Block schematic diagram of the experimental equipment.

Matching half-cells are used at each end of the line, with derivation parameter $(m) = 0{\cdot}6$, so that resistance terminations can be used. The design impedance is 300 ohms. (All the figures quoted above are the nominal design values. The actual equipment made had values slightly adjusted throughout to enable preferred-value components to be used; this accounts for some slight discrepancies in the diagrams.)

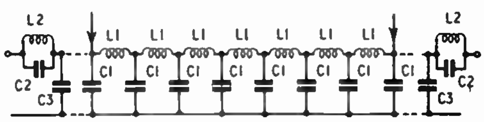

Fig. 20. Delay-line circuit.
C1 = 820 pF C2 = 438 pF C3 = 246 pF
L1 = 77 μH L2 = 23·1 μH

The measured phase-shift/frequency responses of the delay line between the tapping points and one end are shown in Fig. 21. The phase errors due to imperfect design and construction do not exceed about ±10 deg. This is evidently good enough for practical purposes.

11.2. *Channel Equipment*

The circuit arrangement of each channel between the signal inputs (corresponding to the transducer array in the real equipment) and the delay line is shown in Fig. 22. All five channel units are identical.

The modulator circuit used in both stages is the transformerless balanced rectifier modulator described in reference 3, p. 75; this was chosen mainly for convenience, but it is reasonably stable in respect of conversion loss and its phase-accuracy is high and predictable. In the first-stage modulator a balancing potentiometer was used, not so much to reduce carrier-leak, as to minimize a second-order inter-modulation product at 950 kc/s which seriously spoils the scanning patterns obtained, and must be removed in an operational system.

The filter following the modulator has to give a high performance; it is required to pass 1000 kc/s (together with side frequencies when pulsed signals are used), but reject 1050 kc/s (carrier leak) and 1100 kc/s (unwanted sideband) with an attenuation of at least 40 db. This choice of 40 db is, of course, arbitrary, but it was estimated that a poorer performance would so spoil the output waveforms that it might be difficult to check results with theory. The circuit arrangement of the filter is included in Fig. 22, and its insertion-loss/frequency response is shown in Fig. 23. The high-pass portion is included to attenuate the input signal of 50 kc/s, which is not balanced out by the modulator.

Since the phase-shifts in all channel units must be the same if proper scanning patterns are to be obtained, the filter is made of high-accuracy components, the tolerance being 1 per cent. The amplifier valve following the filter is required to make up for the losses in the circuit.

The tappings on to the delay line must be of high impedance to avoid spoiling the performance, and consequently a coupling valve was used in the output of the channel unit.

12. Results Obtained

It must be appreciated that in this first experimental equipment it was not easy to make adjustments, particularly of phase, in order to get all channels lined up to the identical overall amplification and phase-shift. Moreover, there was serious intermodulation distortion as mentioned in the previous section. Therefore the scanning patterns obtained—which should look like that of Fig. 16 except with 5 lobes instead of 9—are not perfectly shaped. They show clearly, however, that the system performance is satisfactory and that a practical sector-scanning pulse echo-ranging system can be made on this basis.

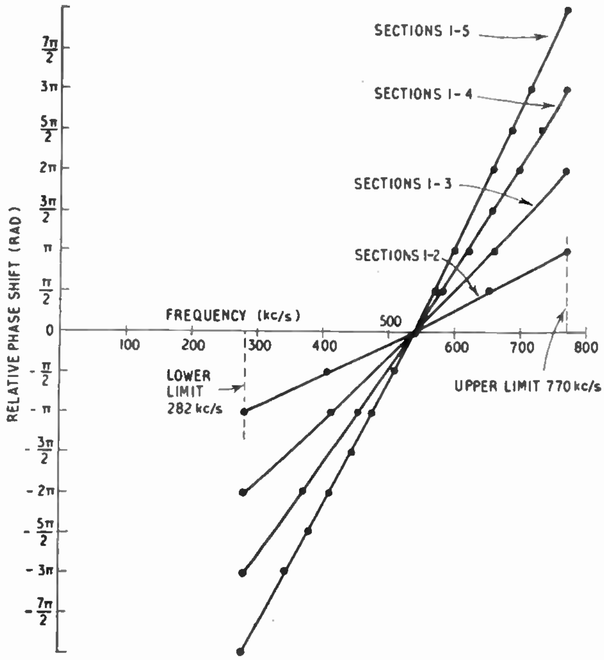

Fig. 21. Delay-line: measured phase-shift/frequency characteristics.

When a steady, continuous, input signal is used, a scanning pattern (as the output signal-envelope from the delay line is called) is obtained which shows a main lobe in the position on the time-base corresponding to the direction of the received target signal—or in the present experiment, corresponding to the phase-shift setting of the signal deflection simulator. Fig. 24 shows the pattern obtained, for example, with the signal coming from a direction approximately one beam-width to the left of the normal to the array. In this case, the sweep frequency was 100 c/s.

When the input signal consists of a short pulse, the scanning period must not be greater than the pulse duration, as shown earlier. Fig. 25 shows the pattern obtained when a 0·15 msec pulse is scanned three times within its duration. The time-base was sinusoidal in this test*. Of course, to give a proper bearing display, the three scans should be superposed so that the three peaks coincide; but they are here presented serially so that it can be seen that even at this high rate of scanning there is no observable scanning distortion due to the delay line. Distortion due to differential wavefront delay is not represented here since the deflection simulator merely gives suitable phase relationships and does not give time delays.

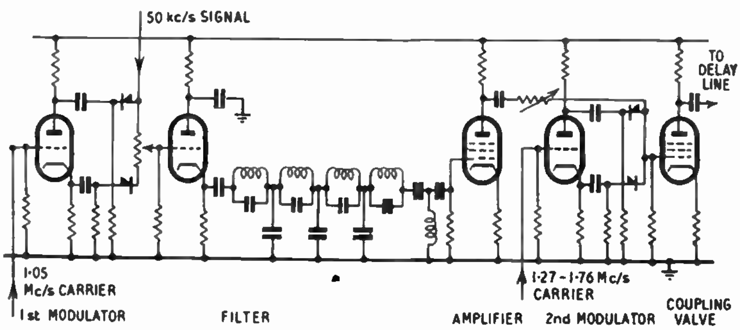

Fig. 22. Circuit of channel units.

It should be noted that for this experimental system (without differential wavefront delay) conditions A and C of Part 2 give limiting scanning speeds (A) much

* As pointed out in Section 2.1.1 of Part 1, the sinusoidal sweep of frequency is much easier to produce than a saw-tooth sweep, and avoids certain transient effects. There is a lot to recommend its use in practice.

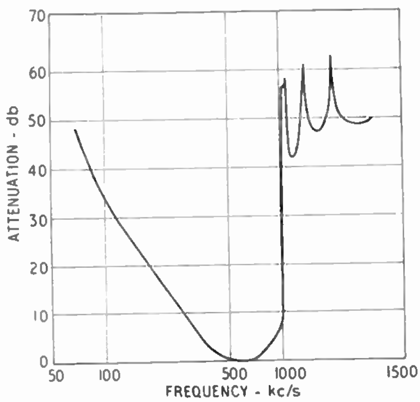

Fig. 23. Loss/frequency response of filter in channel units.

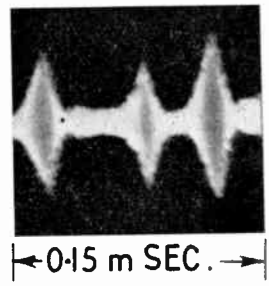

Fig. 25. Scanning patterns for pulse signal,

scanning rate = 20,000 scans/sec
pulse duration = 0·15 msec

[The frequency-sweep was sinusoidal and not saw-tooth; this accounts for the unequal spacing of the peaks. Signal direction = +90° (in electrical degrees).]

less than 125,000 scans/sec and (C) not greater than 25,000 scans/sec, respectively. Thus, theoretically, the rate of 20,000 scans/sec used for Fig. 25 is approaching the limit. The experiment therefore confirms that the theory has not overlooked any important cause of distortion or failure.

13. CONCLUSIONS

Part 1.—A method is described of providing a rapid scan of an acoustic beam (e.g. asdic) over a sector many times the beamwidth. No mechanical operations are involved, the beam being deflected by connecting the sections into which the transducer is divided to points along a delay line after suitable frequency-changing. The swinging of the beam is accomplished by translating the signal frequency received from the transducer to another frequency before applying it to the delay line. The local frequency used in the translation process is swept over a range by the time-base used for the bearing axis in a cathode-ray sector-scan display, so that the frequency applied to the delay line varies from a low to a high value. If the delay line has a phase-shift which increases with increase of frequency, then the beam deflection increases from a low to a high value as the frequency changes. Other arrangements of the scheme are also discussed.

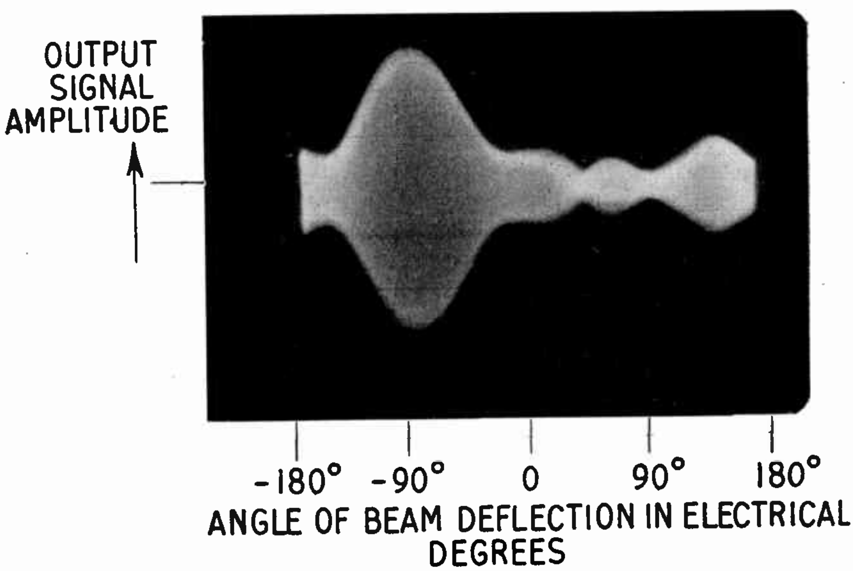

Fig. 24. Scanning pattern for steady input signal, scanning rate = 100 c/s. Signal direction corresponds to −90°.

Suitable electronic arrangements are described, and it is shown that the maximum scanned sector is approximately n times the 3 db beamwidth of the acoustic beam, n being the number of sections into which the transducer is divided.

Part 2.—Problems associated with the use of the scanning system in an echo-ranging equipment having a short-duration pulse are studied, and it is shown that a loss of information can occur due to the association of time-delay with phase-shift in the delay line, and due to the fact that the beam is deflected without mechanical rotation of the array. Methods of reducing these effects to negligible proportions are discussed.

Part 3.—An experimental equipment has built to demonstrate the principles of the scanning system in the laboratory, and this, and the results obtained with it, are described. It is confirmed that the principles are sound, and that a practical system for underwater acoustic applications is feasible.

Note. A full-scale operational trial of such a system will shortly be made, and the results published in due course. It is, moreover, appreciated that the system can be applied to radio and radar[7], and the results of an investigation into this application will also be published later.

14. Acknowledgments

The work was done entirely in the Electrical Engineering Department of the University of Birmingham, but the authors are indebted to the National Institute of Oceanography for a substantial financial contribution towards the cost of the work. Patent rights in the system have been taken over by the National Research Development Corporation. For useful scientific and technical discussion the authors are indebted to several of their colleagues in the Department, and particularly to Mr. J. M. Layton and Mr. J. W. R. Griffiths.

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THE OPTIMUM DESIGN OF ELECTROSTATICALLY DEFLECTED CATHODE-RAY TUBES*

by

Hilary Moss, Ph.D. (Member)†

SUMMARY

The optimum design is defined as the one which, for a given spot size and beam current, has minimum defocusing and maximum deflection sensitivity. The following parameters are supposed to be fixed: (1) over-all tube length; (2) screen diameter; (3) maximum accelerating potential; (4) cathode loading. By making use of the equations of motion in a parallel plate deflector system and an expression for limiting current density due to Langmuir, a solution is reached as follows: (a) make the electron gun as short as possible so that the deflectors can be mounted as far as possible from the screen; (b) make the deflectors as large and as sensitive as possible.

