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An Important Centenary

THIS is the time of the year when Exhibitions predominate, and after the almost overwhelming number of Shows and Conventions which have claimed his attention during the recent months, the professional engineer may now be looking forward to an uninterrupted spell of his normal activities, but before returning to his desk or laboratory there is one more Exhibition which he should not overlook.

It is the special Exhibition at the Science Museum organized by the Museum Authorities with the co-operation of the Institution of Electrical Engineers, the General Post Office and a number of leading manufacturers to mark the centenary of the laying of the first submarine telegraph cable across the Straits of Dover on August 28, 1850, by the Brett brothers. This Exhibition, which remains open until October 28, is an important one, for the event it celebrates may be said to be the birth of international communications, of which the B.B.C.'s recent Dover-Calais Television experiment as a centenary commemoration is a normal development.

Accustomed as we are today to the wonders of modern radio and to the rapidity with which we can communicate with almost any part of the world, it is difficult for us to realize the seclusion in which our ancestors lived a hundred or more years ago before the invention of the railways and the telegraph. The electric telegraph itself arose from the needs of the railways for a means of communication from station to station. The first telegraph system was that from Paddington to West Drayton in 1838 and in the ten years that followed the main network of railways and telegraph systems were being pushed towards completion. In 1846 the London to Dover telegraph had been completed and it was not unnatural that the engineers of the day should look across the channel as a logical extension to the inland telegraph system.

The wonder and amazement with which the laying of the Dover-Calais cable was greeted is expressed in a leader of *The Times* of August 31, 1950, which concludes with the following:

"In the case of the submarine electric telegraph, the first and obvious effect of this instantaneous communication between the two most civilized and powerful nations of the world will be to unite them so closely in community of interests as to secure their co-operation in all designs that may promote the advancement of humanity and maintain the peace of the world."

Such an expression of high hope will be read today with some heartsearching and may create in many of us a desire to return to the good old days.

The Exhibition itself is an exhibition of the submarine cable—a British invention upon which has been founded a cable making and cable laying industry which has remained predominately British.

In a specially prepared handbook (H.M.S.O., price 2s. 6d.) entitled "100 Years of Submarine Cable," G. R. M. Garrett, of the Science Museum, traces the fascinating history from the Dover-Calais cable in 1850 and the subsequent transatlantic cable to the present day, when upwards of half a million miles of submarine cables span the world.

The Video Output Stage

By EDWARD T. EMMS, B.Sc., A.R.C.S. and EMLYN JONES, B.Sc., A.M.I.E.E.

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Part I

THE television video output valve acts as a wide band amplifier working into a capacitance which is formed by the C.R.T. input capacitance, the output capacitance of the video valve, and stray capacitance due to wiring, etc. In addition, the synchronizing separator and noise limiter are often fed from the video valve anode, and the effective input capacitance of these stages must then be added.

The video valve is usually driven from the demodulator valve, although it may form its own demodulator as in the case of the anode bend demodulator video stage.

The two most common types of video valve circuit employed are the anode compensated stage and the cathode compensated stage.

The requirement of the video stage is that it shall reproduce, with minimum distortion, the input waveform (or the envelope of the input waveform in the case of the anode bend demodulator video stage). This implies that the circuit shall have an indicial response (i.e., a response to unit step) of high speed without excessive overshoot.

The various types of video valve stage in common use will be examined with a view to discussing their relative advantages and disadvantages. To this end, all video stages will be designed to have a critically damped indicial response. This means that the response to unit step will have the shortest time of rise consistent with no overshoot. In practice the video stage may be allowed to exhibit some overshoot in its indicial response if the indicial response of the R.F. stages exhibits no overshoot or very small overshoot.

We shall discuss mainly the response of the video stage to rapid changes of the input waveform; this is chiefly determined by the H.F. response of the stage. The average level of this waveform, however, contains information about the average brightness of the picture and this information must be transmitted to the C.R. tube. The D.C. amplification of the stage should therefore be equal to its amplification in the middle of its frequency band. With the particular waveforms used in television this can be achieved, if desired, by the use of D.C. restoration.

Circuit Capacitances

Likely values of circuit capacitances are given below:—

	Minimum	Maximum
Video Valve Anode	2.0 pF	12 pF
Valve Holder and Wiring ...	5.0 pF	10 pF
C.R.T. Input Capacitance ...	5.0 pF	9 pF
Synchronizing Circuits ...	3.0 pF	10 pF
D.C. Restoring Circuits and Noise Limiters	0.0 pF	9 pF
	15.0 pF	50 pF

Thus we may say

15 pF represents an absolute minimum
30 pF represents a reasonable average
50 pF represents a probable maximum
} for the total capacitance existing in the output circuit of the video valve.

Build-up Time of the Video Stage

The build-up time will be taken to signify the time for the response to rise from 10 per cent to 90 per cent of the final value, when unit step is inserted.

The required build-up time of the video stage depends on the television system under consideration, and the number and quality of coupling circuits in the receiver.

In a typical receiver, coupling networks as below may be employed.

No.	Type
1.	Aerial to H.F. grid.
2.	H.F. Anode to Frequency Changer grid.
3.	Frequency Changer anode to 1st I.F. grid.
4.	1st I.F. anode to 2nd I.F. grid.
5.	2nd I.F. anode to demodulator anode.
6.	Demodulator cathode to Video Valve grid.
7.	Video Valve anode to C.R.T.

Nos. 1-5 are band-pass couplings.
Nos. 6-7 are low-pass couplings.

It may be shown that the speed of rise is reduced by approximately $1/\sqrt{2}$ every time the number of band-pass couplings in the system is doubled.[1] For this purpose a compensated video frequency coupling ranks as a band-pass coupling, while if staggered single tuned couplings are employed in the R.F. stages, two of these rank as one band-pass coupling.

Thus we see that, depending on whether single tuned circuits or transformers are used in the R.F. stages, there are between $4\frac{1}{2}$ and 7 equivalent band-pass circuits in the receiver. Each of these must be designed for a build-up time t', where

$$t' = T/\sqrt{n}$$

n = number of equivalent band-pass couplings
T = required overall build-up time.

For the above values of n ($4\frac{1}{2}$ and 7) we find $t' = 0.47T$ and $t' = 0.378T$ respectively, the average being $t' = 0.425T$. Actually, some overshoot is permissible, so that as a representative figure we shall take

$$t' = 0.5T \qquad (1)$$

This still leaves a value for T to be found. In order to estimate T we shall have to investigate the speed of the transmitted waveform, and also the desired definition. In order to ascertain the speed, the complete amplitude and phase responses would be required. However, an estimate of the speed may be obtained using the very approximate relationship

$$B\tau = 0.35 \qquad (2)$$

where B = 3 db semi-bandwidth (as defined by Fig. 1)
τ = build-up time.*

Thus for the B.B.C. transmission we have:

* These figures refer to the London transmission; the band width of the Midland's transmitter is somewhat wider.

Bandwidth at 3 db down = 2.75 Mc/s.
Build-up time (calculated) = 0.128 μS.
Duration of picture element = 0.173 μS.

It will be seen that the B.B.C. transmission has something in hand—the transmitted waveform has a build-up time smaller than the duration of a picture element.

If the build-up time of the receiver is made equal to the build-up time of the transmitted wave, the overall build-up time will be approximately $\sqrt{2} \times 0.128\ \mu S = 0.180\ \mu S$, which is only 4 per cent in excess of a picture element duration. Such a receiver will reproduce the 8 Mc/s. bars on the B.B.C. test pattern "C", at about 60 per cent of full contrast.

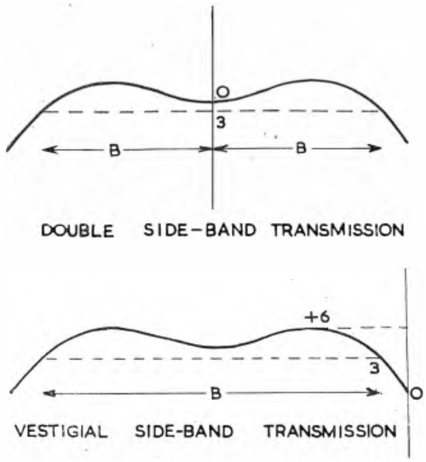


Fig. 1. Definition of B

We may arbitrarily specify the following criteria: The definition is

very good when the receiver build-up time is *one-half* the transmitted wave build-up time.
good when the receiver build-up time is *equal* to the transmitted wave build-up time.
average when the receiver build-up time is *twice* the transmitted wave build-up time.

Table 1 may now be calculated:

TABLE I

Definition	T(μS)	t′ (μS)
Very good	0.07	0.035
Good	0.13	0.065
Average	0.26	0.130

T build-up time of receiver.
t' = build-up time of any one coupling network (assumed equal).

The Required Output Voltage Swing

The voltage output swing required is determined by the picture tube employed, and more particularly by the maximum permissible beam current which it may pass without serious defocusing. The maximum beam current for direct-view picture tubes is usually 100 μA, but peaks up to 200 μA may occur. On the other hand, acceptable pictures have been obtained in a darkened room with only 5 μA mean beam current.

All Mullard direct-view picture tubes require an input of approximately:

16 volts p-p for 100 μA beam current.
21 volts p-p for 200 μA beam current.

Taking 24 volts as a representative figure to allow for tolerances and adding the correct synchronizing percentage of 30 per cent we require a *p-p* output voltage of 30 volts. This figure may be taken as a design figure for most direct view video stages.

For the Mullard projection tube MW 6-2 the corresponding design figure may be taken as 84 volts *p-p*.

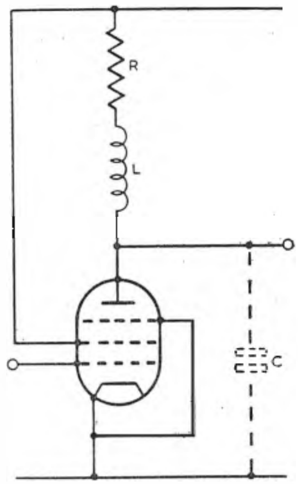


Fig. 2a.
First order shunt anode compensation

Polarity of Output Voltages

The video valve may be biased near cut-off and be driven positive by a white signal, or it may be biased just clear of grid current and be driven negative by a white signal.

In the first case the output signal is applied to the cathode of the C.R.T., while in the second case the signal is applied to the C.R.T. grid. Both methods are in common use.

Whereas the average power dissipation in the valve is much lower in the first case, the peak power can be greater; for example, a receiver may be left with the gain control at a maximum, so that the video valve completely overloads on the arrival of a signal. The video valve is then driven well into grid current during the picture periods.

In the second case, the maximum power dissipation occurs in the absence of any signal, and falls to a low value on overload.

The Anode Compensated Video Output Stage

The basic circuit for first order shunt compensation is shown in Fig. 2a.

This type of compensation employs only one compensating element, the inductance L in series with the load resistor R. Higher orders of compensation may be employed using more elements. Table II gives some of these circuits with design details for critical damping so that the indicial response exhibits no overshoot. Fig. 2b shows the indicial response obtained. The build-up times for these circuits are as follows:

Circuit	Build-up Time
1	2.19RC
2	1.57RC
3	1.39RC
4	1.32RC
5	1.27RC
6	1.40RC
7	1.24RC

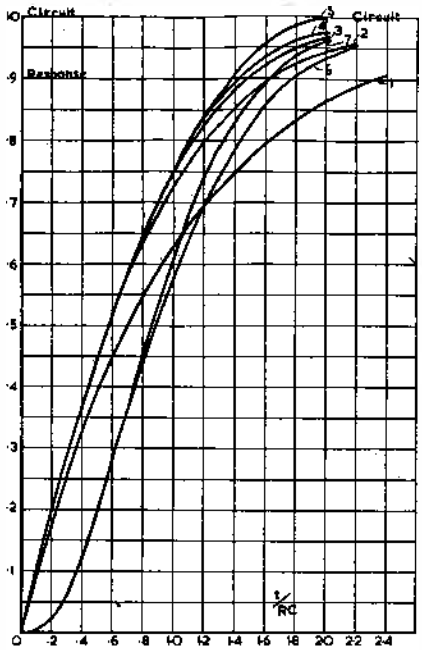


Fig. 2b.
The indicial response of anode compensated stages. (numbers refer to the circuits of Table II)

It will be noticed that the first order shunt compensated stage gives a build up time 1.4 times faster than that obtained for the uncompensated stage at the same gain (i.e., the same value of load resistor). Alternatively, the same build-up time may be obtained with 1.4 times the gain of the uncompensated stage. This, of course, is a well worth while improvement

Second order shunt compensation again shows a marked improvement over first order shunt compensation and in this case it is possible to provide C_1 (see Table II) by appropriate winding of L_1 so that it has the correct self-capacitance. This has been effected, in one design, by winding L_1 on a metal former.

High order compensation does not yield so marked an improvement, and it is doubtful whether the increased complexity and expense merits using a higher order compensation than second-order shunt compensation.

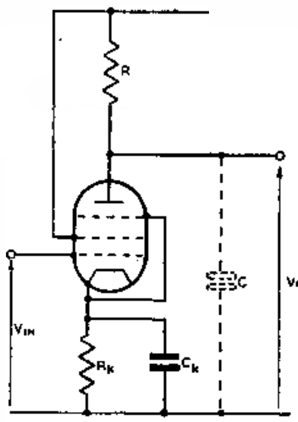


Fig. 3a.
Cathode compensation

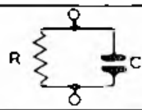
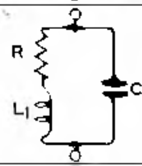
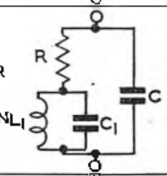
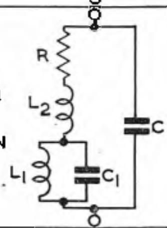
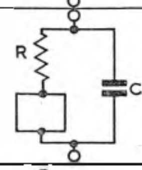
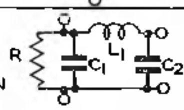
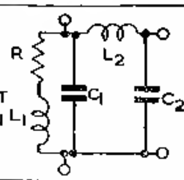
Further considerations may influence the designer's choice of video compensating circuit. For example, in a high-sensitivity receiver it may be deemed necessary to employ series compensation, so that the series choke may also serve to keep R.F. from the lead to the cathode-ray tube which might otherwise give rise to instability by being fed back to the input to the receiver. The presence of R.F. at the anode of the video valve, due to insufficient attenuation from the demodulator, should always be borne in mind when a complete design is essayed.

The Cathode Compensated Video Stage

Since the cathode compensated video stage is not so well known as the anode compensated video stage the principle of its operation will be described in detail. The basic circuit is shown in Fig. 3a.

If the valve delivered a unit step of current to the load, the output voltage would rise according to an

TABLE II

NAME	CIRCUIT	DESIGN DETAILS	RESPONSE TO UNIT STEP $x=t/RC$
1 UNCOMPENSATED			$1-\varepsilon^{-x}$
2 FIRST-ORDER SHUNT COMPENSATION		$L_1=\frac{1}{4}CR^2$	$1-(1+x)\varepsilon^{-2x}$
3 SECOND-ORDER SHUNT COMPENSATION		$L_1=\frac{8}{27}CR^2$ $C_1=\frac{1}{8}C$	$1-(1+2x+2x^2)\varepsilon^{-3x}$
4 THIRD-ORDER SHUNT COMPENSATION		$L_1=\frac{1}{4}CR^2$ $C_1=\frac{1}{4}C$ $L_2=1/16CR^2$	$1-(1+3x+4x^2+\frac{16}{3}x^3)\varepsilon^{-4x}$
5 INFINITE-ORDER SHUNT COMPENSATION			$0<x<2$ $x-\frac{x^2}{4}$
6 FIRST-ORDER SERIES COMPENSATION		$C_1+C_2=C$ $C_1=1/9C$ $C_2=8/9C$ $L_1=3/8CR^2$	$1-(1+3x+\frac{9}{2}x^2)\varepsilon^{-3x}$
7 MIXED FIRST ORDER SERIES & SHUNT COMPENSATION		$C_1+C_2=C$ $C_1=\frac{1}{5}C$ $C_2=4/5C$ $L_1=1/16CR^2$ $L_2=25/64CR^2$	$1-(1+4x+8x^2+8x^3)\varepsilon^{-4x}$

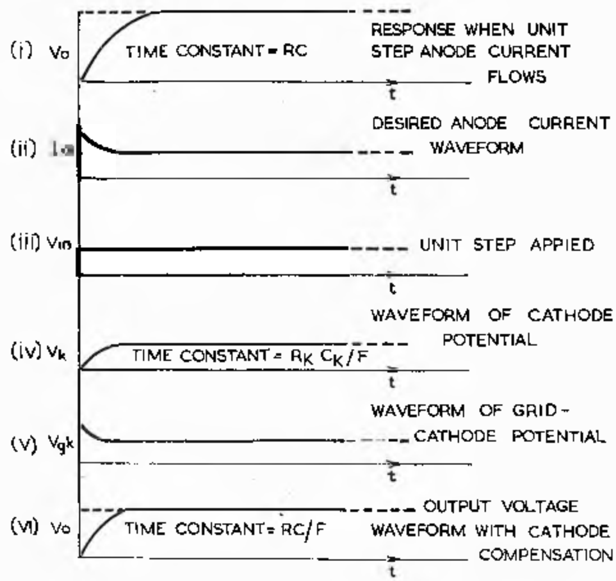


Fig. 3b.
Waveforms of cathode compensated stage

exponential curve with time constant RC, Fig. 3b(i). If, instead of a unit step of current, a valve current such as shown in Fig. 3b(ii) were delivered to the load, it is clear that the build-up time of the output voltage waveform would be reduced, and by appropriate choice of this anode current waveform no overshoot need be introduced. This particular anode current waveform is produced by an appropriate choice of the cathode time constant R_kC_k. If the input voltage waveform to the stage is a unit step, the cathode potential also rises according to an exponential curve with a time constant R_kC_k/F, where F is the feedback factor ($F = 1 + g_kR_k$).

This cathode waveform is shown in Fig. 3b(iv). The actual grid-cathode voltage applied to the valve is the difference between this and unit step and is shown in Fig. 3b(v). Since the anode current of a pentode is proportional to the applied grid-cathode voltage the anode current waveform is of the correct shape required as indicated in Fig. 3b(ii). An

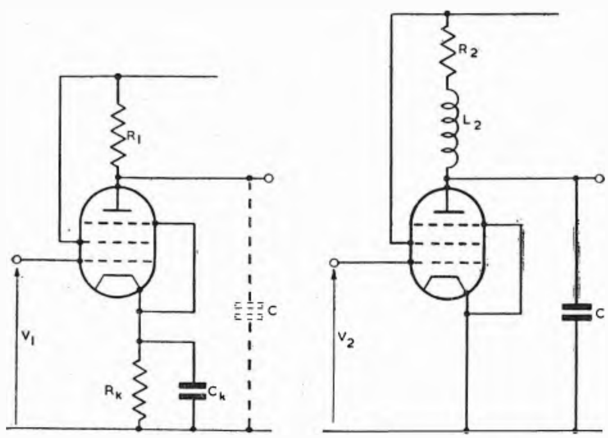


Fig. 4a.
Comparison of alternative circuits

analysis of the circuit indicates that, for an indicial response which has no overshoot with the smallest build-up time, we require

$$RC = R_kC_k \quad \ldots\ldots (3)$$

and the indicial response becomes an exponential with a time constant RC/F (Fig. 3b(vi)). The build-up time is therefore $2.19RC/F$.

A Comparison of the Anode Compensated and Cathode Compensated Video Output Stages

Since first order shunt compensation and cathode compensation are the most widely used video output circuits, it it worth while making a comparison of the two methods.

To make this comparison, we shall assume that the two circuits work into equal capacitative loads (C) and design them to have:

(i) the same build-up time.
(ii) the same output voltage.

Requirement (i) yields

$$\frac{2.19R_1C}{F} = 1.57R_2C \quad \ldots\ldots (4)$$

Where R_1 denotes the anode load in the cathode compensated video stage and R_2 denotes the anode load in the anode compensated video stage (Fig. 4a).

Requirement (ii) yields:

$$\frac{g_m R_1 V_1}{F} = g_m R_2 V_2 \quad \ldots\ldots (5)$$

Where V_1 and V_2 are the required input voltages to the cathode compensated and anode compensated video stages respectively.

Division of (5) by (4) gives

$$V_1 = 1.4\, V_2 \quad \ldots\ldots (6)$$

Thus we conclude that no matter what degree of cathode compensation is employed, the input voltage to the cathode compensated stage must be 1.4 times that applied to the anode compensated stage if conditions (i) and (ii) are to be satisfied. In other words, the gain of the cathode compensated stage is 3 db less than that obtained with the anode compensated stage.

If unit step were inserted into the circuits to obtain the same voltage waveform out (shape and magnitude) this would imply that the grid-cathode voltage of the cathode compensated stage would swing 1.4 volts for every volt that the grid-cathode space of the anode compensated stage would swing. This is a fact that is often levelled at cathode compensated stages as a disadvantage, the argument being that a cathode compensated stage overloads at the grid before an anode compensated stage, using the same valve and with the same build-up time and output voltage. However, this is not so, due to the finite build-up time of the input wave to the video stage in a television receiver.

In practice, the input wave is not unit step but has a finite build-up time, this being usually longer than the build-up time of the indicial response of the video stage. As a reasonable input voltage wave we shall adopt that shown in Fig. 4b.

This input wave is specified by:

$$\begin{aligned} f(t) &= 0 \qquad -\infty < t < 0 \\ &= t/\sigma \qquad 0 < t < \sigma \\ &= 1 \qquad \sigma < t < \infty \quad \ldots\ldots (7) \end{aligned}$$

It may then be shown that the maximum grid-cathode voltage occurring in the cathode compensated stage is given by:

$$\hat{V}_{gk} = 1/F\left\{1 + \frac{R_1C}{\sigma}\cdot\frac{F-1}{F}\left(1-e^{-F\sigma/R_1C}\right)\right\} \quad .... \ (8)$$

The build-up time of the input-voltage wave and the indicial response of the video stage are 0.8σ and $2.19R_1C/F$ respectively, so we have

$$p = \frac{\text{build-up time of input wave}}{\text{build-up time of indicial response of video stage}}$$

$$= \frac{0.8F}{2.19}\cdot\frac{\sigma}{R_1C} \quad \text{........................} \ (9)$$

and from (8) and (9) it follows :—

$$\hat{V}_{gk} = 1/F\left\{1 + (F-1)\cdot\left(\frac{1-e^{-2.75p}}{2.75p}\right)\right\} \quad ...... \ (10)$$

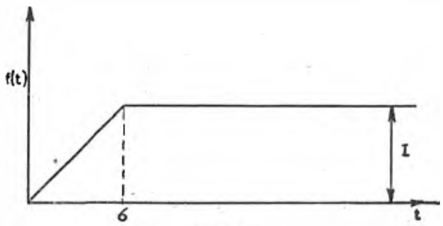


Fig. 4b.

Assured input voltage waveform

Fig 4(c) shows the grid-cathode waveform for $F = 2$, $p = 1$, while Fig. 4(d) shows the peak grid-cathode voltage obtaining f r various values of F and p (when 1.4 volts peak are inserted).

It may be shown that if the build-up time of the input wave is greater than the build-up time of the indicial response of the video stage (which is usually true) the peak grid-cathode potential swing required for the cathode compensated stage is less than that required for the anode compensated stage for the same output voltage (build-up time and magnitude), provided F is greater than 1.7.

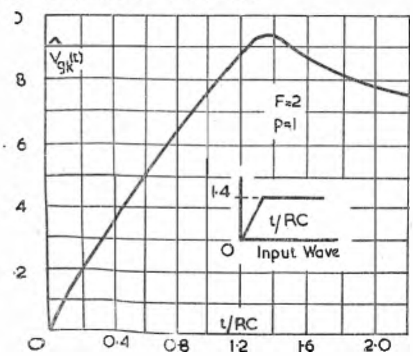


Fig. 4c.

The grid-cathode voltage waveform for F=2 p=1 for the cathode compensated stage

The Required Characteristics of the Video Valve

Since we have calculated the required build-up time of the video stage, and we know the required output voltage swing, we are in a position to calculate (a) the required load resistance, R, (b) the required $p-p$ current swing in the valve, for those circuits most used in practice.

(1) The uncompensated anode load (Circuit 1, Table II).
The build-up time is given by $t' = 2.19CR$.

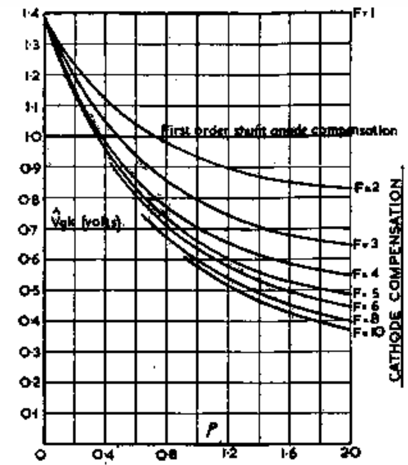


Fig. 4d.