LIST OF SYMBOLS

V	Maximum volt-velocity of the electron beam.
$\overline{V}$	Voltage between deflector plates.
$\nabla\overline{V}$	$=\overline{V}/d=$Potential gradient between deflector plates.
ρ_c	Current density at cathode surface.
ρ_0	Current density in focused spot.
e	Electron charge.
m	Electron mass.
k	Boltzmann constant.
T	Cathode temperature in degrees Kelvin.
v_z	Axial velocity of electron corresponding to volt-velocity V.
l	Length of deflector plates.
d	Separation between deflector plates.
$2s$	Mean beam width in deflector plates.
L	Distance between exit end of deflecttor plates and screen.
α	Semi-angle of cone of rays from triode.
λ	Angle of deflection, i.e., angle between tube axis and centre ray in deflected beam.
θ	Semi-angle of cone of rays converging to focused spot.
M, N, K	Scaling factors.
D	Deflection of beam at screen.

1. Introduction

The over-all design of any cathode ray tube is essentially a compromise between a very large number of conflicting variables. When discussing these matters it is most important that a very clear statement be made of which of these variables are being regarded as fixed parameters and which are being allowed to vary in accordance with some design principle. Much confusion is caused—and occasionally heated arguments generated—by misunderstanding as to what postulates are being made.

The present author has treated the geometrical design of electron guns at length in a number of previous papers.[1,2,3] For the most part, these discussed the actual electron-ray generator, i.e. the gun proper rather than the gun in association with its deflector system. In this paper the over-all system, including the deflectors and the envelope is studied, so as to arrive at important principles concerning the over-all design of the best tube.

In nearly all practical cases there are three variables which are fixed by operational considerations. They are the over-all length, the

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screen size, and the maximum available accelerating potential. These are imposed by the user.

The fourth parameter, which is also fixed, is set by the designer. This is the maximum permissible cathode loading, i.e. the maximum current density at the cathode surface. Although knowledge of the relationship, if any, between cathode loading and life is very far from complete, there is general agreement, based entirely on observation, that very high loadings seem to lead to short life. Most designers regard 1 amp/cm^2 as about the maximum limit advisable.

There are thus four constraints and within them it is now proposed to show how the optimum tube can be designed by following certain simple principles.

2. The Correct Division of the Fixed Over-all Tube Length Between the Gun and the Deflectors

Figure 1 shows a sketch of the essential geometry. S_1 is the plane of the crossover (or object) and S_2 is the plane of the final image or spot in a cathode-ray tube. We are postulating a fixed over-all length so that the distance between S_1 and S_2 is fixed. We also postulate fixed screen size so that the maximum deflection $II'=D$ is also fixed.

In 1937 Langmuir[4] was able to show that the maximum image current density ρ_0 was given by—

$$\rho_0=\rho_C\left\{\frac{eV}{kT}+1\right\}\sin^2\theta \quad ....(1)$$

From this formula it may be seen that for constant loading ρ_C and constant final accelerating potential V (both of which are postulated as fixed) the maximum image current density is proportional to $\sin^2\theta$. It is thus defined once θ is defined. On Fig. 1 the cone of rays reaching the screen S_2 at this angle is drawn, and by eqn. (1) the image density is thus once and for all defined. Its actual value is not relevant. It is being set by the operational requirements and the knowledge of the practical values which are achieved for given angles.*

In Fig. 1 are also shown the idealized trajectory of the rays from the crossover O. Two paths are shown: one is OAI and the second OA′I. These two different paths are due to changes in the generating gun. The position of the final focusing lens has been varied and the angle of the rays (α) being handled from the crossover is also different. These alterations affect, of course, the magnification of the gun and (possibly) the absolute size of the image or spot. However, for the present, these variations are of no interest since the Langmuir equation (1) shows that the image *density* is the same for both sets of trajectories because the final convergent angle (2θ) of the cone of rays forming the spot is constant. The matter of maintaining constant spot *size* as distinct from constant spot *density* will be dealt with in Section 3.

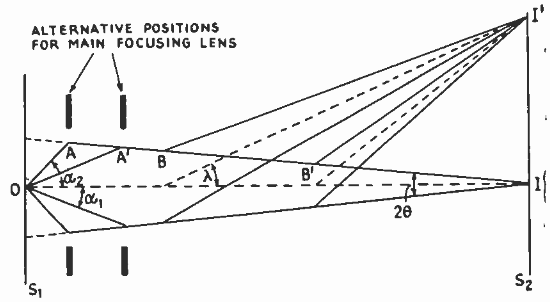

Fig. 1. Geometrical diagram for the space between crossover and screen of cathode-ray tube.

Transferring attention to the deflector system, the idealized trajectories of two deflection geometries are shown in Fig. 1. In the first the deflection starts at point B and forms a deflected image I′, while in the second it starts at B′ and forms the same image at I′ Obviously an infinite number of such trajectories is possible depending on whereabouts the deflector plates are located along the beam axis. The problem is to investigate the optimum position.

Since the image current density has been fixed by maintaining constancy in the final convergent angle of the beam, the choice of the position of the deflectors is wholly determined

* The practical values are lower than those given by eqn. (1) which defines the theoretical upper limit. This does not affect the reasoning advanced. From the present viewpoint a much less explicit equation would suffice, namely, that $\rho_0=\Phi[\rho_C, V, \theta]$ where Φ is an arbitrary function.

by considerations of minimizing the deflection defocusing and maximizing deflection sensitivity. The question is, in effect, "Is it better to deflect a somewhat smaller beam cross-section through a larger angle, or a larger beam cross-section through the smaller angle?"

To study this, the electron deflection in a parallel plate condenser as shown in Fig. 2 is investigated, neglecting fringing fields and assuming that the lateral deflecting field is sharply bounded by the planes XY, X′Y′.

It will also be postulated that the deflection is "balanced" in the manner shown in Fig. 2, so that the median plane of the deflectors remains always at final anode potential V.

The assumption that the deflecting field is sharply bounded by the planes XY and X′Y′ is equivalent to the following situation. An electron following a track such as ABCD moves at volt-velocity V from A to B. At B it *suddenly* jumps to a new volt-velocity. The slight convergence of the beam is neglected so that the velocity change at B is wholly axial. From Fig. 2 it is seen that this new volt-velocity

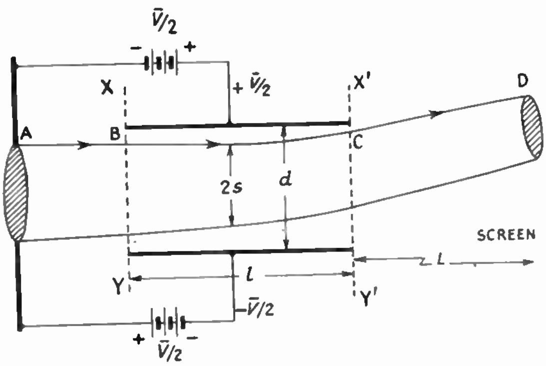

Fig. 2. Diagram of deflection in a parallel plate condenser.

just *inside* the plate system at B is $V+\bar{V}.s/d$. Since the field between XY and X′Y′ is wholly normal to the tube axis, this *axial* velocity remains constant throughout the plate region. Changes in the electron velocity during passage through the plates are wholly lateral. On emerging from the plates at C the *axial* volt-velocity component of the electrons is *suddenly* reduced and reverts to V. There is no change in the tangential velocity component across the boundary X′Y′.

These facts are now expressed symbolically. Let v_1 be the new axial velocity of an outer edge electron at the *top* edge of the beam when just to the right of XY. Let v_2 be the velocity of an outer edge electron at the *bottom* edge of the beam when just to the right of XY. Then

$$\tfrac{1}{2} m v_1^2 = \left\{ e\,[V+\bar{V}.s/d] \right\} \quad \text{.........2(a)}$$

$$\tfrac{1}{2} m v_2^2 = \left\{ e\,[V-\bar{V}.s/d.] \right\} \quad \text{.........2(b)}$$

For an electron following the top track ABCD the plate transit time is l/v_1. The lateral acceleration of any electron within the plates is $(e/m).\nabla\bar{V}$. Thus the radial velocity of an electron at the top edge of the beam on leaving the plates is $(e/m).\nabla\bar{V}.(l/v_1)$. Just *after* passing the exit plane X′Y′ its angle to the tube axis will be arc tan $(e/m).\nabla\bar{V}.(l/v_1).1/v_z$. At the screen, distant L from X′Y′, this angular deflection gives a displacement of $(e/m).\nabla\bar{V}(l/v_1).L/v_z$.

Assembling these results, and using the identity $\nabla\bar{V}\equiv\bar{V}/d$ reveals that, at the screen, the displacement between extreme electrons from the opposite top and bottom beam edges is given by

$$\Delta D = \frac{e}{m}\cdot\frac{\bar{V}}{d}\cdot l\left[\frac{l}{2}\left\{\frac{1}{v_2^2}-\frac{1}{v_1^2}\right\}+\frac{L}{v_z}\left\{\frac{1}{v_2}-\frac{1}{v_1}\right\}\right] \quad \text{.........(3)}$$

By substituting from 2(a) and 2(b) into (3), neglecting $\bar{V}^2.s^2/d^2$ in comparison with V and writing

$$\left\{1-\frac{\bar{V}^2}{V^2}\cdot\frac{s^2}{d^2}\right\}^{\frac{1}{2}} = 1-\frac{\bar{V}^2}{V^2}\cdot\frac{s^2}{2d^2}$$

gives, after some reduction,

$$\Delta D = \frac{1}{2}\,\frac{\bar{V}^2}{V^2}\,\frac{1}{d^2}\cdot l\,(L+l)\,.\,s \quad \text{.........(4)}$$

Similarly it is readily shown that the deflection at the screen of a central ray in the beam is given by

$$D = \frac{\bar{V}.l}{2d.V}\left\{L+\frac{l}{2}\right\} \quad \text{.........(5)}$$

and also that the deflection of the central ray in the beam at the exit plane X′Y′ is

$$D\Big|_{X'Y'} = \frac{\overline{V}.\,l^2}{4d\,.\,V} \qquad \text{.........(6)}$$

A careful study of eqns. (4), (5) and (6) will answer the question as to the proper location for the deflectors.

Suppose first that parallel deflectors of length l and spacing d are placed in position 1, Fig. 3. It will be assumed that l is chosen so that at the maximum defined deflection the beam just grazes the exit edge of the plates, i.e. l has its maximum value, consistent with avoiding plate cut-off.

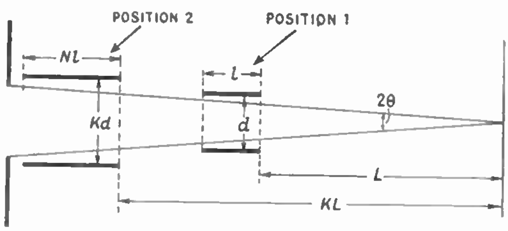

Fig. 3. Illustrating effect of moving the parallel plate system.

The plate system is now moved back towards the gun into position 2. The distance L now becomes KL where K is some factor greater than unity. From Fig. 3 it can be seen that the new plate spacing becomes Kd. Suppose now that the new plate length becomes Nl and the new deflector plate voltage $M\overline{V}$.

In the new plate position, as in the old, it is reasonable to make the new plate length Nl of maximum value, i.e. just below the start of beam cut-off. In position 1 the mean beam width in the deflectors is $2s$ so that the mean spacing between the undeflected beam edge and the plate is $(d-2s)/2$. In position 2 the mean beam width is very nearly $2Ks$ and the plate spacing is Kd. Thus the new gap between the undeflected beam edge and the plate is $(Kd-2Ks)/2$, i.e. is multiplied by K. By eqn. (6) this must mean that $\overline{V}\,.\,l^2/d$ is to be multiplied by K. Hence the first requirement is

$$\frac{MN^2}{K} = K \qquad \text{.........(7)}$$

Again the total beam deflection is to be unchanged. Thus using eqn. (5) the second requirement is

$$\frac{MN\,(KL+N\,l/2)}{K\,(L+l/2)} = 1 \qquad \text{.........(8)}$$

Substituting for M in (8) from (7) gives

$$N = K^2\left(1 + \frac{l}{2L} - \frac{Kl}{2L}\right)$$

Since $2L \gg l$ we find that $N \cong K^2$. Similarly it follows that $M \cong K^{-2}$. Thus the new plate length becomes $K^2 l$ and the new deflector plate voltage $\overline{V}/K^2$.

Using eqn. (4) the new defocusing of the plates in position 2 is

$$\Delta D = \frac{1}{2}\cdot\frac{\overline{V}^2}{K^4}\cdot\frac{1}{V^2}\cdot\frac{1}{K^2 d^2}\cdot K^2 l\,(KL+K^2 l)\,.\,Ks$$

Again assuming $L \gg l$ this shows that the new defocusing is approximately K^{-2} times the old. In addition since, the new sensitivity is increased approximately K^2 times it follows that the new plate system in position 2 is markedly superior. Table 1 summarizes these important results.