The peak grid-cathode voltage for the cathode compensated stage

Loading Capacitance (pF)	Anode Resistance (R) required (ohms)			p-p Current Swing required (mA) for 30 v.p.p.		
	Build-up Time			Build-up Time		
	average $t'=0.13\,\mu s$	good $=0.065\,\mu s$	v. good $=0.035\,\mu s$	average $=0.13\,\mu s$	good $=0.065\,\mu s$	v. good $=0.035\,\mu s$
15	3,960	1,980	1,060	7.6	15.1	28.2
30	1,980	990	532	15.1	30.2	56.3
50	1,190	595	320	25.2	50.3	93.7

(2) The first-order shunt compensated anode load (Circuit 2, Table II).
The build-up time is given by $t' = 1.57CR$.

Loading Capacitance (pF)	Anode Resistance (R) required (ohms)			p-p Current Swing (mA) required for 30 v.p.p.		
	Build-up Time			Build-up Time		
	average $t'=0.13\,\mu s$	good $=0.065\,\mu s$	v. good $=0.035\,\mu s$	average $=0.13\,\mu s$	good $=0.065\,\mu s$	v. good $=0.035\,\mu s$
15	5,520	2,760	1,490	5.5	10.9	20.1
30	2,760	1,380	744	10.9	21.7	40.3
50	1,660	830	447	18.0	36.1	67.0

(3) The cathode-compensated stage $CR = C_kR_k$ (Fig. 3(a)).
The build-up time is given by $t' = \dfrac{2.19RC}{F}$

Loading Capacitance (pF)	(F)	Anode Resistance (R) required (ohms)			p-p Current Swing (mA) required for 30 v.p.p.		
		Build-up Time			Build-up Time		
		average $t'=0.13\,\mu s$	good $=0.065\,\mu s$	v. good $=0.035\,\mu s$	average $=0.13\,\mu s$	good $=0.065\,\mu s$	v. good $=0.035\,\mu s$
15	1.5	5,920	2,960	1,600	5.1	10.2	18.8
	2.0	7,930	3,970	2,130	3.8	7.6	14.1
	3.0	11,900	5,950	3,190	2.52	5.0	9.4
	4.0	15,850	7,930	4,260	1.89	3.8	7.1
	5.0	19,800	9,900	5,320	1.52	3.0	5.7
30	1.5	2,960	1,480	800	10.2	20.4	37.6
	2.0	3,970	1,980	1,070	7.6	15.2	28.2
	3.0	5,950	2,970	1,600	5.1	10.2	18.8
	4.0	7,930	3,960	2,130	3.8	7.6	14.2
	5.0	9,900	4,950	2,660	3.0	6.0	11.4
50	1.5	1,780	890	480	16.9	33.8	62.7
	2.0	2,380	1,190	640	12.6	25.2	47.0
	3.0	3,570	1,780	960	8.4	16.8	31.3
	4.0	4,770	2,380	1,280	6.2	12.4	23.5
	5.0	5,940	2,970	1,600	5.1	10.2	18.8

From the above tables it will be seen that a valve giving 22 mA $p-p$ swing will satisfy the first order shunt-compensated stage and the cathode-compensated stage for:—
(a) minimum capacitance and minimum time of rise,
(b) average capacitance and average time of rise,
(c) maximum capacitance and maximum time of rise.

If we add approximately 30 per cent to the above requirement to allow for the unusable bottom bend of the $I_a - V_g$ characteristic, the current at $V_g = -1$ for a valve of practically universal application, in direct-view receivers would be about 30 mA.

There are two possible extreme applications:

(a) working from 170 V H.T. line,
(b) working from 330 V H.T. line.

For the first case we require $I_a = 30$ mA at $V_g = -1$. This implies that the value of I_a at $V_g = -1$ when operating from a 330 V line will be about 85 mA, and the anode dissipation would then have to be about 28 watts. It will be understood that it is not practicable to employ a decoupled voltage dropping resistor to avoid this, because of the need to avoid overamplification of the D.C. component. Clearly two valves are really required to cover all cases.

For 170 V H.T. line application ($V_{g2} = 170$ V)

I_a at $V_g = -1 = 30$ mA
$W_a = 5$ watts
$W_{g2} = 1.25$ watts (assuming $I_{g2}/I_a = 0.25$)

For 330 V H.T. line application ($V_{g2} = 330$ V)

I_a at $V_g = -1 = 30$ mA
$W_a = 10$ watts
$W_{g2} = 2.5$ watts (assuming $I_{g2}/I_a = 0.25$)

A valve of universal application, although an excellent idea, is not necessary in many cases. The use of a smaller current valve, employing cathode compensation, may result in a definite economy in H.T. current required, and also yield a more economic design. This view has been adopted in the design of a video stage for the Mullard Projection system.

(*To be continued*)

References

[1] Kallman, Spencer & Singer. *Proc. I.R.E.*, 33, 1945, 169

PHOTOGRAPHING SOUND WAVES

ATTEMPTS by W. E. Kock and F. K. Harvey to "photograph" sound waves at Bell Telephone Laboratories are described in the July, 1950, issue of *Bell Laboratories Record*.

The tip of a long horizontal arm was made to describe an arc of a circle in the vertical plane in front of a loudspeaker emitting a reasonably constant and pure note. A small probe microphone mounted on the tip and facing the loudspeaker picked up the note, which was then fed to an amplifier. The output of this amplifier was used to light a pea-lamp mounted alongside the microphone. (See Fig. 1).

In order to scan the vertical sound field in front of the speaker the horizontal arm, driven from a small electric motor and gear train, was set in motion, and at the same time the whole apparatus was moved slowly away from the sound source.

Scanning took place in a darkened room, and a time exposure was made during the whole process by a camera facing at right angles to the direction of motion of the trolley.

Interesting results were obtained with an ordinary telephone handset (Fig. 2) and with the loud speaker in combination with acoustic lenses and prisms. Radiation from a variety of sound sources was tested, as well as diffractions around interfering objects and diffusion by divergent lenses.

A film strip was made by repeating a set of four frames, each one a quarter-wave different in phase position—the result being a presentation of waves moving outwards from a lens, converging to a focus, and then diverging again.

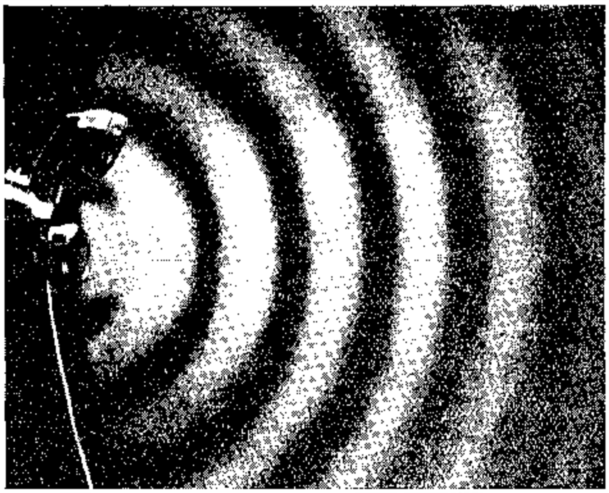
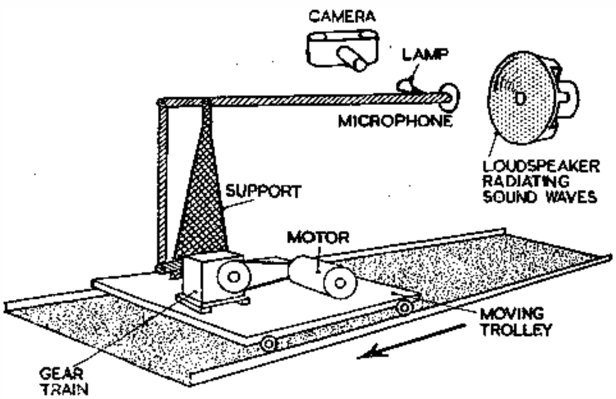


Fig. 2.
Results obtained with a telephone type handset. After publication of this picture in the July issue of the "Record" several observant readers protested that the sound "waves" appeared to be originating in the carbon transmitter. An explanatory note in the August issue remarks that during the experiments it was argued that a reversal of the transmitting and receiving positions should not affect the photographed pattern.
The scanning receiver at the end of the rocking arm was therefore made a sound source, and the handset was arranged with the carbon transmitter taking the place of the receiver. As expected, the resultant "sound" picture was the same. It is this which is reproduced above

Fig. 1.
A sketch of the apparatus used in the experiment. The gear train imparts a rocking movement to the horizontal arm, moving the lamp vertically



The Performance of Screening Rooms

By J. MIEDZINSKI, B.Sc.
and S. F. PEARCE, B.Sc., A.Inst.P.

INDUSTRIAL and medical radio frequency equipment have, for many years been recognized as serious sources of interference with radio reception. A survey by A. Turney[1] has shown that fields from the former equipment may be of the order of 100 μV/metre at distances up to about half a mile from the site of equipment having an unscreened work circuit. The field from a totally screened equipment was considerably lower, and the following describes a continuation of this work to determine the efficacy of simple forms of screening in reducing the field. Tests have been carried out on four examples and from the results obtained an attempt to deduce formulæ explaining the behaviour of these rooms has been made. Features affecting the performance of the rooms are also discussed.

The formulæ derived should prove useful in predicting the performance of a given screen or, in a less direct way, should enable the most economical design of screen to be obtained for a desired reduction in the field.

The efficacy of a screen depends in general on the type of radiator (or receiver) enclosed and on the size of the room, as well as on the quality of the screen itself. For the frequencies used in these tests the dimensions of the screening rooms were small compared with the wavelength. Schelkunoff[2] has shown that for this case the attenuation will be less for a magnetic dipole than for an electric dipole, and the former type of radiator was, therefore, used in the tests in order to deal with the less favourable case.

Screening Rooms and Test Equipment

The four rooms examined were of similar construction and differed only in the screening material used. A prefabricated construction was employed, each room consisting of six sections forming the walls, roof and floor which could be bolted together by 5/16 in. Whitworth bolts. Each section comprised a wooden frame on which the screening material was mounted, and contact between the various sections of the screening was obtained by wrapping the material round the edge of the framework so that adjacent sections were held in contact by the bolts. The room dimensions were approximately 6 ft. 6 in. by 6 ft. 6 in. by 7 ft. The door, constructed similarly to the rest of the room, was reinforced round the edges with thin steel channel, and opened and closed by sliding in a steel channel along top and bottom of the aperture. A similar channel was fitted along the door jambs. Contact between the door and the room was maintained by spring loaded balls spaced every 16 in. down the vertical edges of the door. The materials forming the screen for the four rooms were as follows:—

Room No. 1 : Perforated Zinc Sheet :

Sheet thickness:	0.015 in.
Hole diameter:	0.080 in.
Spacing between holes:	0.145 in.
Angle between rows of holes:	60°
Fractional area taken up by metal:	0.56

Room No. 2 : Electro-galvanized Expanded Steel :

Sheet thickness:	0.022 in.
Gap between punchings:	0.031 in.
Mean thickness of galvanizing:	0.00014 in.
Fractional area taken up by metal:	0.27

Room No. 3 : Tinned Iron Mesh :

Wire diameter:	0.020 in. (25 S.W.G.)
Mesh size:	0.25 in. square woven ("4 mesh")

Tinned before weaving, no bonding between wires.

Mean thickness of tinning:	0.00026 in.
Fractional area taken up by metal:	0.16

Room No. 4 : Galvanized Iron Wire Netting :

Wire diameter:	0.027 in. (22 S.W.G.) after galvanizing
Mesh size: hexagonal:	0.60 in. between centres

Galvanized after weaving.

Mean thickness of galvanizing:	0.0016 in.
Fractional area taken up by metal:	0.09

Transmitters of conventional design were used. The radiating element was a tuned unbalanced screened loop. The loop current was measured by a thermocouple ammeter connected in series with the loop at the earthy end of the circuit. Four sizes of loop were used to cover the frequency range of 0.75 Mc/s. to 24 Mc/s. The power supply for the transmitter was obtained from batteries and a rotary transformer. In this way the screening properties of the room could be examined without reference to the efficacy of any filtering in leads entering the room.

The receiver used for measuring the field was an Army receiver R.206 modified for the measurement of field strength.

Measurement of Attenuation

The transmitter, and screening room when present, were mounted on a platform which could be rotated about a fixed axis. The radiating loop was mounted vertically and arranged so that the diameter was on the axis of rotation and approximately at the centre of the room. The receiver was set up at a fixed distance from the transmitter site and the polar diagram obtained by rotating the transmitter platform and noting the readings for various angular positions of the platform.

The tests were carried out at six frequencies harmonically related in the range of 0.75 Mc/s. to 24 Mc/s. Polar diagrams for the unscreened transmitter were obtained at these frequencies, and the procedure repeated for the transmitter enclosed by the screening rooms in turn.

A good earthing system was provided at the transmitter site and the effect of earthing of the transmitter and screening room examined.

The results of the field measurements were plotted in the form of polar diagrams. At each frequency the readings were adjusted to refer always to a given loop current so that all diagrams could be plotted on a single graph, and the attenuation for the various screening rooms read directly from the graph. It was immediately apparent, however, that the attenuation so obtained would depend to some extent upon the position on the polar diagram at which comparison between screened and unscreened transmitter was made, as the diagrams departed slightly from the theoretical shape. The magnitude and nature of the departure varied from case to case and were best observed on diagrams relating the ratio $E\Psi / E\pi/2$ to Ψ. $E\Psi$ is the field at the receiving site in a direction making an angle Ψ with the axis of the dipole. By this method of plotting the well-known "figure 8" diagram should be obtained with the longitudinal axis lying in the plane of the loop ($\Psi = \pi/2$).

Two departures from the ideal figure were apparent from these diagrams, (i) a small angular displacement of the longitudinal axis from the direction $\Psi = \pi/2$ and (ii) a difference in the diameters of the two halves of the figure 8. These can be explained by the presence of fields from (i) a horizontal magnetic dipole with its axis at 90° to that of the main dipole and moment not greater than 1/10 that of the main dipole and (ii) a vertical electric dipole of moment not greater than 1/20 that of the main dipole. Both effects were random and uncontrollable. However, so long as the stray fields are small compared with the main field, the mean value of readings within the angle $\Psi = 90° \pm 30°$ and $\Psi = 270° \pm 30°$ is practically unaffected by the presence of the stray field. The diameter of the ideal circles is approximately 1/2 db greater than this mean value. The attenuation calculated from the diameters of these circles for each of the rooms is shown in Fig. 1. The theoretical curves derived as described in the following section are shown for comparison.

Room No. 1 showed a marked difference in behaviour from the other rooms. Fig. 2 shows the results from three series of measurements made on this room and illustrates the following features: (a) the first series of measurements showing a wide departure from the theoretical value, (b) and (c) the final results obtained after effecting considerable improvement in the contacts between the room sections and showing the effect of setting the plane of the transmitting loop parallel or perpendicular to the side containing the door. These points are discussed further in the last section.

The condition of earthing of the transmitter and screening room is not specified in these figures as it had little or no effect on the field.

The effect of displacing the loop from the centre of the room was examined but only a small variation of the polar diagram was observed.

Theoretical Analysis

In the following, formulæ are developed for calculating the attentuation due to a screening room from a knowledge of certain electrical parameters of the room. To simplify the analysis the room and transmitter will be considered as a sphere with a magnetic dipole at the centre. Since the material from which the rooms were constructed is not in the form of a continuous sheet a solution from the direct application of Maxwell's equations is not practicable. An alternative approach to the problem is developed below.

At radio frequencies any magnetic shunting becomes negligible, and the attenuation of the field from the dipole is due entirely to the currents induced in the screening and may be divided between the following two cumulative actions:

(i) Attenuation due to currents induced around the circumference of the screen. This may be regarded as a short circuited transformer action or reflexion loss.

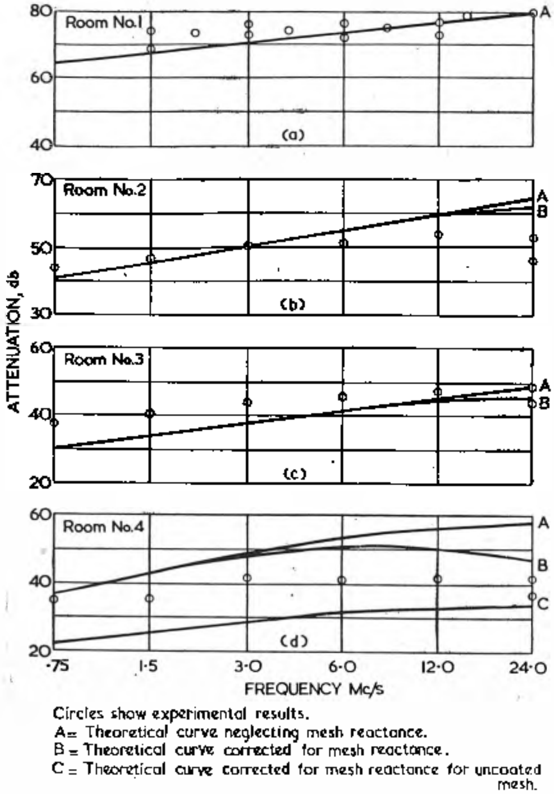
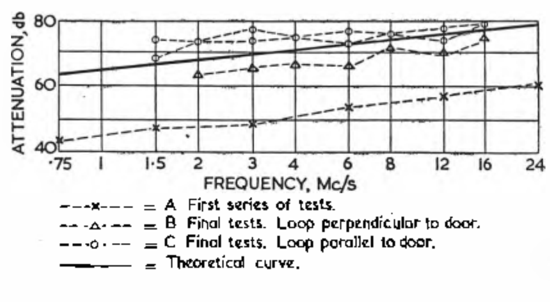


Fig. 1.
Attenuation-frequency characteristics of screening rooms

Fig. 2.
Test results on Room No. 1



(ii) Attenuation due to eddy currents induced in the thickness of the metal, i.e., skin effect or transmission loss, which is an important factor only for a screen of continuous sheet. In the case of a mesh, skin effect occurs in the metal alone and the transmission through the air holes may be considered as practically unaffected over the short distance of the mesh thickness. The mesh may therefore be regarded as a system of transmission lines of two kinds, one representing the air path and one the path through the metal, feeding a common terminating impedance. The number of lines of each kind may be taken as proportional to the fractional areas S_1 and S_2 of the air and metal paths respectively. If α_1 and α_2 be the respective attenuation constants the transmission loss is given by:

$$A^2 = 1/[S_1 \exp(-2\alpha_1 d) + S_2 \exp(-2\alpha_2 d)]$$

where A is the field attenuation and d = thickness of the screen. The attenuation constant in the metal will invariably be much greater than in the air, so if S_1 is not vanishingly small we have

$$S_1 \exp(-2\alpha_1 d) >> S_2 \exp(-2\alpha_2 d)$$

and $A^2 \approx \exp(2\alpha_1 d)/S_1$

i.e., $A \approx \exp(\alpha_1 d)/\sqrt{S_1}$

If the linear dimensions of the holes are large compared with the thickness of the screen the propagation through them will not differ greatly from that in free space, i.e., $\alpha_1 d$ will be very small, and

$$A \approx 1/\sqrt{S_1}$$

For the rooms tested the minimum value of S_1 was 0.44 so that the maximum value for A was $\sqrt{2.27}$, i.e., 3½ db. This applies to room No. 1. For other rooms the values of A were as follows: No. 2—1.5 db; No. 3—0.7 db; No. 4—0.3 db, i.e., for the last two rooms A may be regarded as unity to a first approximation. The attenuation due to skin effect may therefore be neglected and the transformer action or reflexion loss is all that need be considered. This effect may be evaluated from circuit theory or the field impedance concept. The first method of approach is particularly convenient when the screening surface is within the induction field of the radiator, i.e., when the radius of the sphere is less than $\lambda/2\pi$. Also it involves the use of more generally understood engineering concepts and since the radius of the rooms examined was always less than $\lambda/2\pi$ this method has been adopted here. Both arguments of course give identical formulæ.

The field strength E due to a magnetic dipole is proportional to its moment M. For two concentric coaxial dipoles the total external field is

$$\bar{E} = \bar{E}_1 + \bar{E}_2$$

where the bar denotes a complex quantity and the suffices denote primary and secondary dipoles respectively. The attenuation of the primary field due to the coupled secondary dipole is then

$$\bar{A} = \bar{E}_1/(\bar{E}_1 + \bar{E}_2) = \bar{M}_1/(\bar{M}_1 + \bar{M}_2) = 1/(1 + \bar{M}_2/\bar{M}_1) = 1/(1+\bar{a})$$

where $\bar{a} = \bar{M}_2/\bar{M}_1 = m_2\bar{I}_2/m_1\bar{I}_1$

and m = magnetic moment per unit current in the circuit.

For two coupled circuits we have the relation

$$\bar{I}_2/\bar{I}_1 = -Z_{12}/Z_{22}$$

where Z_{22} = self-impedance of secondary circuit, and Z_{12} = mutual impedance betwen the circuits.

For the case considered

$$Z_{12} = j\omega L_{12}$$

(L_{12} = mutual inductance) and Z_{22} may be written as

$$Z_{2i} + Z_{2e} = Z_{2i} + j\omega L_2$$

where Z_{2i} = internal impedance of secondary circuit, i.e., impedance associated with the fields in the conducting surface.

Z_{2e} = external impedance of secondary circuit, i.e., impedance associated with the field in space.

L_2 = external inductance.

Then $\bar{I}_2/\bar{I}_1 = -j\omega L_{12}/(Z_{2i} + j\omega L_2) = L_{12}/L_2(jZ_{2i}/\omega L_2 - 1)$

At low frequencies Z_{2i} is simply R_2, the resistance of the secondary circuit. In general Z_{2i} may be related to R_2 by a complex factor $\bar{K}$ such that $Z_{2i} = \bar{K}R_2$. It follows that:

$$Z_{2i}/\omega L_2 = \bar{K}R_2/\omega L_2 = \bar{K}D_2$$

where $D_2 = 1/Q_2$ is the dissipation factor. Hence we may write

$$\bar{a} = m_2 L_{12} / m_1 L_2(j\bar{K}D_2 - 1) = a_0 / (j\bar{K}D_2 - 1)$$

The quantities a_0 and D_2 may be evaluated for a spherical sheet and for a mesh of a spherical form on the assumption that the current distribution remains essentially the same as in a sphere of continuous sheet.

The case of a conducting spherical surface with a dipole in the centre has been solved analytically by H. Kaden[4] who has shown that the secondary field H_2 in the sphere, due to the currents induced on its surface, is parallel to the axis of the dipole and uniform over the volume of the sphere. It may further be shown[4] that the distribution of the induced currents to produce this field is given by

$$i_2 = \hat{i}_2 \sin\theta$$

and $H_2 = 2\hat{i}_2/3$

where i_2 = linear current density along a circular path in a plane perpendicular to the axis of the dipole.

$\hat{i}_2$ = equatorial value of i_2.

θ = ½ the linear angle subtended by the current path at the centre of the sphere.

Further: $I_2 = \int_0^{\pi} i_2 r_2 \sin\theta \, d\theta = 2r_2\hat{i}_2$

$$m_2 = M_2 / I_2 = (1/I_2)\int_{I_2} S_2 dI_2 = 2\pi r_2^2/3$$

$$m_1 = N_1 S_1$$

$$L_2 = (3/2I_2^2)\int_{V_2} \mu_0 H_2^2 dV_2 = 2\pi\mu_0 r_2 / 9$$

$$L_{12} = \mu_0 H_2 N_1 S_1 / I_2 = N_1 S_1 \mu_0 / 3r_2$$

where r_2 = radius of sphere

S = area enclosed by current path.

N = number of turns.

V_2 = volume of sphere.

μ_0 = permeability of free space $= 4\pi \times 10^{-7}$ H/metre.

Hence $a_0 = m_2 L_{12} / m_1 L_2 =$

$$(2\pi r_2^2/3)(N_1 S_1 \mu_0/3r_2) / (N_1 S_1)(2\pi\mu_0 r_2/9) = 1$$

and $\bar{A} = 1 / [1 + 1/(j\bar{K}D_2 - 1)] =$

$$-(1 - j\bar{K}D_2) / j\bar{K}D_2 = jQ_2/\bar{K} + 1$$

i.e., $A = |\bar{A}| \simeq Q_2/K$ where $K = |\bar{K}|$

The attenuation is therefore equal to the quality factor of the room provided K does not differ greatly from unity.

The value of R_2 the effective resistance of the room, required to calculate Q_2, is given by

$$R_2 = (1/I_0^2)\int_{S'} R_s' \, i_2^2 dS' = 2\pi R_s' / 3$$

where S' refers to the surface of the sphere and R_s' represents the effective surface resistance of the mesh. R_s' may be related to R_s, the surface resistance of the material from which the mesh is made, by the quantity C_n, where

$$R_s' = R_s / C_n = 1/\sigma t C_n$$

where σ = volume conductivity of the mesh material.
t = thickness of an equivalent conducting layer having uniform radial current distribution.

For a continuous sheet

$$\frac{t}{K} = (\delta/\sqrt{2})\sqrt{\frac{\cosh 2p - \cos 2p}{\cosh 2p + \cos 2p}}$$

where $p = d/\delta$ and δ = depth of penetration.