This answers the first question and leads to the first principle:

Principle (1)

"The electron gun should encroach on the *minimum* amount of tube length so that the deflectors are as far from the screen as possible."

3. Manipulation of Gun Parameters to Maintain Spot Size

Principle (1) makes a system of high magnification. It is therefore necessary to study how the spot size may be kept below some defined maximum value.

From Fig. 1 it is seen that if the final focusing lens plane is moved from A′ back to A, the semi-beam angle of rays from the triode must be increased from α_1 to α_2 so as to generate the same angle of convergence 2θ in the rays to the image. This may be readily achieved by a reduction in the first anode to modulator spacing[2]. If the cathode to modulator spacing is now adjusted so as to maintain constant cut-off voltage, then it can be shown that the triode giving the semi-beam angle α_2 (at zero modulator voltage) has virtually the same modulation characteristics as the original triode, giving semi-beam angle α_1 (also at zero modulator voltage). These design factors are discussed in detail in reference 2, while an abbreviated account is contained in reference 5.

All that happens when the first anode to modulator spacing is reduced *and the cut-off voltage is maintained constant—by readjustment of the cathode to modulator spacing*—is that the total currents from the triodes are almost identical at all points on their modulation characteristics, but the solid angle into which these currents are projected is changed. It can also be shown that the constancy of modulator hole diameter coupled with the identity of the modulation characteristics gives constancy of cathode loading[2]. Furthermore, the crossover size varies in such a manner that the final image size remains constant. This is quite accurately true if the beam angles α_1, α_2 are small. Such is often the case since on electrostatic tubes the beam width in the deflectors is limited severely by an aperture stop in the final anode. Hence, the conclusion is simply to make a very short gun with a very small first anode to modulator spacing. This compensates for the increased magnification. Thus, the final spot is maintained of constant diameter.*

Principle (2)

"The electron gun is designed with a very small modulator to first anode gap so as to give a large beam angle α from the triode and a very small crossover."

4. Optimum Design of the Deflector Plates

So far parallel plate deflectors have been considered. The analysis of the defocusing mechanism postulated a parallel system, but a little thought will show that the derivation of eqn. (4) is not likely to be greatly upset if other plate shapes are considered. It is henceforth assumed that eqn. (4) will apply to a non-parallel plate system.

So far the analysis has shown that the absolute maximum amount of the fixed tube length should be used to effect deflection. Suppose that the plates have been mounted as far as possible towards the crossover. From eqn. (4) it is seen that the deflection focusing is proportional to the square of the deflecting voltage, while from eqn. (5) the actual deflection is linearly proportional to the deflecting voltage. It thus seems reasonable to conclude that the best deflecting system will be the one having the greatest sensitivity, i.e. the one requiring the minimum deflecting voltage for a given spot

* Of course this fact must follow from the following simple consideration, using the Langmuir equation: Keeping ρ_C and θ constant has kept the spot density constant. But the beam current has been maintained constant. Thus, by definition of density, the spot size must also be constant.

Table 1

	Variable	Multiplier
Basic Operations	Screen deflection (D)	$\times 1$
	Exit end of plates to screen distance (L)	$\times K$
	Separation of deflector plates (d)	$\times K$
Subsidiary operations to preserve constant "closeness to plate interception" at constant deflection	Deflector voltage (V)	$\times K^{-2}$
	Plate length (l)	$\times K^2$
Consequences	Deflection defocusing (ΔD)	$\times K^{-2}$
	Plate sensitivity	$\times K^2$

movement. Hence our third and final principle is reached:

Principle (3)

"The deflector plates should be constructed to give maximum sensitivity."

5. Attainment of Maximum Deflector Plate Sensitivity

Theoretical discussion of the plate profile to achieve this has been given by Maloff and Epstein[6]. However, the author has found that the far simpler procedure illustrated in Fig. 4 gives a plate profile virtually indistinguishable from the optimum. Here ZZ′ is the tube axis. S_2 is the plane of the screen. XY normal to ZZ′ represents the position of the entry edge of the deflector plates. X′Y′ is the position of the exit edge. I′ is the position of maximum spot deflection at the screen. Draw the straight line I′T where YT=TY′. This cuts X′Y′ at N. Erect the normal to I′T through N. This cuts XY at K which is the centre of curvature of the plate. The profile so defined is pulled away along a normal to the tube axis by an amount just in excess of s (where $2s$ is the beam width) so as to avoid beam interception. Experiments made by the author have shown that the above method results in a plate profile giving maximum sensitivity within the error of observation.* If, for example, a plate so designed is dusted over with a thin coating of some fluorescent powder, it will be observed that when the beam strikes, it does so uniformly along its whole length. Clearly this is indicative that maximum sensitivity for a given plate length has been achieved. Any seeming discrepancy between this technique and the analysis indicated by Maloff and Epstein is resolved by the fact that their theoretical curve very closely approximates a circle except for exceedingly long deflector systems. It must, of course, be remembered also that their analysis involves certain assumptions (including neglect of the fringing fields) which to some extent vitiate the exact truth of their working.

* This construction has the following basis in theory. The beam trajectory in a parallel plate system is parabolic. A tangent to this parabola at the exit plane intersects the tube axis half way along the plates. This provides justification for setting YT=TY′.

6. Effect of the Second Axis of Deflection

If the tube were to have only one axis of deflection, the preceding analysis shows that the best deflection plate system would be a very long one, filling almost all the space between the electron gun proper and the fluorescent screen. For this very long plate, the approximate construction given for optimum sensitivity

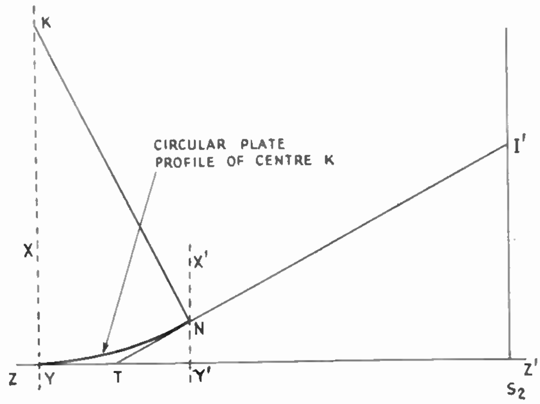

Fig. 4. Geometrical construction for obtaining optimum plate profile.

would be in error and it would be necessary to employ the more exact methods of Maloff and Epstein[6].

In practice, of course, two orthogonal axes are invariably needed. The length of the first set of deflectors is thus limited by the consideration that they must not encroach too much on the space to be occupied by the second set. The choice of the relative length of the two plate systems cannot therefore be determined without a precise specification for the tube. In practice, it is not usual for the length of the first set of deflectors to exceed about 2 in.

7. Conclusions

The principles of the design of optimum electrostatically-deflected tubes have been stated. They result in tubes of the general appearance indicated in Fig. 5(*b*). The gun is fat and stubby. The deflector plates are very large and the tube neck diameter correspondingly so. Fig. 5 (*a*), on the other hand, indicates the general appearance of a rather poor system. The gun is long and rather thin and the deflector plates quite small.

The historical aspects of this matter are of some interest. The first tubes designed for high performance after the manner indicated in Fig. 5 (*b*) were, to the author's knowledge, made by the General Electric Company in Wembley, England, around 1939. Certainly their designers had a good understanding of the advantages which were given by this type of construction although the author is not aware of whether they had at that time formalized their reasoning to the extent set forth in this paper. In the intervening years this type of design has been widely adopted in Europe, but only recently has it been taken up in America.

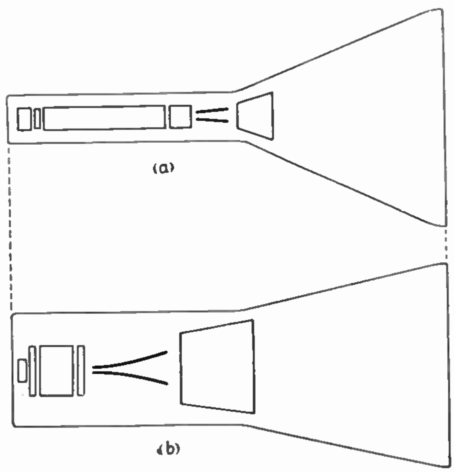

Fig. 5. Diagrammatic sketches of (*a*) low performance tube and (*b*) high performance tube.

8. Acknowledgment

The author is much indebted to Dr. J. W. Coltman of the Westinghouse Central Research Laboratories in Pittsburgh who read the original manuscript and suggested significant improvements.

9. References

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APPLICANTS FOR ELECTION AND TRANSFER

As a result of its July meeting the Membership Committee recommended to the Council the following elections and transfers.

In accordance with a resolution of Council, and in the absence of any objections, the election and transfer of the candidates to the class indicated will be confirmed fourteen days after the date of circulation of this list. Any objections or communications concerning these elections should be addressed to the General Secretary for submission to the Council.

Direct Election to Full Member

O'BRIEN, William Joseph. *Leatherhead.*

Direct Election to Associate Member

DUMMER, Geoffrey William Arnold, M.B.E. *Great Malvern.*
GASKELL, Major Waryn Thomas Mogridge, B.Sc., Royal Signals. *Germany.*
HARI SINGH, Lt. Col., M.Sc., B.Sc., B.A., Indian Corps of Signals. *Jubbulpore.*
LOADER Dennis, B.Sc. *Weston-super-Mare.*
MASON, William, B.Sc. *Dinas Powis.*
POWELL, Claud. *New Malden.*
SIMPSON, David, Dip.Eng. *Camberley.*
VARLEY, William Eric Clifford. *Sanderstead.*

Transfer from Graduate to Associate Member

BARTON, Martin Leonide, Dip.El. *Stamford, Conn., U.S.A.*
DEAR, Leslie Donald. *North Hillingdon.*
DODD, Ivor David, B.Sc. *Pontypridd.*
LAMBERT, Lionel Edward. *Stevenage.*
PASSMORE, Sqdn. Ldr. Patrick Martin, R.A.F. *Henlow.*
PHILLIPS, James Hugh. *Cookham.*
WITHERS, William Christopher Robin. *Hampton Hill.*

Direct Election to Associate

DAWANCE, Eugene Julian. *Paris.*
GLEDHILL, Ronald. *St. Helens.*
KITCHEN, Ronald. *Braintree.*
SCOTT, Major Malcolm Douglas, Royal Signals. *Singapore.*
STONER, Paul Henry. *Barkingside.*
URWIN Robert Mather. *North Shields.*
YARNOLD, William Bertram, B.Sc. *Malvern.*

Direct Election to Graduate

DAVIS, Colin. *Ruislip.*
HANSFORD, Douglas John. *Bridport.*
KEEP, Brian Dennis, B.Sc.(Eng.). *Surbiton.*
STEWART, Jeremy Ian. *Huddersfield.*
WHITMORE, Dennis Ainsworth, B.Sc. *Carshalton.*

Transfer from Studeut to Graduate

CAMPBELL, Gordon Jarvis. *Oslo.*
MALHOTRA, Plt. Off. Shadi Lal., B.Sc., M.Sc., I.A.F. *New Delhi.*

STUDENTSHIP REGISTRATIONS

AHMED, Ahtesham. *Cambridge.*
ALWANI, Kishin Kakumal. *Bombay.*
AVAVIND, K. P. *Madras.*

BAMGBOYE, Christopher Olumuyiwa. *London, N.13.*
BORGES, Rev. William Joseph, S.J., B.Ph., B.Th. *Southampton.*
BROOKES, Malcolm John. *Heston.*
BUTTON, Lieut. William Gordon, R.N. *Dover.*

CHATTAWAY, Gary Roland. *Nuneaton.*
COX, Ronald. *Sheffield.*

DARE, Isaac Oluwole. *Ibadan, Nigeria.*
DUNKERLEY, Robert Selwyn. *Coventry.*

GRIFFITH, Geraint Wynne. *Southampton.*

IUSTER, Nelu. *Hadera, Israel.*
JOSHI, Ganesh Bahirao. *Bombay.*

KONKAR, Vishwanath Anant. *Poona.*

LAWTON, David Harold. *Oldham.*

LOVEWELL Stanley Sydney. *Newport, Mon.**

McAVOY, James. *Walton-on-Thames.*
McGARRY, Douglas Wilfrid. *Woodford Green.*
MILAS, Athanasios. *Athens.*
MORGAN, Bernard John. *Tredegar.*

O'CONNELL, Terence Kevin. *Alresford.**
OWEN, Irving Percival Davison. *Bromley.*

PANTAZIS, Dimitrios. *Athens.*
PANTHALKAR, Vasant Rao. *Bangalore.*

RAJPUT, Dhir Singh. *New Delhi.*

SPAILE, Edgars Viktors. *Bass Hill, New South Wales.*
STICKLEY, Michael Peter. *Nairobi.*
SURTI Kantilal Khushaldas, B.E. *Bhilai.*

TAYLOR, John Keith. *Chelmsford.*

VACALIS, George. *Athens.*

WIESENFELD, Josef. *Ramat-Gan, Israel.*

* Reinstatement.