C_n is a function of the mesh size and shape. Most forms of mesh or perforated sheet may be simulated by two (or more) interlaced sets of parallel strips or wires, at an angle β to each other. The surface resistance of such a mesh depends on the direction of the current flow (unless $\beta = 90°$) but a close approximation to the mean value is given by the resistance in the direction of one set of wires, and when $d > \delta$

$$R_s' = R_s c / b \;\therefore\; C_n = b/c$$

where c = spacing between the wires.
b = width of each strip or diameter of wire.

The expression for attenuation then becomes

$$A \approx Q_2/K = \omega L_2 / KR_2 = C_n \omega \mu_0 r_2 / 3R_s = (2\pi/3) f \mu_0 r_2 \sigma t (C_n/K)$$

At frequencies such that $d < \delta/2$, when the conductor is electrically thin and there is no skin effect, t is constant and equal to d. The internal reactance is then negligibly small and $K = 1$. In this region, therefore,

$$A = (2\pi/3) f \mu_0 r_2 \sigma d C_n$$

i.e., the attenuation is proportional to frequency.

At frequencies such that $d > 2\delta$ (which was the case during the tests) the skin effect is practically fully developed and

$$t = \delta = 1\sqrt{(\pi f \mu \sigma)}$$

(where $\mu = \mu' \mu_0$ and μ' = permeability relative to free space). At the same time the internal reactance increases to a value numerically equal to R_2, and $\bar{K}$ becomes equal to $1 + j$, i.e., $\bar{K} = \sqrt{2}$. The attenuation is then

$$A = 2C_n r_2 \sqrt{\pi f \mu_0 \sigma / \sigma'} \;/\; 3\sqrt{2}$$

Writing $\sigma = \sigma' \sigma_0$, where σ_0 is the conductivity of copper $= 58 \times 10^6$ mhos/metre, and putting $\mu_0 = 4\pi \times 10^{-7}$ H/m, we obtain when $r_2 < \lambda/2\pi$* and $d > 2\delta$

$$A = 10^4 \, r_2 C_n \sqrt{\sigma' f / \mu'} \;/\; \sqrt{2}$$

where f denotes frequency in Mc/s. The attenuation is now proportional to $(\text{frequency})^{\frac{1}{2}}$.

The formula just derived was obtained on the assumption of uniform secondary field inside the sphere. This will be correct for a very close mesh, but with an open mesh the field will be distorted at the mesh surface. It is convenient to consider this distorted field as composed of the uniform field of a continuous sphere, having superimposed on it a local distortion field due to the non-uniformity of the conducting surface. The uniform field gives rise to the reactance ωL_2 as calculated previously, while the distortion field gives rise to an additional reactance, which will be capacitive for wires perpendicular to the direction of the electric field, and inductive for wires parallel to it. The former may be neglected for the frequency range covered by this report, but the latter can explain the flattening, or even falling off, of the attenuation characteristics observed with some of the rooms.

Macfarlane[5,6] has developed the following formulæ for the surface reactance of an infinite plane grating when energised by a plane wave, with the electric field parallel to the conducting strips

$$X_s' = \sqrt{\mu_0/\epsilon_0}\,(c/\lambda)\,\ln \operatorname{cosec}(\pi b/2c)$$

provided that $c << \lambda$ and $b/c << 1$

This can be written as

$$X_s' = (1/2)\sqrt{\mu_0/\epsilon_0}\,(2c/\lambda)\,\ln \operatorname{cosec}(\pi b/2c)$$
$$= (1/2)\sqrt{\mu_0/\epsilon_0}\; x$$

$(1/2)\sqrt{\mu_0/\epsilon_0}$ may be recognized as the surface impedance of surrounding space to an infinite plane sheet with uniform current distribution ($\sqrt{\mu_0/\epsilon_0}$ on each side of the sheet, the two sides being in parallel), i.e., external surface impedance of the sheet. This impedance may be denoted in general as Z_{se}, giving the expression

$$X_s' = Z_{se}\, x$$

Pending a rigorous solution, it will be assumed that the function x is invariant with respect to the shape of the surface formed by the grating, provided that this surface is unbounded and the conductors are everywhere parallel to the electric field. The value of X_s' is then found in each case by evaluating and substituting for Z_{se} the impedance corresponding to the particular surface and current distribution.

It was shown that the external impedance of a spherical surface energized by a spherical magnetic wave is

$$Z_{2e} = j\omega L_2 = j2\pi\omega\mu_0 r_2/9 \quad (r_2 < \lambda/2\pi)$$

Also, the internal resistance was found to be

$$R_2 = 2\pi R_s/3$$

The relation between total and surface quantities must be the same for internal and external impedances, thus we can write

$$Z_{2e} = 2\pi Z_{se}/3$$

hence $Z_{se} = j\omega\mu_0 r_2/3$
for a sphere.

The mesh surface reactance X_s' now becomes

$$X_s' = (\omega\mu_0 r_2/3)x$$

This reactance is, of course, external to the conductor, but as it is a function of the mesh size and shape it will be more convenient to associate it with the internal impedance, and include its effect in the correcting factor $\bar{K}$ which now becomes

$$\bar{K} = [(1+j)R_s' + jX_s']/R_s'$$
$$= 1 + j(1 + X_s'/R_s')$$
$$= 1 + j[1 + (\omega\mu_0 r_2/3R_s')x]$$
$$= 1 + j(1 + Q_2 x)$$

and
$$x = (2c/\lambda)\,\ln \operatorname{cosec}(\pi b/2c)$$
$$= f\,(2c/300)\,\ln \operatorname{cosec}(\pi b/2c)$$
$$\approx 10^2\,(f/6n)\,\ln \operatorname{cosec}(\pi C_n/2)$$

* Errors are small up to $r_2 = \lambda/2$ but increase rapidly for values of $r_2 > \lambda/2$.

where n is the mesh number, i.e., the number of conducting strips per inch. In practice the conducting strips may have thickness comparable with their width, or round wires will be used. Now it can be proved[1,3] that with $C_n << 1$ a round wire is equivalent to a strip of width b' equal to twice the wire diameter. Generalizing, the conductor may be considered as an ellipsoidal cylinder of mean diameter

$$s = (b + d)/2$$

The expression for mesh reactance is now

$$X_s' = 10^{-3}\ (Q_2 f/6n)\ ln\ \text{cosec}\ [(\pi/2)(2s/c)]$$
$$= 10^{-3}\ (Q_2 f/6n)\ ln\ \text{cosec}\ (\pi C_n'/2)$$

where $C_n' = (1 + d/b)C_n$

For very thin strips $C_n' = C_n$, and for round wires $C_n' = 2C_n$.

To extend the application of the attenuation formula, the case of a composite material must also be considered. If Z_{s1} and Z_{s2} be surface impedances (internal) of the outer and inner material respectively, d_1 and δ_1 the thickness of, and depth of penetration in, the coating and Z_{st} equivalent surface impedance of the composite material, then

$$Z_{st} \approx Z_{s2} \text{ when } d_1 << \delta_1$$
$$Z_{st} \approx Z_{s1} \text{ when } d_1 \geqslant \delta_1$$

and in the region of transition from one metal to another, the values of σ' and μ' for the inner material may be used together with a correcting factor C_t, where

$$C_t = |Z_{s2}/Z_{st}| = (Z_{s2}/Z_{s1}) \sqrt{\frac{\cosh\ (2p_1+\beta) - \cos\ 2p_1}{\cosh\ (2p_1+\beta) + \cos\ 2p_1}}$$

where $p_1 = d_1/\delta_1$ and $\beta = ln[(Z_{s2}+Z_{s1})\ /\ (Z_{s2}-Z_{s1})]$

To summarize the foregoing, the attenuation of a screening room, constructed from a mesh of conducting material, may be given by

$$A = Q_2/K = 10^4\ r_2\ (C_n/K)\ \sqrt{\sigma' f/\mu'}$$

where $2r_2$ = diameter of equivalent sphere = geometric mean of the room dimensions[2],
σ' = conductivity relative to copper,
μ' = permeability relative to free space,
f = frequency in Mc/s.
$K = |1 + j[1 + 10^{-3}\ (Q_2 f\ /6n)\ ln\ \text{cosec}\ (\pi C_n'/2)]|$
$C_n = b/c$ where b = width of strips or diameter of wire,
c = spacing between wires,

If the material from which the mesh is constructed is composite (e.g., tinned copper, galvanized iron), the above value for A must be multiplied by the factor C_t, given above, and σ' and μ' then refer to the inner material.

Discussion and Conclusions

Comparison of the theoretical and measured attenuation characteristics in Fig. 1 shows a general agreement in shape and magnitude. This would suggest that errors due to approximations in the theoretical argument are of the same order as the experimental errors.

Outstanding features of Room No. 1 were its instability and very high peak values of attenuation. The external inductance, L_2, for all rooms was of the order of 1μH, i.e., a reactance of about 6Ω at 1 Mc/s. It follows that a resistance of less than 6mΩ is required to produce the theoretical attenuation of 60 db at this frequency. With such a low value of total resistance in the circuit, the contact resistance between its parts becomes a matter of primary importance. Moreover, when the low surface resistance has been obtained by the use of zinc in the form of perforated sheets, the flat and easily oxidized surfaces are particularly liable to produce a bad contact. Even after inserting copper braiding between most of the joints, the readings varied over a range of 10 db. For the same reason attenuation was highest with the loop parallel to the door when large currents could be induced in the door even if it were completely isolated from the rest of the room.

The characteristics of the other rooms show a flattening or even falling off in attenuation at higher frequencies. This may be explained by the mesh reactance X'_s becoming comparable with, and even greater than, the resistance R'_s. For each mesh there will be a frequency range within which X'_s is much less than R'_s. In this region the conductivity of the mesh material will determine the attenuation and the spacing of the wires is immaterial if C_n is kept constant. On the other hand, for frequencies at which $X'_s >> R'_s$ the attenuation becomes independent of the conductivity and inversely proportional to the spacing of the wires (again if C_n is kept constant), the formula becoming

$$A = \omega L_2/X'_s = \omega L_2\ /\ \omega L_2\ x = 1/x = 6n\ 10^3\ /\ f ln\ \text{cosec}\ (\pi C_n'/2)$$

Over a range of frequencies from very low to very high values the attenuation characteristics will vary from proportionality to frequency through a region of proportionality to (frequency)$^{\frac{1}{2}}$ to proportionality to (frequency)$^{-1}$. The frequencies at which transition from one law to another will take place are defined by the conductivity and permeability of the screening material and by the mesh dimensions. If the total resistance of the screen is kept constant, the upper frequency limit of the region of proportionality to (frequency)$^{\frac{1}{2}}$ may be raised by using a closer mesh. Thus if two screens be made, one of a close iron mesh and the other of a more open copper mesh, the value of R'_s being the same in both cases, the two attenuation characteristics will coincide at lower frequencies, but that for the close mesh of inferior material (iron) will be much better at higher frequencies.

The effect of mesh reactance was not observed in Room No. 1 and has not been calculated since the approximations used in deriving the formula for X'_s are not valid when C_n approaches unity.

The attenuation characteristic of Room No. 2 falls off at higher frequency more rapidly than expected. This suggests that the mesh reactance is greater than that given by the formula for X'_s. The flat diamond shape of this mesh must distort the field close to the mesh more than the square mesh having the same C_n. This result is, therefore, to be expected.

The rather low theoretical value for Room No. 3 suggests that too high a value of μ' (64) was assumed for iron in the calculation. The effect of a non-magnetic coating on a ferro-magnetic material is shown in Fig. 1(d) where the theoretical curve for the same mesh uncoated is plotted (curve C). The coating of this mesh was very uneven, having been obtained by dipping. The mean thickness (obtained by chemical analysis) was used in the calculation of curves A and B, but the experimental results suggest that the effective thickness was rather less.

As only one size of room was examined, the effect of room size on attenuation could not be observed. It has been shown theoretically, however, that A varies as r_2 provided $r_2 < \lambda/2\pi$ (and $X'_s < R'_s$). The dependence of A on r_2 becomes more complicated when $r_2 \geqslant \lambda/2\pi$, but the formula developed can be used up to $r_2 = \lambda/2$ (i.e., $f = 150$ Mc/s. for $r_2 = 1$ m) with an error not greater than -3 db.

The analysis has been made for the case when no conductors pass through the walls of the room. It provides not only a qualitative explanation of the behaviour of the room, but leads also to a fairly simple formula for calculation of the attenuation. Several approximations have been made and the analysis is not regarded as rigorous. Close agreement, however, has been obtained with experimentally determined values. In particular, the analysis applies only when the discontinuities introduced by joints between the various sections of the room may be neglected. Neglecting these discontinuities the performance of the room naturally improves as the impedance of the material used in its construction is made smaller. In practice, however, a limit is imposed by the difficulties met in obtaining a sufficiently low contact resistance between the sections as the impedance of the material is reduced. Thus when a very high attenuation is required it may be more economical to use a double screen of higher impedance material.