INVESTIGATION OF HORIZONTAL DRIFTS IN THE E REGION OF THE IONOSPHERE IN RELATION TO RANDOM FADING OF RADIO WAVES*

by

B. Ramachandra Rao, M.Sc., D.Sc.† and M. Srirama Rao, M.Sc., D.Sc.‡

SUMMARY

Regular determinations of E region wind velocities (V) and frequency of fading (N) were made from simultaneous fading records at three spaced receivers taken on two pulsed radio frequencies, 2·3 and 2·8 Mc/s respectively, between December 1954 and March 1955. A linear relation between V and N was obtained for each wavelength, and the experimental relation $V = 1{\cdot}86\ N\lambda$ deduced. This result agrees with theory and the significance of the constant of proportionality is discussed.

When all possible causes of periodic fading are eliminated, it is usually observed that a pulsed radio wave reflected from the ionosphere undergoes fading which is found to be random. Ratcliffe[1] has presented a theory for this random fading based on the assumption that the ionosphere is an irregular reflector in which the irregularities are assumed to be in continual random motion. Subsequent work has shown considerable evidence of the presence of regular and systematic horizontal drifts at ionospheric levels and radio methods have been evolved for measuring these drift velocities. Making certain plausible and simplifying assumptions on the relationship between the random and systematic horizontal drifts in the ionosphere, McNicol[2] has obtained the approximate relation

$$V \cong 2{\cdot}8\, N\lambda \qquad \ldots\ldots\ldots(1)$$

where V is the drift velocity, N the frequency of fading and λ the wavelength of the radiation. A detailed theoretical study of the diffraction of a radio wave by an irregular ionosphere having random or systematic horizontal drifts has also been made by Booker, Ratcliffe and Shinn[3]. Briggs, Phillips and Shinn[4] have discussed the nature of fading due to random and systematic horizontal drifts and have shown that these two causes of fading may be separated by a suitable analysis of fading records obtained at three spaced receiving points on the ground. From the results of such experiments it is believed that a horizontal movement of the reflecting layer is the main cause of fading. Accepting this view Briggs[5] deduced the relation

$$N = \frac{2V}{\lambda} \sin \theta_0 \qquad \ldots\ldots\ldots(2)$$

where θ_0 is the semi-angle of the cone of the downcoming rays. It will be noticed that except for the constants, this relation is the same as relation (1) given by McNicol.

Studying the scintillation of radio stars at three spaced receiving points simultaneously, Maxwell and Dagg[6] determined ionospheric drift velocities in the F2 region and showed that over the range of velocity of 30–100 m/sec, the fading rate N and drift velocity V are linearly related, but they have not made any attempt to study the dependence of V and N on the wavelength of the radiation.

In the present investigation the authors have attempted an experimental verification of the linear relation between V and $N\lambda$ expected theoretically by measuring the horizontal ionosphere drifts and speeds of fading at two different radio wave frequencies.

The ionospheric drift measurements were made by the spaced receiver method of Mitra[7] after slightly modifying the technique by using

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† Ionosphere Research Laboratories, Physics Dept., Andhra University, Visakhapatman, South India.

‡ National Research Council, Canada; formerly at Andhra University.

U.D.C. No. 551.510.535.

a three beam oscillograph for recording the fading patterns. The full details of the equipment used in this laboratory as well as the method of analysis of records for drift velocities were given elsewhere[8]. The frequency of fading N in each record has been determined by the simple method given by Rice[9]. The records were taken on two different frequencies of 2·8 Mc/s and 2·3 Mc/s giving reflections from the E region during the month of December 1954 to March 1955. All the records were taken simultaneously during daytime between 0900 and 1600 hours I.S.T. The various values of the drift velocities are then separated into groups falling in the ranges of 0–10, 10–20, 20–30, etc. m/sec and the average values of drift velocities and fading frequencies in each group are estimated. As the most probable drift velocities are in the range 30 to 60 m/sec, there are fewer observations having very low or very high velocities and as such the average values of V and N for these ranges are less precise.

The values thus obtained for the average velocities V and fading frequency N in each group for records taken on wave frequencies 2·8 and 2·3 Mc/s are presented in Table 1.

The observations on 2·3 Mc/s are comparatively less in number as the signal strength is much less on this frequency. Fig. 1 shows a plot of these observations and it will be noticed that nearly all the points for each frequency lie nearly on a straight line passing through the origin, thus confirming the linear relationship between V and N expected from theoretical considerations. The most interesting feature of this graph is the fact that the gradients of the straight lines for the observations on the two frequencies are different, being less for the lower frequency. From the relations (1) and (2), it can be easily seen that the gradients of these straight lines given by $60\,N/V$ should be a linear function of the frequency of the wave. Hence the ratio of the gradients of the straight lines in Fig. 1 should be the same as the ratio of the corresponding wave frequencies if the theoretical relations (1) and (2) are valid. The experimentally determined value of the ratio of the gradients of observations on 2·8 and 2·3 Mc/s waves is found to be 1·26 and this value compares well with the value of 1·22 obtained for the ratio of the two wave frequencies. Considering the small difference between the two ratios as due to experimental errors, this agreement can be regarded as a confirmation of the expected linear relation between V and N. Taking the mean value of the constant of proportionality determined from the two graphs, we can write the final relation as

$$V = 1{\cdot}86\,N\lambda \qquad \ldots\ldots\ldots(3)$$

Using relations (2) and (3), the angle of the spread θ_0 of the downcoming waves is calculated and is found to be 7·72 deg. This value is in agreement with the value of 7·2 deg., determined by Briggs[5] by following an entirely different method involving the study of fading records showing amplitude variation with frequency over a 1 Mc/s range. This value of θ_0 determined by the present authors is also in

Table 1

Sl No.	Wave frequency 2·8 Mc/s V in m/sec	60 N in c/min	Wave frequency 2·3 Mc/s V in m/sec	60 N in c/min
1.	19	7·0	30	8·0
2.	29	8·0	44	12·7
3.	35	12·5	53	13·3
4.	45	13·8	72	16·7
5.	55	17·5	85	20·7
6.	64	21·7	110	25·1
7.	72	22·2		
8.	80	23·2		
9.	96	28·0		
10.	107	31·7		
11.	115	35·2		

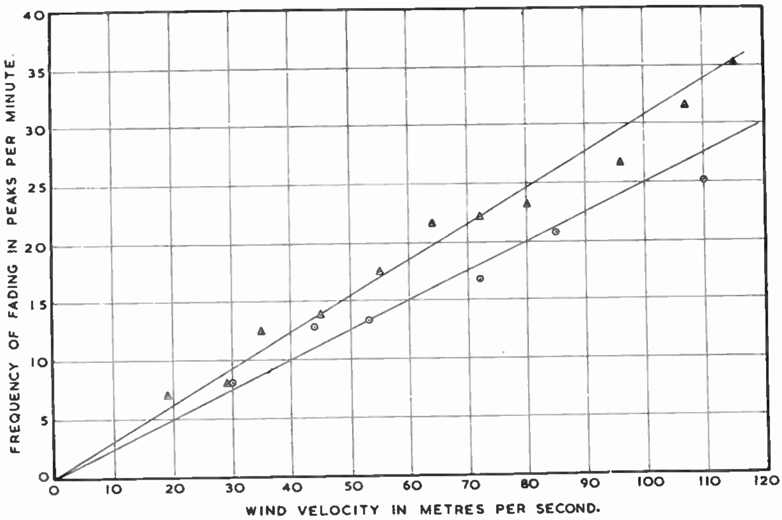

Fig. 1. Variation of frequency of fading with drift velocity.

agreement with the value obtained by Briggs and Phillips[10] from observations of fading at spaced receiving points.

In conclusion, it may be mentioned that the confirmation of the linear relation (3) supports the idea that the random fading of pulsed radio waves reflected from the ionosphere is mainly due to horizontal systematic drifts of the layer, the fading due to random changes being less important.

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INSTITUTION NOTICES

Birthday Honours List

The Council of the Institution has congratulated the following Member whose name appeared in the Birthday Honours List. The names of other members included in that List were published in the June *Journal*.

Harold Owen Ellis (Member) was appointed a Commander of the Most Distinguished Order of St. Michael and St. George.

Colonel Ellis has recently been appointed Postmaster General of East Africa; he was previously Director of Posts & Telegraphs in Lagos, Nigeria, and was a Postmaster General in Nyasaland. In 1952 he received the O.B.E. for his services to telecommunications in Nyasaland.

Institution Visits

The second visit arranged this year by the Technical Committee took place on July 24th last, when a party of members was shown over the B.B.C.'s Technical Receiving Station at Tatsfield.

The Station has a very wide range of duties, and these include regular measurement of frequency and modulation depth of B.B.C. home and external services, as well as television and v.h.f. transmissions, and identification and measurement of signals interfering with B.B.C. transmitters. The Station also carries out important measurement and recording work in connection with ionospheric investigations for the I.G.Y. (These were referred to in a discussion contributed by the Engineer-in-Charge, Mr. H. V. Griffiths, published in the July issue of the *Journal* (page 412).)

The equipment of the Station includes both omni-directional and directional arrays of various types, frequency standards and measurement equipment, field strength measuring sets, direction finders, atmospheric noise measuring equipment, and numerous communications type radio receivers, including special triple diversity receivers for "exalted carrier" and single sideband reception.

The Technical Committee is most grateful to the Engineer-in-Charge and his staff, and to the B.B.C. Engineering Information Department, for the excellent arrangements made for the visit.

As the applications for this visit exceeded the number which could be accommodated, the B.B.C. has kindly agreed to receive another party later on in the year should sufficient support be forthcoming. Members wishing to be considered for such a visit are invited to write to the Institution. Other visits planned for later this year include the factory of Vauxhall Motors Ltd. at Luton on December 4th.

City and Guilds of London Institute Annual Report

A continued increase in the total number of candidates accepted for its Examination is shown in the 1957 Annual Report of the Department of Technology of the City and Guilds of London Institute which has recently been published. This increase is particularly marked in the Telecommunications Engineering Group, the total number of entries received for all the thirteen papers being over 44,000, which is 5,000 more than in 1956; however, the number of full Technological Certificates issued in this subject remains at the same comparatively low level of 80.

An analysis of the entries shows that the numbers of candidates entering for the final year's subjects, Telecommunications Principles Grade V, and Radio Grade IV were 303 and 366 respectively.

Details are also announced of the changes in the Telecommunications Group, which will be succeeded by a course designed for Telecommunications technicians. These changes will affect the acceptance of examinations recognized as exempting from the Institution's Graduateship examination. Further advice on the changes will be given in the Annual Report, which will be published in October.

Circulation of the Journal

As members will be aware, the Institution's *Journal* is one of the few technical journals in the radio and electronics field whose circulation is certified by the Audit Bureau of Circulations. The certificate which has just been published for the first six months of 1958 shows that the average circulation per issue was 6,912. This does not, of course, include sales and other circulation since June.

TELEMETRY AERIALS FOR HIGH-SPEED TEST VEHICLES*

by

R. E. Beagles (Associate Member)†

SUMMARY

The telemetry services in use are mentioned and the problems peculiar to propagation of these signals to and from high-speed test vehicles are discussed. A survey of typical external and suppressed aerials is made. The methods of testing commonly employed are described.

1. Introduction

During the last decade it has become the practice in the aircraft industry to use a series of models (known as Test Vehicles) to evaluate the performance of guided missiles and supersonic aircraft. These models vary in size from some fractional value to full scale, which may mean from a few feet long to the size of a small aeroplane. Fitted with motors or booster rockets they can be ground or air-launched at ranges in the U.K. and Australia.

The purpose of the tests varies throughout a project, for which there may be more than one size of vehicle, and will extend from initially establishing aerodynamic parameters to control and guidance experiments. Measurements within the vehicle can only be made by radio methods whilst in the observation of trajectory radio extends and often supplants optical tracking. Experiments are expensive and it is important that the maximum information is obtained.

Radio methods are used to observe any quantity in the vehicle by using a 24-channel telemetry system[1], to measure velocity by Doppler techniques[2], and to record trajectory by the use of an airborne marker beacon. In addition the control, landing or destruction of the vehicle can be carried out via a radio link.

It is not often possible, or even desirable, to put all these services in for one test, and so each experimental firing is intended to solve some specific feature of a programme. Because the telemetry system is itself not under test, the utmost system reliability must be maintained to prevent loss of signal and hence invalidation of an experiment. Not only does the equipment therefore have its own power supplies, but the design and installation is carried out separately by a group of specialists. This work is never obvious in the finished product. In this paper it is hoped to cover one aspect of this field, that is, the provision of aerial radiating systems for all forms of telemetry.

2. General

A radio aerial has been defined[3] as the structure associated with the region of transition between a guided wave and a free-space wave or vice versa. The characteristics of an ideal aerial could be listed as follows:—

(i) it should transfer maximum energy with minimum loss.

(ii) the energy should be directed in any desired direction.

(iii) it should have a wide frequency coverage.

(iv) it should be polarized in any desired plane.

(v) it should have a small physical size suitable for mounting in any location.

In practice the first point will be covered by having the aerial resonant and matched to the transmission source. The direction of the radiated energy will be dependent on the radiation pattern required which also determines whether any power gain and increase in range can be obtained by making the aerial directional. With a wide bandwidth one type of aerial can be used for any of several frequencies, whilst there may be a minimum bandwidth if a transmitted pulse shape is to be retained. The polarization limitation does not normally hold in this instance for receiving aerials are circularly polarized; occasionally,

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U.D.C. No. 621.396.677:623.4.084

however, measurement of vehicle roll is carried out by observing the change of plane of polarization. The physical construction calls for a special approach and accounts for a large proportion of the difficulties met with in this work. Although techniques established for aeroplanes are still followed, the thin wings and narrow fuselage intensify the problems of connecting feeders and fitting aerial inserts. Airframe designers are reluctant to incorporate large portions of unstressed insulating materials. Fortunately most telemetry systems are in the u.h.f. or microwave band leading to small and more easily resonated units. Since the ideal aerial does not exist, its practical realization is a compromise, a detailed consideration of which follows.

3. The Radiation Pattern

On a sea range or over short distances most telemetry systems need only radiate rearwards. For long ranges there may be receiving stations down range requiring forward radiation and, considering also vehicle roll, the logical extension is an omnidirectional radiator. A straight and level flight cannot be relied upon and it is at the time of erratic behaviour that the telemetry is most important to ascertain the cause. Unfortunately, no omnidirectional radiator is available, for however it is mounted there will be shadow from some part of the supporting structure, although this may to some extent be overcome by doubling up on the installations.

The safety aspect must also be remembered for it is necessary to be able to destroy the vehicle in any attitude and even to know exactly where it is! This may sound surprising but with small models and a conglomeration of large booster rockets jettisoned during flight, radar tracking can be mislead (Fig. 1).

The radiation pattern is therefore most important and warrants careful study. There will be three conditions to consider[4]; where a wavelength is long compared with the test vehicle dimensions, of comparable length, and where it is short. All three conditions will be apparent in most projects at some stage. With the wavelength short, the vehicle appears as an infinite ground plane giving a reasonably predictable pattern, but the effects of shadow from the structure are severe. Conversely, with the wavelength relatively long, energy flows round the obstacles but the actual construction of the aerial can be awkward. When the wavelength is of comparable length, dimensions of wings and fins become fractions of a wavelength, leading to large induced currents in leading edges, etc. This in turn leads to parasitic radiation which can be in, or out of, phase, with the original signal, rendering calculation of the pattern almost impossible. The patterns then have to be determined by measurement.

Fig. 1. Experimental firing of guided weapon "Sea Slug". The photograph shows the jettisoning of the four booster rockets. (By courtesy of Armstrong Whitworth Aircraft.)

Fig. 2 shows patterns obtained during a test with two short spike aerials mounted diametrically opposite on the rear of the vehicle fuselage. In (*a*) where the wavelength is very short, shadow occurs to the front of the vehicle. With the wavelength longer (*b*), the pattern approaches the theoretical figure of eight. With the wavelength long compared with the overall dimensions (*c*), the spike aerial is now very inefficient and an appreciable amount of radiation comes from the fuselage itself.

When establishing such patterns it is as well to remember the final structure undergoing tests. Vehicles are often launched from aircraft and telemetry information is required

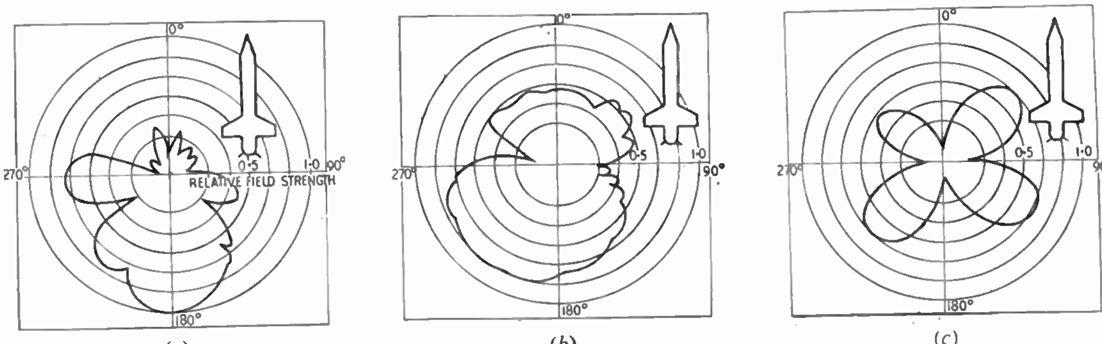

Fig. 2. Change in radiation pattern with variation of wavelength.

(*a*) Wavelength less than one fifteenth of the vehicle overall dimensions. (*b*) Wavelength less than one fifth of the vehicle overall dimensions. (*c*) Wavelength greater than the vehicle dimensions.

prior to, and at the time of launch. Obviously the vehicle aerials will be affected by the parent aircraft. Again the booster rockets may be arranged in many forms and can lead to almost complete aerial screening.

The position of the aerials must be decided upon by a compromise between the structural designer, the stress man and the aerodynamicist. It is desirable to have several types of aerial of the same frequency suitable for different location. An attractive situation on a leading edge may be eliminated by very high temperatures at this point (hundreds of degrees centigrade). It must also be borne in mind that these are experimental vehicles to establish structural design parameters and a rapid solution is required. It is necessary for a decision to be reached early in the project so that structural alterations can be incorporated in the finished airframe. In these circumstances the installation may be far from ideal from the aerial designer's point of view.

4. Types of Aerial

It is possible to divide those in use into two broad categories, aerials external to, and those suppressed within, the structure.

4.1. *External Aerials*

Of the external aerials, trailing wires, mast supported, and whip aerials are impracticable because of flutter, drag and similar mechanical weaknesses, but the short spike is suitable. This is attractive because it involves only minor modifications to the structure and the pattern can be reasonably predicted.

It is therefore necessary to decide whether there is a structurally sound unit which will withstand the extreme conditions of acceleration, vibration and strain, but will not upset the validity of the experiment in any way. The spike should be about a quarter of a wavelength long so that it is resonant, and with a carefully designed base the input impedance can be varied about a fifty ohm point by adjusting the length of the spike. Accelerations of 50*g* are experienced in ground-launched models whilst roll will put a transverse strain on the aerial so that bending moments at the base must be carefully calculated. With these points in view a conical spike of circular cross-section is preferable, and simple to manufacture.

Fitting the aerial on the vehicle immediately increases the drag and this is prohibitive where the total from all aerials exceeds ten per cent. of the total vehicle drag. Some improvement is gained, at a sacrifice of electrical efficiency, by reducing the overall length of the spike. It is also possible to shape the spike into a wedge, rhombic or aerofoil cross-section but this inevitably increases the base mounting area, giving a larger input capacitance. More important, however, is the need now for accurate alignment on small vehicles where the aerial forms an appreciable fraction of the total control surface. Misalignment will cause the vehicle to spin or roll sometimes at several revolutions a second. Finally, the handling of the vehicle during the launching period should be considered for potential hazards to personnel.

Figure 3 shows a spike aerial designed for use at speeds up to Mach 2.

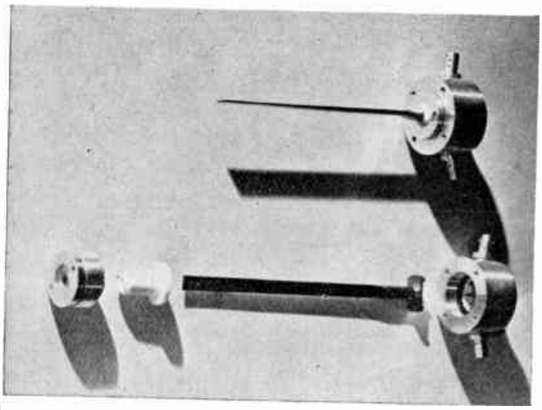

Fig. 3. Missile "spike" aerial suitable for speeds up to Mach 2. (By courtesy of Armstrong Whitworth Aircraft.)

4.2. *Suppressed Aerials*

The simplest way to suppress the aerial is to bury it in an insulated portion of a wing or fin. On the small vehicles a complete stabilizing fin machined out of Tufnol (or similar material) with a metal radiating element inlaid is possible. This is occasionally done, but in most cases the fin is so thin that the material is not rigid enough to withstand vibration or flutter. On larger vehicles where the forces involved prevent the use of a complete surface, a portion may be replaced with insulant. The wing tip aerial popular in aircraft is an example, but again with the wings being thinner the mechanical difficulties of joining, strength, rigidity, etc., restrict their use.

The slot provides a very good basis for a number of different types of aerial. Basically a half-wave rectangular slot is cut in a metal sheet and a connection by transmission line is made across the slot. The impedance at the centre is about five hundred ohms and falls to zero at each end of the slot. A convenient match for fifty-ohm cable is made as shown in Fig. 4 by feeding at a short distance from either end. Currents flow round the slot spreading out over the sheet, and radiation occurs on both sides of the sheet. In an infinite sheet there will be zero current at the edges giving the familiar figure-of-eight pattern of the half-wave dipole. The comparison with the dipole is exact when the direction of the vibration of the electric and magnetic fields is interchanged[5, 6], in other words a horizontal slot is vertically polarized. With the sheet less than infinite, appreciable currents may reach the edges and diffraction takes place at the transition from metal to dielectric. Thus in practice, for a small surface (in terms of wavelength), the radiation pattern is greatly affected by the shape of the surface.

In these circumstances the slot should be considered as an impedance transformer into the radiating structure. Folding changes the centre impedance but may lead to a more con-

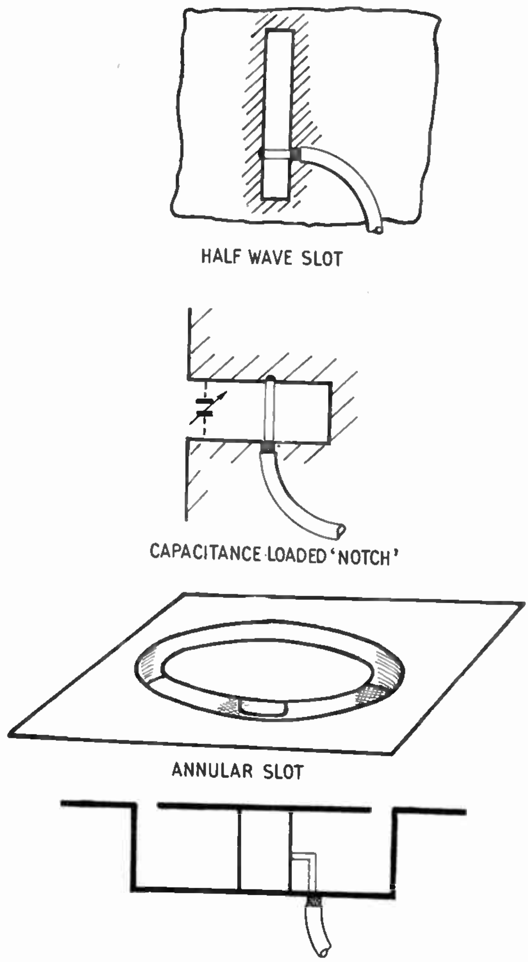

Fig. 4. Suppressed aerials. Possible slot configurations.

venient shape and a further reduction in size can be gained by capacitance or dielectric loading. Dielectric filling is necessary any way to give minimum aerodynamic drag but this does reduce the aerial bandwidth. For radiation on one side as would be required in a fuselage mounting, the slot must be boxed in on one side by a quarter-wave rectangular cavity. With minimum dimensions this introduces an additional reactance across the slot, so that for the same frequency of operation it must be longer than a half wavelength.