In general, the best performance will be obtained by using a material of high conductivity, low permeability and small mesh size. In some circumstances, however, the properties of the conductor from which the mesh is made may have little or no effect on the performance of the room. A particular point to note is that whereas in a room constructed from continuous sheet the skin effect improves the performance, the contrary takes place when the screening material is in the form of a mesh.

Since completing the above work the analysis has been carried out in a more rigorous manner and has been extended to cover all frequencies, and experiments are being devised to check the theory at the higher frequencies. It is hoped that data on the performance of screening rooms constructed from a conductor in the form of a mesh can be tabulated from the results of these tests and thus provide useful information for designers.

Acknowledgments

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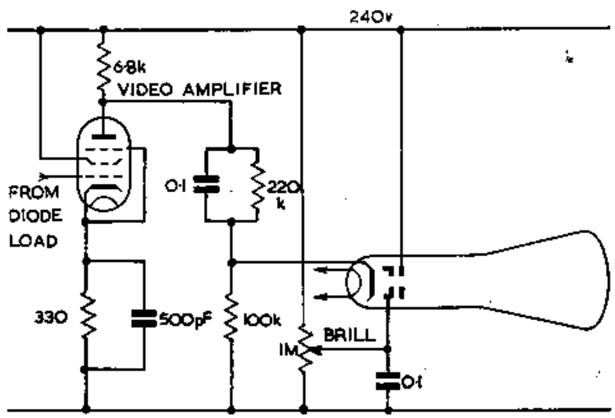
Semi-Stabilized Black Level for Television Receivers

By C. H. BANTHORPE. (*Central Equipment, Ltd.*)

THE mean potential of the video waveform of a television receiver is normally above black level, and during usual programmes stays fairly steady, only changing markedly for short intervals. If, therefore, the C.R.T. is A.C. coupled to the video stage, the brilliance control has to be advanced if the contrast control is advanced. If, however, it is D.C. coupled, the control must be reduced if the contrast is increased. It is possible therefore by using A.C./D.C. coupling for the black level to stay constant for different contrast levels, the brilliance control merely setting black level.

This simple arrangement will only work if the mean potential of the video waveform does stay steady, and, as stated, in practice this is nearly so. With this circuit, once the brilliance is pre-set, increasing the contrast makes the white whiter, and below blacks blacker, which is what is required if, for instance, the picture is viewed with more light on it during an afternoon transmission. If the picture is viewed in darkness, the contrast control would be reduced, the whites becoming less white, and below black less black.



The circuit has other advantages, one being that it is possible to obtain the $A1$ voltage for the C.R.T., if a tetrode, without having a high H.T. line, since the cathode of the C.R.T. is now considerably below H.T. and the $A1$ to cathode potential increases during whites, reducing defocusing. Finally, if the D.C. component is fed to the cathode of the C.R.T. via a long time-constant, the brilliance flutter which occurs because of reflexions from aircraft can be greatly reduced. The use of a long time-constant in this way seems to have been first suggested by E.M.I. in *Wireless World*, April 1949.

A Valve Voltmeter Circuit

By R. KITAI, M.Sc. (Eng.) (Rand)

THE voltmeter described was designed for use in the students' laboratory, the requirements to be met being versatility in voltage and frequency ranges, and low cost. In view of the fact that the instruments were to be used mainly by inexperienced students, particular attention was paid to overload protection and rugged construction, and a somewhat insensitive meter-movement (0-1 mA) was therefore aimed at.

Specifications

(i) A.C. Voltage Ranges

Four ranges giving full-scale deflexions of 1, 5, 25 and 100 volts R.M.S. A.C. The instrument is peak-reading with a scale calibrated to read 0.707 of the peak of a sinusoidal A.C. voltage. The 1 V range is non-linear, but by off-setting the zero (as described below) the linear meter scale may be used to give direct readings of 0.2-1 V on this range. A.F. voltages (20 to 20,000 c/s.) are applied to terminals on the meter panel. R.F. voltages (5 kc/s. to approximately 300 Mc/s.) are applied to a probe containing a separate D.C. peak conversion circuit.

(ii) D.C. Voltage Ranges

Four ranges giving full-scale deflexions of 1, 5, 25 and 100 volts. All ranges are linear. The input resistance is 50 megohms.

(iii) Protection

The current through the milliammeter cannot exceed approximately 1.3 mA on all A.C. ranges, and cannot exceed approximately 2 mA on all D.C. ranges for any overload condition.

(iv) Accuracy

Within 2 per cent of full-scale deflexion on all ranges. The H.T. supply voltage is regulated to provide operation at mains voltages between 200 and 250.

(v) Zero Adjustment

If correctly zeroed on the 1 V range A.C. or D.C., the instrument is automatically correctly zeroed on all other ranges without the aid of additional pre-set controls.

(vi) Meter Scales

The same scales are used for A.C. and D.C. voltage readings.

Circuit

Referring to Fig. 1, it will be noticed that two similar circuits are used for the conversion of A.F. and R.F. voltages to peak D.C. voltages. Few components are needed in the circuits, while the important advantages of a separate circuit for R.F. measurements are threefold:—

(a) A short time-constant circuit can be used giving a fast response of the indicating meter to fluctuations in the R.F. signal voltage.

(b) Difficulty in mechanical construction is avoided, it being unnecessary to provide a special receptacle in the apparatus into which the probe is plugged when connexions are made to terminals on the meter for A.F. voltage measurements. The probe is simply stored in the removable lid of the instrument and is plugged in when needed.

(c) Since a short time-constant circuit is used, a small capacitor C_1 results in a very compact probe. (The probe constructed is 4 in. long and 1 in. in diameter.)

The calibration of the instrument is not dependent on the component values in the circuits within their frequency ranges.

Since the remainder of the circuit concerns itself with the registering of the D.C. voltages developed across R_1 and R_4, it is possible to measure D.C. voltages developed across a high resistance (R_6) switched into the same circuit. The polarity of the D.C. voltages across R_1 and R_4 necessitates the grounding of the D.C. positive terminal. The instrument is made to read on the same meter scales on A.C. and D.C. by shunting the milliammeter (R_8) when measuring A.F. and R.F. voltages.

Valve V_4 is essentially an impedance transformer operating with a small negative bias under quiescent conditions, an impedance transformation of 50 megohms to 1 megohm being obtained. R_7 is the zero-adjustment control, while R_9 is a screwdriver adjustment used to reset R_7 when valves are changed.

The second impedance transformation takes place in valve V_5 and its circuit, the cathode load being 15,000 ohms. This low impedance circuit permits the use of the 0-1 milliamp meter, with its associated low values of range change resistors. (The highest resistance required in this circuit is of the order of 70,000 ohms. Adjustable 5 watt or 10 watt wire-wound resistors are ideal and can be easily set with the precision required.)

Overload Protection

The voltage-overload protection circuit comprises the resistors R_{16} to R_{21} and the diode portion of V_4. The static bias of the diode is adjusted so that the diode conducts only when the meter current is, say, 1.3 times that required for full-scale deflexion on any particular A.C. range. Since the diode resistance (when conducting) and the resistances R_{15} to R_{20} are considerably lower than 1 megohm in the cathode circuit of V_4, the cathode of V_4 is held virtually at earth potential and the triode section of V_4 cuts off with a much lower input voltage than would otherwise be the case.

A further protection circuit is necessary for the period between switching on and the commencement of electron emission in V_5. Should the H.T. supply build up before V_5 conducts, valves V_6 and V_7 will fire, giving rise to a severely unbalanced bridge and overloading of the meter on the lower voltage ranges. To overcome this the rectifier V_3 is of the indirectly

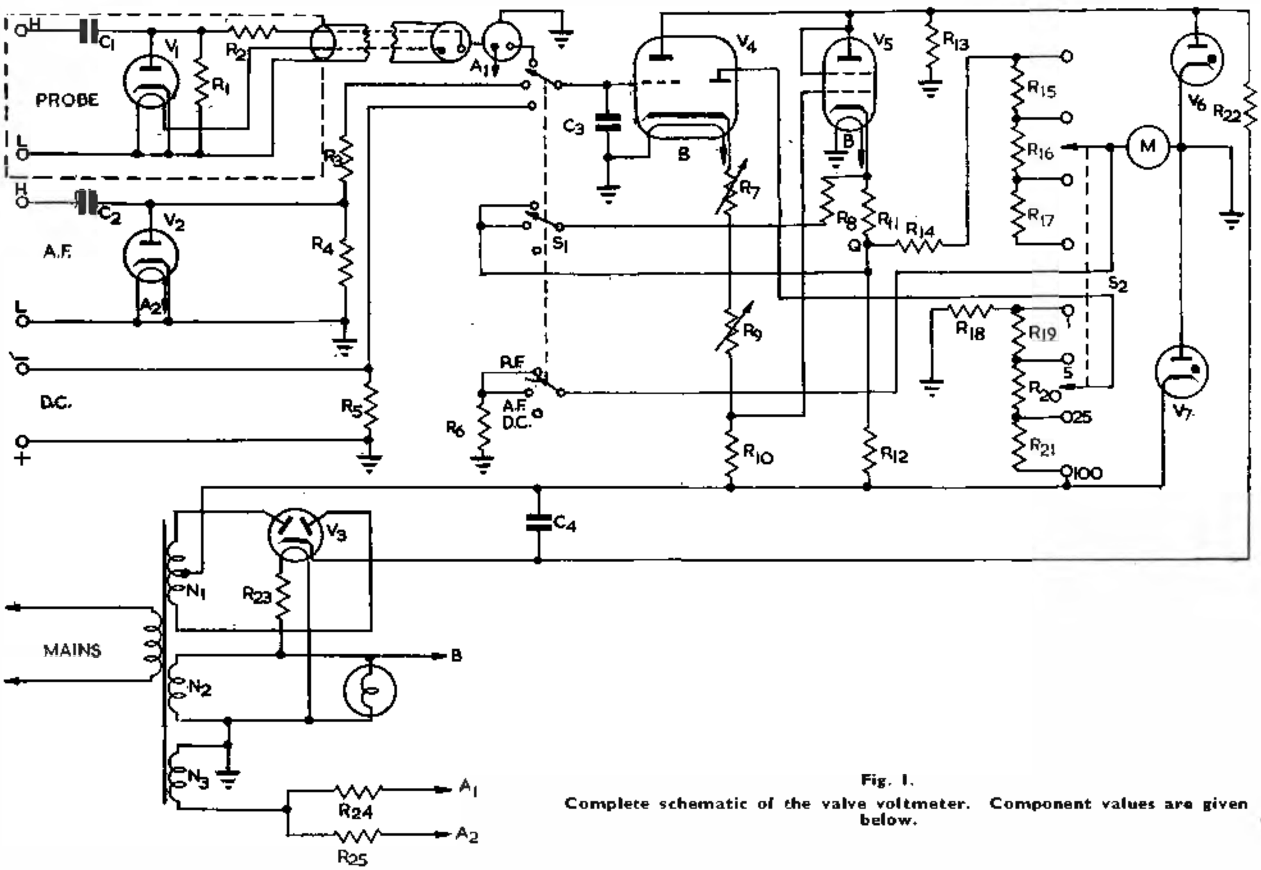


Fig. 1.
Complete schematic of the valve voltmeter. Component values are given below.

- R_1 20M ¼ watt
- R_2 10M ¼ watt
- R_3 10M ¼ watt
- R_4 50M ¼ watt
- R_5 50M ¼ watt
- R_6 Approx. 100Ω wire-wound. (Value dependent on meter used)
- R_7 10k wire-wound rheostat
- R_8 Approx. 2,500Ω wire-wound
- R_9 20k wire-wound rheostat
- R_{10} 1M ½ watt
- R_{11} 400Ω wire-wound
- R_{12} 15k 10 watt wire-wound
- R_{13} 25k 10 watt wire-wound
- R_{14} Approx. 10Ω (Value dependent on meter used)
- R_{15} Approx. 3k wire-wound
- R_{16} Approx. 15k wire-wound
- R_{17} Approx. 60k wire wound
- R_{18} 250Ω ¼ watt
- R_{19} 2k ½ watt
- R_{20} 10k 2 watt
- R_{21} 25k 10 watt wire-wound
- R_{22} 4k 10 watt wire-wound
- R_{23} 2Ω wire-wound
- R_{24} 5Ω wire-wound
- R_{25} 5Ω wire-wound
- C_1 0.002μF mica
- C_2 0.02μF mica
- C_3 0.01μF mica
- C_4 2μF paper
- V_1 Mazda D1
- V_2 Mazda D1
- V_3 6X5GT
- V_4 6Q7GT
- V_5 6V6GT
- V_6 VR90 (OB3)
- V_7 VR150 (OD3)
- M 0-1 milliammeter (Ernest Turner Model 505 scaled to read 0-1, 0-5, 0-25)
- N_1 350-0-350V 60 mA
- N_2 6.3V 2A
- N_3 5V 1A
- S_1 3-pole 3-position switch
- S_2 2-pole 4-position switch

heated type, its heater being under-run at a voltage of 5.9 instead of 6.3 volts, and the building of the H.T. is thereby sufficiently delayed.

Non-Linearity on 1 Volt A.C. Range

The non-linearity of the diode probe and A.F. circuit at low potentials can, of course, be overcome by the inclusion of a special meter scale for this range. It seems more expedient, however, to utilize the large linear portion of this range for direct reading on the meter scale. Curve A in Fig. 2 is the meter deflexion *vs.* applied voltage, the resistance R_{14} being adjusted to provide full scale deflexion with 1 V R.M.S. input, the instrument being initially zeroed with no current through the meter. The region 0.15 V to 1 V of the curve is linear, while voltage readings below 0.15 V are never reliable unless the diode heater currents are carefully regulated. It is possible to obtain voltage readings from approximately 0.2 volt to 1 volt directly on the meter scale by off-setting the zero and readjusting resistance R_{14}. The procedure is as follows:—

On the 1 V range and with zero input voltage, the meter current is adjusted to 0.1 mA. Resistance R_{14} is now adjusted to provide full-scale deflexion with 1 V R.M.S. input, giving the curve B in Fig. 2. On the 0-5 V range the off-setting of 0.1 mA is reduced to 0.02 mA due to the additional resistance (R_{15}) in the meter circuit. This error can usually be neglected, but may also be taken into account by readjusting R_{15} slightly to give the correct full-scale deflexion with this slightly off-set zero.

The value of R_{14} arrived at by this procedure is that value needed for direct measurement of D.C. *voltages on the 0-1 volt range with shunts R_6 and R_8*

out of circuit, the meter being then zeroed on 0 mA, and the entire scale being linear.

On R.F. and A.F. ranges then, the range switch is turned to the 1 volt position and the pointer adjusted to read 0.1 mA with zero input voltage. The instrument is then correctly zeroed for all A.C. ranges and provides direct reading of A.C. voltages between 0.2 V and 100 V R.M.S. On D.C. ranges the range-change switch is turned to the 1 volt position and the pointer adjusted to zero.

Finally four factors instrumental in reducing the complexity and cost of the voltmeter might be mentioned:—

(1) The procedure of zeroing the instrument on A.C. ranges permits the use of the same scale for all D.C. ranges with the removal of two shunt resistances. No special meter scales are required.

(2) The low values of the range-change resistors permit the use of inexpensive wire-wound resistors in common use.

(3) Valves V_6 and V_7 perform the three functions of regulating the H.T. supply, reducing ripple in the H.T. supply, and providing the second fixed potential point.

(4) The instrument is constructed from standard components which are in easy supply.

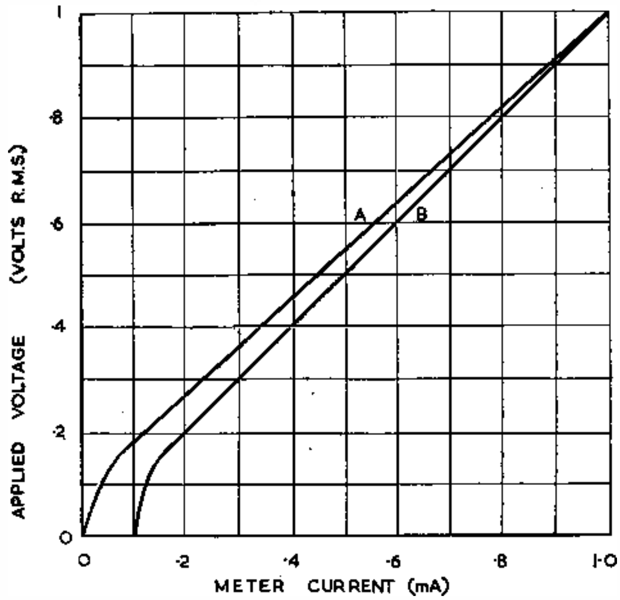


Fig. 2.

Curve B shows how the off-set zero provides direct reading from 0·2V to 1V A.C. on the linear meter scale

Silicones and Other Silicon Compounds

—A GENERAL REVIEW By D. W. THOMASSON

TO judge from the variety of notes on their properties which have appeared in recent years, one might be excused for thinking that the silicones were a panacea for all manufacturing ills. The radio industry seems somewhat doubtful of this, which is understandable, since it frequently suffers from the effects of over-enthusiasm within its own ranks.

Nevertheless, the silicones have many useful applications in radio manufacture, and can be of considerable value if their qualities are properly understood. The necessary understanding can only be gained by examining the origin and nature of this group of materials.

Manufacture

The first step in the manufacture of silicones is the preparation of silicon tetrachloride ($SiCl_4$). This is obtained from sand, coke and chlorine, the coke acting as a catalyst.

The second stage is the replacement of one or more of the chlorine atoms by a hydrocarbon group, forming an organo-silicon chloride. This is modified by hydrolysis and condensation to produce the silicone.

It is clear that a wide variety of organo-silicon chlorides can be produced, since there are many different hydrocarbon groups which can be used, and each offers three alternative chlorides, according to the number of chlorine atoms replaced. The versatility of the silicones is thus made more reasonable, since there is at least one for each organo-silicon chloride.

If all the chlorine atoms are replaced, another range of substances can be obtained, of which ethyl silicate ($Si(OC_2H_5)_4$) is perhaps the most important. This substance has been known for a little more than a hundred years, and, though industrially useful, has given so much trouble through instability that it has been little used.

It is not impossible that this has given rise to some of the distrust of silicones. It will be shown later that a way has been found of making ethyl silicate reliable enough for industrial use, but the silicones need no such provision. Stability is one of their most important features.

Characteristics

To understand why silicones are so stable, it is necessary to know their structure. This is based on a framework of silicon-oxygen-silicon linkages, which is also found in glass, quartz, flint, porcelain, and other well-known silicon compounds. All these exhibit the same stability, which arises from the strength of the basic linkages.

In particular, the silicones have a high heat stability, similar to that of quartz glass. They also have high specific resistances, as might be expected, and are suitable for H.F. insulation purposes.

They range from oils to hard moulding materials in consistency, and include rubbers, lacquers, greases, thermo-setting resins, and paint bases. All these share in some measure the useful features of the natural silicon compounds, while adding a few of their own.

Applications

Each type of silicone has its own special field of usefulness. The oils and greases, for example, have features which make them valuable for use in vacuum apparatus. Some have extremely low vapour pressures and enable very low pressures to be reached. Furthermore, they do not decompose at high temperatures in the presence of air, permitting a more rapid change-over of the pumping system, without waiting for the pump to cool.

The oils and greases are also useful for high temperature lubrication, as they retain their characteristics up to 300° C., while the working range extends far below zero in the opposite direction.

The rubbers retain their elasticity at temperatures up to 270° C., and have been used for gaskets in jet engines. They are useful where elasticity is required in the neighbourhood of hot surfaces, as in valve retainers and cushions.

The resins may be used for wire insulation, when temperatures up to 180° C. are permissible. Such insulation would not suffer from accidental contact with the soldering iron, and would be valuable in many awkward corners where ordinary plastic insulation will not last. The rubbers may also be used for insulation purposes.

For transformers, there are many interesting possibilities. The radio industry has recently been exhorted to improve its standards of tropicalization in respect of such components. The best way of doing this is to employ silicone varnish for wire insulation, the winding being impregnated with similar varnish on completion. For the larger types, immersion in silicone oil would provide the final touch.

The result, with or without the oil, is *real* tropicalization. If the oil is not used, the transformer can be finished with a coat of silicone moisture-repelling varnish. Water will then run away as it is said to run off the proverbial duck's back before permeation can occur. Meanwhile, the working temperature can be doubled.

Glass fibre insulation may be improved by impregnation with silicone varnish. The flexibility and strength are increased without loss of the glass characteristics.*

A last example, iron dust cores are liable to serious damage from damp when protected only by "impermeable varnishes" or thin wax coatings. If the dust is bonded in silicone resin, the effect is the same as would be obtained by sealing in glass. Such cores can withstand the worst humidity conditions.

Other Silicon Compounds

These notes would not be complete without some mention of the uses of ethyl silicate and its derivatives.

The value of ethyl silicate lies in its power to deposit silica. It reacts with water to give silicic acid and ethyl alcohol. On heating, the ethyl alcohol evaporates, and the silicic acid dehydrates to produce a very fine and highly reactive form of silica.

This silica may be used in conjunction with suitable "fillers" to produce a variety of substances. If ethyl silicate, sillimanite, and water are mixed in correct proportions, and poured into a mould, a jelly-like substance is produced. This may be dried and baked to produce a substance which is rock-hard, and suitable for use in precision moulds for casting metal. A milling cutter made by this process by copying from an existing cutter was ready for use without machining. A tolerance of ±0.002 in. per linear inch can be obtained with these moulds.

Other filler materials produce different kinds of finish. An example is a material having the texture and electrical characteristics of porcelain, and the strength of cast iron.

It has been mentioned that certain difficulties arise in the use of ethyl silicate in industrial applications. In its original form it is unstable in the presence of moisture, and may "gel" at any time without warning.

This has been overcome by the production of a series of modified silicon esters, which retain the characteristics of depositing silica by reaction with water, but are standardized for reaction times by the addition of inhibiting agents.

These modified esters, which have a long "shelf-life", are produced under the name "Silesters" by Silicon Developments, Ltd. Several varieties are available, having different reaction times and giving different grades of silica.

The applications of these esters and a thermosetting resin derived from them have not been explored in full as yet. Their use requires considerable care for good results, but it is to be hoped that they will justify their present promise of being valuable materials.

The Other Side

After giving so much credit to the silicones, it is necessary to end with a corrective. All these good points exist, and are of value, but there are bad points, too.

The silicones are relatively new materials, and their acceptance is not encouraged by the lack of adequate working information. It would be a great help if specific details of the characteristics of the different varieties could be obtained.

Another point is the supply difficulty. Silicones are obtainable in this country, but not as readily as one might wish. The vacuum oils can be obtained from W. Edwards & Co., Ltd., and various firms supply the varnishes and resins.

Not all of the varnishes and resins available are of top quality, however, and their use is hampered by the lack of information mentioned above. With better quality products and properly documented information, the silicones would find acceptance far more readily.

Conclusion

The "corrective" was placed last, for it is of great importance. It would not be fair to end, however, without reaffirming the intrinsic value of silicones. The value is there, though clouded by inadequate knowledge of its limits.

As with so many other developments, the Americans have made the silicones their own speciality, if not monopoly. It is to be hoped that the production of these valuable materials will receive deserved attention in future, at the expense of some of the earlier "boom" materials, which have already been tried and found wanting. That would be better than putting up with the best we can get.

* A recent article by Hoptroff and Rosevear (*J.I.E.E.*, III, 97, 191, May, 1950) points out that this type of insulation does not stand up to the "exhaust steam" test, the silicone varnish being "washed out." This is a good example of the limitations of silicones, which necessitate a closer understanding of their properties if they are to be used with success.

Precision A.C. Voltage Stabilizers

By G. N. PATCHETT, Ph.D., B.Sc.(London), A.M.I.E.E., M.I.R.E.

(Bradford Technical College)

Part 11

Elimination of the Unwanted Voltages

IT is obvious that for accurate work it is essential to eliminate the unwanted voltages produced by a lamp bridge. These unwanted voltages were shown to consist of two parts:—

1. A voltage of supply frequency 90° out of phase with the supply.
2. A voltage of three times the supply frequency.

One method that has been used,[36] although not electronic, is of interest. The output from the bridge is fed to a wattmeter movement (a modified energy meter), the other coil of the movement being connected across the supply. As both components are "wattless" they produce no torque on the movement. This method can only be used where a mechanical movement is required.