An obvious application for this type of aerial is for microwave transmissions. When waveguide fed it has advantages from a temperature and drag point of view over a waveguide terminated in a dielectric lens aerial (Fig. 5). However, the large dimensions involved virtually preclude this device from use at frequencies much less than 500 Mc/s.

An immediate reduction in size is gained by the use of a notch aerial developed by Johnson[7]. It consists of an open-ended slot, or notch, cut in the edge of a wing and should be a quarter of a wavelength long. Matching is again achieved by feeding part way up the notch. It is made much shorter physically by capacitive loading at the open end, subject to the same bandwidth limitations, thus making a very convenient unit to use. The capacitance is adjusted to tune the notch to resonance incidentally taking up any production variations. The feed point is a little more critical and is best found by experiment. Built round

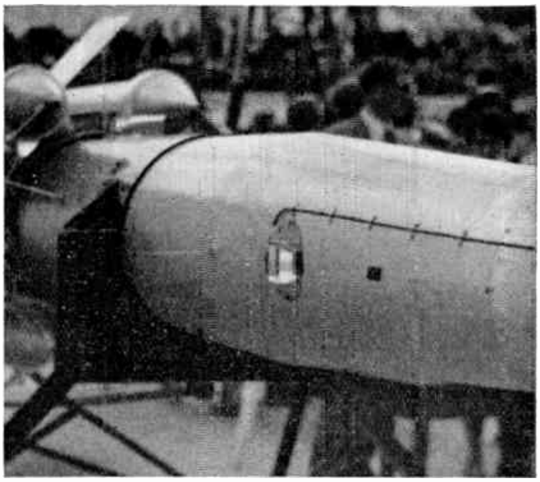

Fig. 5. Typical telemetry installation.

a glass laminate cover the unit can be set up on the bench and then placed *in situ*. The pattern is still dependent upon the structural shape for small surfaces, but is like a unipole on the edge of the sheet when the sheet can be assumed infinite. Plane of polarization is along the major axis of the sheet.

Figure 6 shows the pattern obtained from notch aerials operating at the same frequency but mounted in different positions. Fig. 7 shows the impedance characteristics of a notch 3 in. long.

Another form of slot is the circular diffraction antenna or annular slot[8]. This, as its name

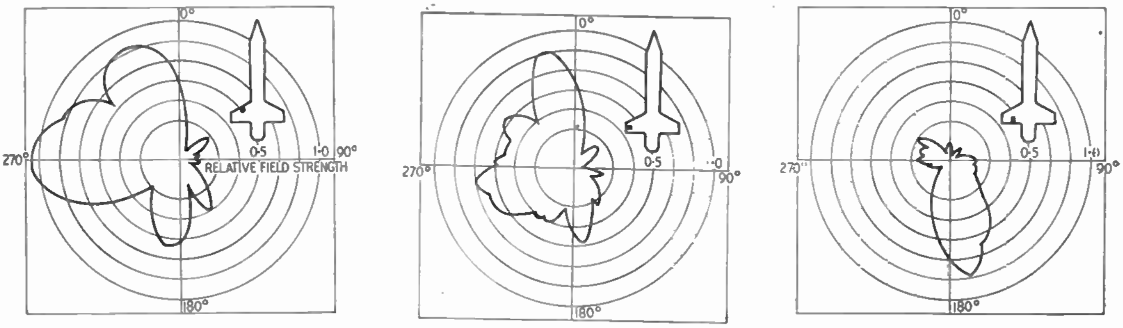

Fig. 6. Change of pattern produced by reposition of the aerials.
(*a*) Notch aerial in the wing leading edge. (*b*) Notch aerial in the wing tip. (*c*) Notch aerial in the wing trailing edge.

implies, consists of a circular slot cut in a metal surface and backed by a shallow cavity. The radiation pattern is determined both by the slot diameter and width of the gap. It is possible, with suitable dimensions, to obtain a multi-lobed pattern, but for a single lobe with minimum directivity (i.e. the nearest approach to an isotropic source) a small slot diameter in terms of wavelength and a large gap are best.

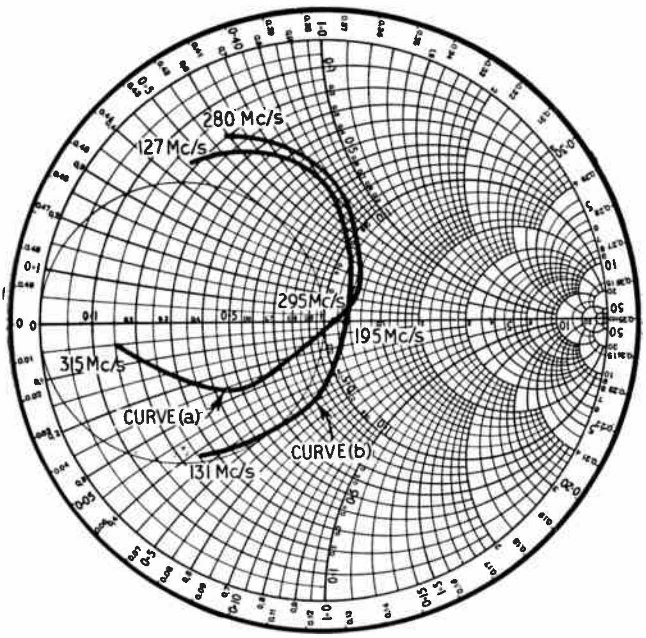

Fig. 7. Impedance characteristics of a 3-in. notch aerial.

(*a*) Notch with 4 pF capacitive loading.
(*b*) Notch with 26 pF capacitive loading.

Thus a slot of 0·22λ diameter and a gap of 0·016λ has a pattern similar to a unipole and is polarized at right-angles to the plane of the aerial. Maximum bandwidth is obtained with a slot diameter of 0·8λ but a bandwidth of 2 per cent. of the resonant frequency obtained with the smaller slot is acceptable for most purposes. For radiation from one side only the slot must be backed by a shallow cylindrical cavity. This is treated as a radial transmission line short-circuited at its inner end and dimensioned so that the line susceptance matches the susceptance of the gap. This usually varies when a dielectric covering is used and filling the cavity with a dielectric for rigidity also introduces a change requiring final adjustment to be made by measurement. Matching of the input cable can be accomplished by any of several ways and Fig. 4 shows a simple method. The input position along the radial line is adjusted for the 50-ohm point. Because of the contained surface currents the radiation pattern of this aerial is dependent on its location to a lesser degree than the previous types.

4.3. *Ground Aerials*

Although not strictly within the context of this paper, a brief reference is made to the ground aerials for general interest. As previously suggested, these must be circularly polarized, and preferably have a high gain. The helix[9] is ideal for this application being simple to design and, at these frequencies, simple to manufacture. The illustration (Fig. 9) shows a test helix of ten turns with a gain of 15db, a beamwidth of 36 deg., and a terminal impedance of 140 ohms.

5. Matching

The matching is best carried out on the aerial itself and has largely been covered in previous sections. When such matching is impossible recourse must be made to a quarter-wave transformer or stub matching. This is to be avoided for, apart from the added complication and increase in the time taken to prepare a test vehicle for firing, the aerial will only be matched at the one (quarter-wave) frequency. The importance of good matching has been continuously stressed. If the aerial is mis-matched the oscillator may be pulled off frequency. The aerial becomes particularly sensitive to adjacent objects such as the launching structure. The transmitted signal may therefore change on firing, leading perhaps to complete loss of signal.

6. Cross Modulation

Frequently up to a dozen or more aerials have to be mounted on a vehicle and with the inevitable proximity of the aerials and transmitter frequencies, cross modulation can take place. The major causes will be radiating feeders due to a poor match, surface currents between adjacent suppressed aerials and direct radiation between aerials. Good matching is essential as mentioned elsewhere and the intermingling of surface currents is just one further point to consider when positioning the aerials.

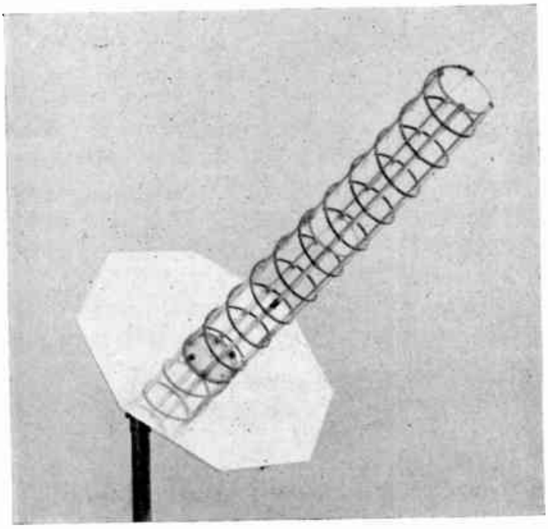

Fig. 8. A ten-turn helix aerial.

This also applies to direct pick-up, some rejection being obtained with a control surface interposed in the path of maximum radiation. Crossed polarization of aerials (i.e. vertical next to horizontal) will give a further reduction of up to 20 db. As a final expedient a coaxial line rejection filter can be inserted in the aerial feeder.

7. Design and Testing

Ideally the design goes through the following procedure. Early in the project, the communication requirements are decided upon, with their appropriate frequencies. The general structure of the proposed vehicle is examined and a number of alternative schemes prepared. The most promising from a structural point of view is then tested. Designs following an established pattern for which there is some information available are proceeded with first. If these are not suitable, resort is then made to aerials which take advantage of any peculiarity of the structure.

The radiation pattern is first obtained. If possible, full-scale vehicles are used, but where this cannot be done because of large size or if outdoor work is prevented for security reasons, scale model techniques may be employed. A brass model is made to some appropriate scale of a tenth, twentieth, etc., and the wavelength is scaled down by the same factor (i.e. the frequency is scaled up). Reflex klystrons provide transmitters of convenient frequency and sufficient power to be used with a simple crystal detector, over the short range required. The scaling factor is therefore chosen to suit these frequencies taken together with a consideration of the practicabilities of making the models. The model, with aerials, is then mounted in a radio darkroom constructed of radio frequency absorbing material (Fig. 9). With the model on a turntable and an automatic plotter, a rapid assessment can be made and modifications to the aerials carried out accordingly. Patterns will be recorded in at least three planes whilst a three-dimensional picture can be created out of wire or Perspex. It is, of course, only necessary to get sufficient energy from the aerials for measurement purposes, matching being impossible without scaled-down cable and stray capacitance.

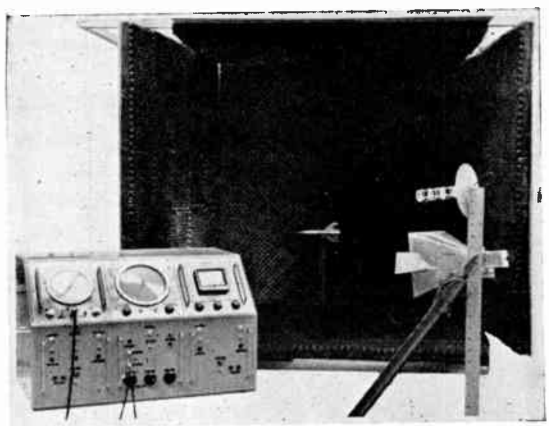

Fig. 9. Radio darkroom showing the automatic pattern recorder, the scale model test vehicle and the klystron transmitter with horn aerials.

Models can give quick results, but their potentialities are limited when it comes to suppressed aerials. Although some work may be possible with the complementary unsuppressed type (substituting a short unipole for a notch or annular slot), it is often preferable to test the aerial itself.

Radio darkrooms are restricted to the higher frequencies so that at lower frequencies a large test area is required, at least 10 wavelengths in radius, clear of all obstructions. This invariably means outdoor work with all the attendant restrictions caused by the weather. A test site is usually maintained with the necessary equipment for raising and rotating the specimen and suitable power supplies for field strength measuring receivers.

Matching must be done on an actual size aerial surrounded by a piece of structure to an extent of a few wavelengths. Many laborious measurements then have to be made with u.h.f. bridge, reflectometer, or a slotted line. One problem, which may have been surmised, is to obtain actual structural specimens sufficiently early to carry out tests. This is a continual difficulty but can be partially overcome by constructing sections of wood. These are then sprayed with copper giving a good conducting surface to which connections are easily soldered.

The final test comes during the first flight of a new aerial system during which a continuous pen record is taken of signal strength with superimposed timing pips. This will immediately reveal whether the desired safety margin has been maintained, but will also provide other interesting information. An irregularly shaped, or poor radiating pattern will lead to a record oscillating in amplitude during vehicle roll. As the flight terminates multipath propagation due to reflections from the ground, will be observed. Complete loss of signal, if not due to power supply failure as shown on the telemetry record, is then correlated to vehicle attitude with the help of the timing pips and the trajectory record.

8. Conclusions

It is regretted that more specific information has not been given, but for obvious reasons this has had to be withheld. A paper of this length and nature can therefore hope to be little more than an introduction to the subject. Detailed information on specific aerials is contained in the references.

9. Acknowledgments

The author wishes to thank Messrs. A. V. Roe & Co., Ltd., for their kind permission to publish this paper.

10. References

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INTERNATIONAL CONFERENCE ON MEDICAL ELECTRONICS

THE idea of an international conference on medical electronics was first put forward by Dr. V. K. Zworykin, director of the Medical Electronics Centre of the Rockefeller Institute for Medical Research, in a letter to the Brit.I.R.E. in November 1957.

The Council of the Institution agreed to send a delegate to the first planning conference, which was held at La Nouvelle Faculté de Medecine, Paris, from June 26th-28th, 1958. The chairman was Professor Fessard, Professor of Electroneurophysiology, Collége de France, and over fifty delegates attended representing France, Great Britain, Holland, Italy, Japan, Sweden, Switzerland, the U.S.A. and Western Germany. In his opening address, Dr. Zworykin spoke of the necessity for co-operation between medical men and engineers and the difficulties experienced due to the lack of a common language.

Since only Italy, Japan and the U.S.A. had existing medical electronics organizations and most of the delegates had no terms of reference allowing them to represent their countries, it was agreed that all the delegates would belong to an interim committee, and that an Executive Committee be appointed.

The following apointments to the Executive Committee were made:—

President:	Dr. V. K. Zworykin (U.S.A.)
Vice Presidents:	Dr. M. Marchal (France)
	Dr. C. N. Smyth (G.B.)
Secretaries:	Dr. C. Berkley (U.S.A.)
	Dr. A. Rèmond (France)
Treasurer:	Mr. B. Shackel (G.B.)

It was also agreed to appoint a number of advisers to the committee and the following immediate appointments were made:—

Prof. O. Wyss (Switzerland)
Dr. C. Sakamoto (Japan)
Dr. P. H. Bekkering (Holland)

Two additional members were co-opted to the Executive, Dr. S. Sherwood (G.B.) and Mr. W. J. Perkins (Associate Member) (G.B.).

The duties of the Executive would be to help the Interim Committee in its aim of co-operation with other organizations in the field and to organize committees on a regional basis, so that eventually they would form the basis of an International Committee. They would also be responsible for the organization of the second conference and the preparation of a bibliography. It was decided that the conference should be held in about a year's time in Paris and that it should be restricted to about 250 members.

It was suggested that the proposed bibliography should be based upon that already produced by the Electronics Centre of the Rockefeller Institute, and members of the Conference were requested to provide future references.

In addition to the formal business proceedings of the Conference a number of short papers were read. These dealt with such subjects as an ultraviolet colour translating microscope, the so-called "radio pill" used for the transmission of internal measurements of pressure, etc., study of eye movement, electroneurophysiology, radiography, medical data processing and the use of computers in medical research and practice.

The Executive Committee expressed their appreciation of the offer by Dr. R. C. G. Williams on behalf of the Institution of Electrical Engineers, and Mr. W. J. Perkins on behalf of the British Institution of Radio Engineers, placing the help and facilities of their respective organizations at the disposal of the next conference.

As members will be aware, the Institution has always taken a keen interest in the application of electronics to medicine. During the past ten years there have been nearly 30 papers read on this important subject both in London and at Local Section meetings, and several of these, as well as other papers, have been published in the *Journal*. These proposals for a definite framework for international co-operation will therefore meet with support from members generally and particularly those professionally concerned with this branch of electronic engineering.

To enable the Institution to co-operate more fully in the work of the Conference, members in this field are invited to notify the General Secretary of their interests; offers of papers are especially welcomed.

GRADUATESHIP EXAMINATION—MAY 1958—PASS LISTS

These lists contain results for *all* successful candidates in the May Examination. A total of 425 candidates entered for the examination.

LIST 1

The following candidates having completed the requirements of the Graduateship Examination, are eligible for transfer or election to Graduateship or higher grade of membership.

Candidates in Great Britain

BENTLEY, Edward Leslie. (S) *London.*
BONNER, John Stafford. (S) *London.*
COLLINS, Cyril Raymond. (S) *Glasgow.*
COOKE, Dick. *Scunthorpe.*
CORBEN, Clifford Bernard. (S) *London.*
CUTLER, George Donald. (S) *London.*
FRAZER, Charles Henry. (S) *London.*
GREEN, Kenneth Henry. (S) *London.*
HANSFORD, Douglas John. *London.*
HARIDAS, Krishnan. (S) *Birmingham.*
HOWARTH, Edwin. (S) *Plymouth.*
JASTRZEMBSKI, Jerzy A. (S) *London.*
KENTLEY, Eric William. (S) *London.*
LARGE, Douglas Blake. (S) *London.*
MEDROW, David G. (S) *H.M.S. Sheffield.*
O'CONNOR, Joseph Francis. *Manchester.*
PAIS, Aloysius Francis. (S) *Bristol.*
PANTHAKY, Jal-Khurshed. (S) *London.*
RICE, Matthew Joseph. (S) *Dublin.*
ROYLE, Basil Leonard. (S) *London.*
RYNKIEWICZ, Arthur E. (S) *London.*
SMITH, James Barnard. (S) *Manchester.*
THOMAS, Philip Robinson. (S) *Plymouth.*
WHITMORE, Dennis Ainsworth. *London.*

Overseas Candidates

ANAND PRAKASH. (S) *Banaras.*
BANSAL, Vijai Kant. (S) *Calcutta.*
BEN-DOR, Baruch. (S) *Tel-Aviv.*
BHATTACHARYA, Dilip Kumar. *Calcutta.*
BRUNNER, Amos. (S) *Tel-Aviv.*
CHEUNG SHIU HUNG. (S) *Hong Kong.*
de BRUYNE, Pieter. (S) *Delft.*
FONG YAN, Alick. (A) *Hong Kong.*
GREENWOOD, Frank. (S) *Adelaide.*
GUPTA, Tara Chand. (S) *Delhi.*
GURCHARAN, SINGH SURIE. (S) *Bangalore.*
JAIN, Naim Chand. (S) *Delhi.*
KARAMJIT SINGH. (S) *Bombay.*
KHADKIKAR, Ganesh D. *Jabalpur.*
KRISHNASWAMY RAO, M. S. *Bangalore.*
KUDDYADY, Chandrashekhar Raghunath. (S) *Bombay.*
KUNDU, Sati Nandan. (S) *Bangalore.*
LIPSCHITZ, Gerhard Gad. (A) *Tel-Aviv.*
MALHOTRA, Madan Mohan. (S) *Delhi.*
MARATHE, Yashvant. (S) *Agra.*
MARIOUW, Rudi Gijsbert F. (S) *Delft.*
MURTY, Suryanarayana Y. (S) *Banaras.*
NICE, Richard Keith. (S) *Essen.*
NIRODI, Suresh. (S) *Bombay.*
PATTABIRAMAN, A. K. (S) *Delhi.*
PULLAPERUMA, Don. G. (S) *Colombo.*
RAJOGOPAL, A. (S) *Bangalore.*
RENTON, Frank Morley. (A) *Montreal.*
SHAH, Rajendra Kantilal. (S) *Bombay.*
SIVASANKARA PILLAI, A. (S) *Trichinopoly.*
TARLOCHAN SINGH. (S) *Agra.*
VIRDI, Harbans Singh. (S) *Lucknow.*
VISWANATHAN, S. (S) *Madras.*
WRIGHT, Frank William. *Kirkuk.*
YOUAKIM, Mansour Francis. *Amman.*

LIST 2

The following candidates were successful in the parts indicated

Candidates in Great Britain

ANWAR, Salahuddin N. (4) (S) *London.*
BOWEN, Joseph Alfred Edward. (1, 2, 3) (S) *London.*
CACHIA, Saviour. (1, 3) (S) *London.*
CAWTHORN, Lewis. (3) (S) *London.*
CHESTER, Michael William. (1) (S) *Lowestoft.*
CHRISTIE, Stanley. (1, 3) (S) *Dublin.*
COLLINS, James Edward. (1) (S) *London.*
CORBETT, John Richard Galliers. (1, 3) (S) *London.*
CURLEY, Michael Joseph. (1, 2, 3) (S) *Dublin.*
DAVIS, Frank. (1) (S) *Birmingham.*
DUNELL, Wilfred Maurice. (4) *London.*
DUNNETT, Paul Wesley. (4) (S) *Bristol.*
FOSTER, David B. (5) (S) *Cambridge.*
FRASER, William Morrison. (1) (S) *London.*
GONZAGA, Victor E. (3) (S) *London.*
GREEN, John Owen. (5) (S) *Cambridge.*
GRIGGS, Brian E. (5) (S) *Cambridge.*
HORE, John Reginald. (4) (S) *London.*
JACKSON, Michael C. (4) (S) *Bristol.*
KEMP, Paul Courtney. (2, 3) (S) *Plymouth.*
KENWRIGHT, William Arthur. (1, 3) *Manchester.*
LEONARD, Allan G. (2) (S) *London.*
LETT, William Frederick. (3) *Birmingham.*
MASSINGHAM, Richard Peter. (1, 2) *Birmingham.*
MURPHY, Arthur S. (3) (S) *Manchester.*
NEED, Richard John. (4) (S) *London.*
NICHOLSON, Horace Leonard. (2) (S) *Glasgow.*
PARKER, William R. (2) (S) *London.*
PERYER, Michael G. (1) (S) *London.*
PYSDEN, Geoffrey T. (3) (S) *Birmingham.*
RIOUAL, Louis Auguste. (5) (S) *London.*
SNOW, James William. (2, 3) (S) *Birmingham.*
SOAMES, Michael R. (5) (S) *Cambridge.*
STEELE, Michael. (3) (S) *London.*
WALLACE, David John. (4) (S) *London.*

Overseas Candidates

AGGARWAL, Govind Saroop. (2) *Delhi.*
AGHARKAR, Tatnakar Vinayak. (4) (S) *Delhi.*
AHUJA, Om Prakash. (3) (S) *Lucknow.*
BALASUBRAMANYAM, T. R. (4) (S) *Madras.*
BANERJEE, Gour. (1) (S) *Madras.*
BHAT, Manohar Ganesh. (2) (S) *Bombay.*
BHOWAL, Kamal Kanti. (1) (S) *Bangalore.*
BHOWMIK, Sital Chandra. (1) (S) *Delhi.*
BROOKS, William G. E. (4) (S) *Toronto.*
BUZZELLI, Giovanni E. (2, 3) (S) *Turin.*
CAPNISTOS, Emman. (3) (S) *Athens.*
COUSINS, Ralph W. (1) (S) *Toronto.*
DEEKSHITULU, Rayaprolu. (3) (S) *Bangalore.*
FRISCH, Abraham. (1) (S) *Tel-Aviv.*
GOVINDARAJULU, R. G. (1, 2) (S) *Trichinopoly.*
GUPTA, Jitender Kumar. (2) (S) *Delhi.*
HANCOCK, Peter Burnell. (1) (S) *Newfoundland.*
HEMAN DAS BALWANI. (1, 2, 3) (S) *Delhi.*
HUTCHINSON, David Norman. (3) (S) *Montreal.*
JASWANT SINGH, Ghoman. (1) (S) *Delhi.*
JEYASINGH, Joshua Daniel Rajomoney. (4) (S) *Bombay.*
JOGBAND, Keshav Raghunath. (2, 3) (S) *Bombay.*
JOSHI, Rajaram Sivasanker. (3) (S) *Delhi.*
KHAMBADKONE, Murlidhar Ramrao. (3) (S) *Bombay.*
KOHLI, Suraj Parkash. (2) (S) *Kanpur.*
KOUTOURIOTIS, Georges. (3) (S) *Athens.*
KRIEGSMAN, Anthonius H. (1) (S) *Delft.*
KUMAR DEVA, Shanti. (2) (S) *Agra.*
MALHOTRA, R. N. (5) (S) *Delhi.*
MARTENS, Alexander E. (4) (S) *Toronto.*
MARTIN, Ronald Henry. (1) (S) *Malta.*
MASHIAH, Baroukh Elie. (3) (S) *Tel-Aviv*
MATHUR, Satya Prakash. (1) (S) *Lucknow.*
MEDIWAKE, Weerakoon Bandara. (1, 3) (S) *Colombo.*
MENON, Venugopalan V. (1) (S) *Bangalore.*
MITRA, Gobinda Lal. (3) (S) *Kanpur.*
NAG, Asim. (2, 3) (S) *Calcutta.*
NARASIMHA IYENGAR, R. Y. (2, 4) (S) *Bombay.*
NARASIMHAN, K. S. (2) (S) *Agra.*
NEUMANN, Shimon Siegfried. (1) (S) *Tel-Aviv.*
NEWMAN, Martin Myer. (3) (S) *Tel-Aviv.*
OKONKWO, Godwin Chukunwike. (1, 3) (S) *Lagos.*
PARKASH LAL CHAWLA. (2) (S) *Delhi.*
RAGOOWANSI, Nandlal Lakhoomal. (2, 5) (S) *Bombay.*
RAJIAH, Mallapuram. (2) (S) *Hyderabad.*
RAMADORAI, T. C. (2) (S) *Madras.*
RAMA RAYA KINI, P. (2, 5) (S) *Bangalore.*
ROESENER, Werner. (1, 3) (S) *Toronto.*
RUPRAI, Balwant Singh. (4) (S) *Hyderabad.*
SADASIVA DASS. (3) (S) *Delhi.*
SAUNDERSON, John Joseph. (2) (S) *Woomera.*
SCHUITEMAKER, Jozef Johannes. (1, 3) (S) *Delft.*
SHINDE, Manohar Bhimrao. (1) (S) *Bombay.*
SIKARWAR, Raghvendra Singh. (2) (S) *Jabulpur.*
SIMMS, Terence. (2, 3) (S) *Hong Kong.*
SRINIVASAN, Subramania. (2) (S) *Delhi.*
STEYN, Marthimus J. (4) (S) *Cape Town.*
SUJAN, Chander Sobhraj. (2) (S) *Bombay.*
SURYA PRASAD, K. M. (2) (S) *Madras.*
SUSHIL, Kumar. (3) (S) *Delhi.*
TALWAR, Satish Kumar. (3) (S) *Agra.*
TEMBE, Shivram Ramchandra. (4) (S) *Bombay.*
TOEG, Haim. (3) (S) *Tel-Aviv.*
UPADHYAY, Ram Adhar. (1, 2) (S) *Delhi.*
USMAN, Mirza Mohd. (2) (S) *Dhahran.*
WALKER, Rainer G. (1) (S) *Toronto.*
WATSON, William John. (1) (S) *Johannesburg.*
YOGINDER, Singh Jaswal. (2) (S) *Delhi.*

(S) denotes a Registered Student

. . . Radio Engineering Overseas

A selection of abstracts from European and Commonwealth journals received in the Library of the Institution. Members who wish to borrow any of these journals should apply to the Librarian, stating full bibliographical details, i.e. title, author, journal and date, of the paper required. All papers are in the language of the country of origin of the Journal unless otherwise stated. The Institution regrets that translations cannot be supplied.

538.66:621.314.2:621.318.42

Use of thermo-sensitive attenuation-coils and transformers. O. Hora. *Automatisace, (Prague)*, **3**, pp. 78-82, March 1958.

The article deals with the use of attenuation-coils and transformers, sensitive to temperature for the measurement of a momentary temperature as well as for checking the upper limit of temperature admissible. There is also a description of several complete electrical signalling equipments. The article ends with an evaluation of the properties of these new elements.

621.3.012.11:621.3.09

A new circle diagram for transformations in transmission line technique. G. W. Epprecht. *Archiv der Elektrischen Übertragung*, **12**, pp. 289-293, June 1958.

Transmission line techniques frequently present the problem of determining the influence of a quadripole on to the following network. This paper discusses the graphical solution of this problem, using the $\zeta - \theta$ diagram as a modified Carter diagram. From the simple phaseless and lossless transformation of the characteristic impedance to the mappings by a general reciprocal quadripole the new diagram leads with rapidity and clarity to a solution of the problems.

621.3.04:621.82.3

Analogous transistor system design and nodal methods of construction with applications to research equipment and prototype evaluation. R. F. Treharne. *Proceedings of the Institution of Radio Engineers, Australia*, **19**, pp. 319-347, July 1958.

Analogous design and nodal methods of construction allow valve engineers to take up the design of transistor circuits quickly and to employ methods of construction which are quick and economical. A block diagram which is physically realizable in valves is usually physically realizable using transistors. Nevertheless circuits which are quite unrelated to valve circuitry will eventually be used where appropriate. A two-transistor circuit unit may be wired directly from a circuit diagram to a chassis consisting of a tag strip. Large systems may be assembled from these simple, economical units. Modular design on them makes efficient use of inherent repetition. Every node in the system is systematically available for testing. Transistor action and circuits are described in terms of the thermionic valve analogy. Examples of the application of the technique of circuit development are given. They include d.c. amplifiers for telemetry, servo-mechanisms and analogue computing, timing generators, a beacon transmitter, a receiver and pulse units. Larger transistor systems are described.

621.314.23:621.941.2-52

The differential transformer and its use in the automatic control of machine tools. J. Zeleny. *Automatisace, (Prague)*, **3**, pp. 68-72, March 1958.

The article deals with elements forming essential parts of various position-servomechanisms, the importance of which, so far as automatization in machine manufacturing is concerned, has considerably increased during the last few years. There is also a description of the principle, function, property and construction of elements.

621.37/38.016.35

Operating stability of electronic equipment. M. Savescu. *Telecomunicatii (Bucharest)*, **2**, No. 2, pp. 75-80, March-April 1958.

An important problem in electronic equipment design is the stability of its functioning with variations of the circuit parameters from their nominal values. A general formulation of the problem and solving methods are given. Some examples are discussed.

621.372.543.3

The design of a triple tuned band-pass filter. R. J. L. Bosselaers and J. Roorda. *Tijdschrift van het Nederlands Radiogenootschap*, **23**, pp. 115-134, May 1958.

The design of a triple-tuned band-pass filter for an i.f. amplifier of a radio receiver, with the three circuits tuned to the i.f., is considered from the points of view of symmetry of the band-pass curve, the shape of that curve and to the amplification at the resonant frequency. From the possible shapes of the band-pass curve the one with one peak only is selected and discussed in detail. The design formulas are derived and worked out with a view on the practical realization of the filter. Finally examples are worked out for a filter with constant bandwidth, and for a filter with variable bandwidth but with restricted change of amplification when the bandwidth is varied respectively.

621.375.1

Band-pass amplifiers, their synthesis and gain-bandwidth factor. Fuad Surial Atiya. *Archiv der Elektrischen Übertragung*, **12**, pp. 251-264, 317-325, June and July 1958. (In English).

The following types of band-pass amplifiers are investigated: (*a*) the synchronous amplifier; (*b*) the maximally-flat, stagger tuned, single-tuned amplifier; (*c*) the quasi-Chebyshev, stagger-tuned, single-tuned amplifier; (*d*) the maximally-flat, stagger-damped, double-tuned amplifier with losses in only one of the two coupled circuits, and with equal losses in the two coupled circuits; (*e*) the quasi-Chebyshev, stagger-damped, double-tuned amplifier as for type (*d*); (*f*) the maximally-flat amplifier composed of stagger-feedback-pairs; (*g*) the quasi-Chebyshev amplifier composed of stagger-feedback-pairs. Versions, composed of identical sections of the above, are also considered. The synthesis and design-formulae of the above types are given, and expressions for the gain-bandwidth factor derived. The gain-bandwidth factor F is plotted against the number of stages m. It is seen that values of F as high as 4 are possible (type (*e*)). The selectivity of the quasi-Chebyshev type is superior to that of the maximally-flat, and far superior to that of the synchronous type. The relative merits of the types considered are discussed, and examples of the most important types are solved.

621.375.132.9:534.6

A new logarithmic amplifier. T. M. BAJENESCU. *Telecomunicatii (Bucharest),* **2**, No. 2, pp. 86-89, March-April 1958.

The block diagram of an installation for reverberation time measurements is given, together with the circuit diagram of a two-stage push-pull logarithmic amplifier. After a discussion on the physical processes involved in this amplifier, design formulae are given together with experimental verification.

621.375.4:621.396.621.54

A transistorized i.f. amplifier for communication receivers. A. S. PETT. *Proceedings of the Institution of Radio Engineers, Australia,* **19**, pp. 351-357, July 1958.

The design of an i.f. amplifier is outlined on a step-by-step basis and the necessary information is presented in the order in which it is required. An equivalent circuit is given and the range of values of its parameters demonstrated. It is shown how allowance is made for these varying parameters in the design formulae. An i.f. amplifier suitable for use in a communications receiver is specified and a design meeting the specification is briefly described.

621.382:621.37/9.002.2

Development of manufacturing techniques of semiconductor equipments. J. M. MERCIER. *L'Onde Electrique,* **38**, pp. 342-346, May 1958.

Pointers to the developments which will make future equipment of better performance and lower price are discussed. The techniques of purification of the materials, the means of manufacturing the active part of the equipment and the techniques of mass production are described.

621.382.3:621.317.75

An oscilloscope accessory for the display of transistor characteristic curves. R. E. AITCHISON. *Proceedings of the Institution of Radio Engineers, Australia,* **19**, pp. 370-373, July 1958.

A method is described for the display of transistor characteristic curves on an oscilloscope. A full wave rectified voltage is applied between the collector and emitter. A constant current is supplied to the base, the value alternating between two values in synchronism with the two phases of the collector supply, one for each half of the rectified sine wave. Using an oscilloscope as a display unit, two of the family of curves $(I_{ce}/V_{ce})\ I_b$ = constant, are produced. Such a display is particularly suitable for production testing of transistors during manufacture or for pre-selection before insertion in circuits. A simple modification enables the matching of a pair of transistors. All waveforms necessary for operation of the unit are obtained to the necessary accuracy with simple circuits using silicon rectifiers, Zener diodes, and two voltage regulator gas discharge tubes.

621.382.3:621.374.3

Trausistor monostable multivibrator for use with counting registers. R. E. AITCHISON. *Proceedings of the Institution of Radio Engineers, Australia,* **19**, pp. 315-318, July 1958.

The use of monostable multivibrators using transistors for the operation of counting relays is considered. A suitable design is given in which the pulse width can be readily controlled and is relatively independent of variations of transistor parameters. A specific example has a maximum counting ratio of 20 pulses per second, requiring a standby current of 6 mA and a maximum current of 18 mA from a 12 V supply, and is suitable for use with transistor decade scaling units, in equipment for the measurement of frequency or nuclear radiation.

621.382.333.32

A storing and switching transistor. W. v. MUNCH and H. SALOW. *Nachrichtentechnische Zeitschrift,* **11**, pp. 293-299, June 1958.

A storing switch can be produced by inserting a tungsten point into the collector contact of a *npn*-barrier transistor during the alloying process. The resulting input characteristic is similar to that of a thyratron. The switch transistor in a blocked condition has a differential input impedance of 1 megohm while the impedance in a conducting state is less than 10 ohms. An essential characteristic of a switch transistor is its high switching velocity in comparison with other semi-conductor devices.

621.385.029.6

On the theory of the electron wave tube with elliptic cross-section. P. MATTILA. *Acta Polytechnica Scandinavica, Electrical Engineering Series I, Helsinki,* 1958, 78 pp. (In English).

The theory of the electron wave tube of elliptic cross section is derived, based on small signal theory. In the case of the elliptical tube the E and H waves contain Mathieu's functions. The development of electron beam waves on two completely mixed beams is studied in detail. Comparison of the elliptic and circular cross-section electron wave tubes is made. One practical example is worked out in detail. At small values of eccentricity it is possible to obtain nearly the same amplification with elliptic cross section electron wave tubes as with circular cross-section tubes. With eccentricity approaching unity the amplification falls abruptly approaching zero as its limit.

621.391

The reliability of binary code transmission by means of various types of modulation. H. J. HELD. *Nachrichtentechnische Zeitschrift,* **11**, pp. 286-292, June 1958.

Mathematical formulae for the probability of a faulty binary step transmission are compiled and compared with one another. A simple and general approximate formula for the probability of errors which is valid for all these methods is given. This formula contains the signal/noise ratio as well as three constants depending on the method of keying. A statistical differentiation theory is used to show that the method of phase reversal keying is an optimum binary transmission method. Laboratory arrangements for an experimental determination of the error probability in various binary transmission methods are described briefly and the results of measurements are reported.