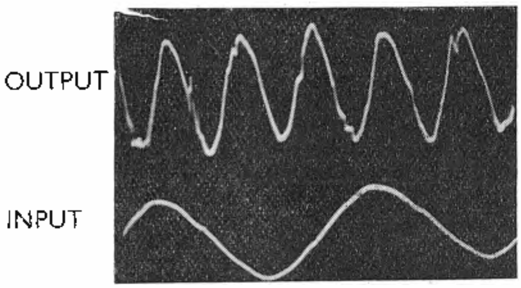


Fig. 5.
Lamp bridge input and compensated output

From the analysis it is seen that the lamp (assuming tungsten filament) behaves as a resistance and capacitance in series (or in parallel) and it should therefore be possible to eliminate the quadrature component by a suitable capacitor in parallel with the resistance arm. This method is quite successful and Fig. 5 shows the output with compensation applied in this way. The value of the capacitor required varies from 0.003 to 0.03 microfarads, depending on the lamp and voltage applied to the bridge. The practical value was found to be somewhat different to that calculated, and this was assumed to be due to the inevitable stray capacitances across the arms of the bridge. For the compensation to hold it must be arranged that the stray capacitances remain constant. This arrangement has the disadvantage that at high frequencies the bridge is a long way from balance, due to the low reactance of the capacitor, and high frequency ripple is likely to occur as shown on Fig. 5. This could be overcome by a capacitor in series with the resistor instead of in parallel, but, unfortunately, the value of capacitance required is too large to be a practical proposition. Some improvement can be obtained by connecting a large capacitor across a section of the resistance arm, say 1/10th of the total. This method of compensation was found to be satisfactory so long as the amplification following the bridge was not too great, i.e., the indicator across the stabilizer input, or, if across the output, the stabilization was not less than about 1 per cent.

Another rather better method from the point of view of high frequency ripple is shown in Fig. 6. The reactance of C' is arranged to be much smaller than R', so that the voltage across C' lags the supply voltage by 90°. By arranging the correct values of C and R', the voltage across C' can be made to neutralize the unwanted quadrature component of the bridge output. This arrangement has been found very satisfactory. None of these methods reduces the figure of of merit of the bridge.

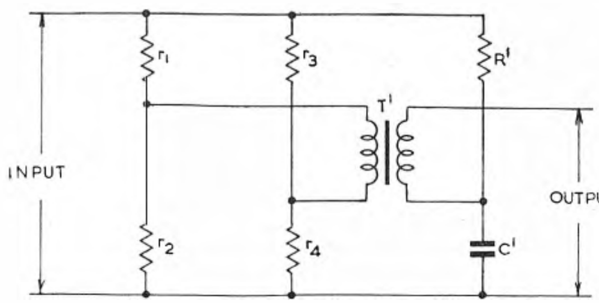
Having compensated for the quadrature component it now remains to eliminate the third harmonic output. The arrangement first used was that of having two similar bridges fed off the supply so that their input voltages were 60° out of phase. This causes the third harmonic output voltages to be 180° out of phase and so cancel out. Although the arrangement was successful it was found to be too complicated and difficult to adjust. It was found easier to eliminate this harmonic output by means of a filter tuned to reject 150 c/s. Efficient L-C filters for this frequency are difficult to make, so a parallel-T R-C filter was used. This is shown in Fig. 7 and has been dealt with by a number of authors.[37,38,39] It can be shown that the filter rejects a frequency of

$$f = \frac{1}{2\pi . C.R}$$

and that at any angular frequency ω the ratio of output voltage to input voltage is

$$|e_0 / e_i| = \frac{1 - (\omega / \omega_0)^2}{\sqrt{1 + 14 (\omega / \omega_0)^2 + (\omega / \omega_0)^4}}$$

Fig. 6.
Circuit for neutralizing the quadrature component of the bridge output



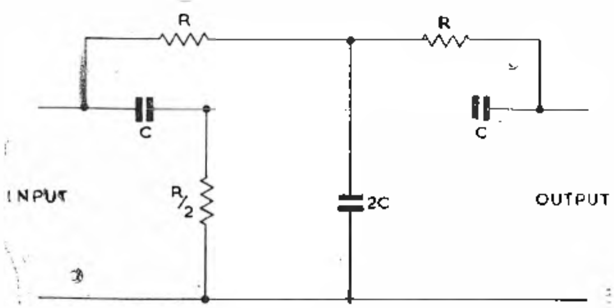


Fig. 7.
Parallel-T filter for removal of third harmonic. Typical values: R = 100k Ω, C = 0.0106 μF

with a phase angle

$$\tan^{-1} \frac{4}{\omega/\omega_0 - \omega_0/\omega}$$

where ω_0 is the angular frequency rejected. In the case considered, ω_0 is made $2\pi 150$ and therefore at 50 c/s.

$$e_0/e_1 = 0.55 \text{ at an angle} -56° \ 18'.$$

This, unfortunately, reduces the overall figure of merit of the bridge, but the loss of wanted voltage can easily be made up in the amplifier if the unwanted components are completely eliminated. A result obtained by this method is shown in Fig. 8(a) where it is seen that the harmonic is completely eliminated. The remaining "hash" is due to rather bad high frequency ripple on the A.C. supply and the use of a capacitance across one resistor arm for quadrature compensation. Fig. 8(b) shows the output for a 1 per cent change in bridge voltage, both records being to the same scale. This method has been used successfully with the second method of quadrature compensation in complete stabilizers, both with the indicator connected across the input and the output terminals of the stabilizer. With the indicator across the output it was not found practicable to get a stability better than about ¼ per cent due to the unwanted harmonic voltages. Although it is theoretically possible to eliminate the quadrature and third harmonic unwanted voltages, when the ratio of unwanted voltages to wanted voltage is high, one finds in practice that the above unwanted components are not true sine waves, due to harmonic in the supply and second order effects in the bridge which have been neglected, and it becomes extremely difficult to eliminate the higher frequencies without causing other troubles. It can be shown theoretically, by an analysis similar to that in the appendix, that the b^{th} harmonic in the supply produces unwanted harmonic components of $b + 2$ and $b - 2$, as well as a voltage of the b^{th} harmonic.

Glynne[16] has used quite an interesting method of eliminating the unwanted components. He uses a double bridge and an L-C circuit tuned to 150 c/s. The operation of the circuit is rather involved and the output is only 40 per cent of that from a single bridge. The initial adjustment appears to be difficult and it would appear only possible to adjust the bridge with a cathode ray oscillograph. A change in one side of the bridge relative to the other also upsets the compensation. The methods given by the author are easily adjusted without the need for an oscillograph, and may be used with only slight modification for the other non-linear devices to be considered.

Effect of Ambient Temperature

The effect of ambient temperature when lamps are operated on A.C. is similar to that on D.C.[21] and negligible if they are operated above about one-tenth of their normal voltage.

Stability

The effect of vibration is, in general, the same as that when the lamps are operated on D.C.,[21] where it is shown that normal high voltage lamps are very unstable when subject to vibration and changes in bridge balance of up to 5 per cent have been obtained. To overcome this trouble special lamps must be used.

The determination of the operating point, design of the resistance arms and the response time are exactly similar to the D.C. case and have already been considered.[21] No compensation for the incremental response time is possible as in D.C., since otherwise a large proportion of the supply voltage would appear in the output.

Summarizing the properties of a lamp bridge we have:—

Disadvantages :—

1. Rather small figure of merit.
2. Large unwanted voltage.
3. Instability due to vibration unless special lamps are used.
4. Large power consumption for many applications.

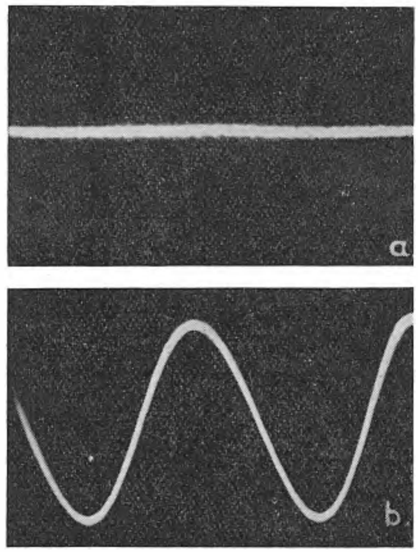


Fig. 8.
(a) Efficiency of the circuit of Fig. 7 in eliminating harmonic ripple
(b) Resultant ouput, to same scale, after a 1 per cent charge in bridge voltage

Advantages :—

1. Effect of ambient temperature may be made negligible.
2. Long life if operated below the normal voltage.

In view of the above it is seen that a lamp bridge may be used across the input to a precision stabilizer, but the stabilizer then suffers from the disadvantages of this position but is not suitable for use across the output of a stabilizer unless the accuracy is low (not generally sufficient for a precision stabilizer).

Barretters

Figure of merit

Over a small range it can be shown that a barretter also follows the law $V = kI^n$, but the value of n is now much higher than a lamp and values up to 40 are possible. The figure of merit therefore approaches unity.

Effect of Ambient Temperature

A barretter characteristic changes considerably with ambient temperature so that both the bridge balance voltage and the figure of merit alter. The latter is important if the bridge is connected across the input to the stabilizer. The magnitude of the change of balance voltage depends on the barretter and the operating point, but is of the order of 0.1 per cent per C°.

Stability

The effect of vibration on barretters appears to be small, but depends on the type of construction. The stability with time also appears to be good; in a particular case a change of only 0.002 per cent per hour taking place. A barretter is very sensitive to draughts and appreciable changes occur due to draughts caused by persons moving in the vicinity. This can largely be overcome by the use of a suitably designed box, but the design is not easy as the power dissipated by the barretters is large. No effect similar to this could be detected with tungsten filament lamps.

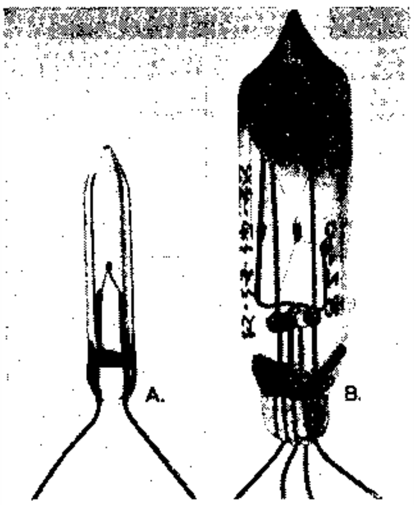


Fig. 9.
Sealed Thermistors. A-directly heated. B-indirectly heated
Courtesy Standard Telephones and Cables, Ltd.

Response Time

It was soon found that barretters are very sluggish in action and take several *minutes* after a change of voltage to settle down to the new equilibrium condition. In a particular case only 90 per cent of the change occurred in 15 seconds and the final value was not reached for 3 to 4 minutes. It was expected that the incremental response time (the time for 63.3 per cent of the change to take place) would be large, but this is not the case. The time of response is of the order of 0.1 second and obviously, if quoted alone, is very misleading. Morton[40] states that there is no appreciable time lag with a barretter, but this appears to be incorrect and it depends on how accurately the results are taken. It is assumed that the initial rapid rise in resistance is due to the comparatively small mass of the filament and the much slower rise in resistance is caused by the gradual rise in the gas temperature, to the new equilibrium temperature, under the new power input conditions.

Magnitude of the Unwanted Voltages

Qualitatively the analysis of the unwanted voltage in a lamp bridge can be applied to a barretter bridge. Experimentally it is found that the magnitude of the unwanted voltage is of the same order as that of a lamp bridge operating at the same voltage. This is presumably due to the fact that the incremental response time is similar, the long time lag having little effect on the rapid cyclical variations in heating.

From the above it is seen that the ratio of the wanted to unwanted voltages is about three times that of a lamp bridge under the same conditions of operation. Similar methods may be used for eliminating these unwanted voltages as with a lamp bridge.

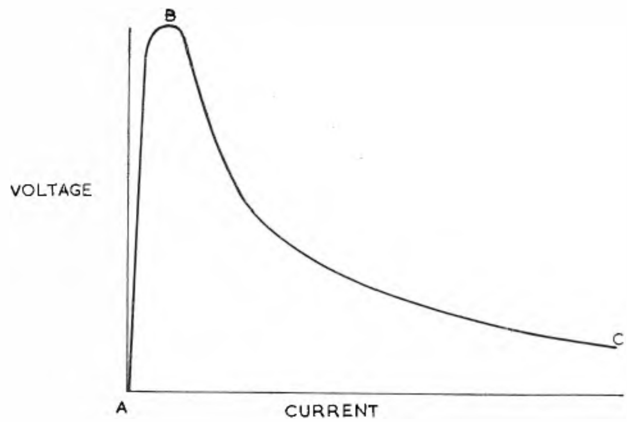


Fig. 10.
Thermistor Characteristic

We may summarize the properties of a barretter bridge as follows:—

Disadvantages

1. Large effects due to ambient temperature.
2. Large changes due to draughts.
3. Long time lag.
4. Power consumption large for most applications.

Advantages :—

1. Larger figure of merit than lamp bridge.
2. Larger ratio of wanted to unwanted voltages than lamp bridge.

In view of the disadvantages it appears that a barretter bridge is not suited for use in a precision stabilizer.

Saturated Diodes

The author has had no experience of bridges using saturated diodes, but this principle is used in America[41] and by other authors.[9,42] It is essential to use a special diode for the purpose and it is interesting to note that one has just been placed on the market in this country.

Thermistors

The type of thermistor used for this purpose consists of a small bead of semi-conducting material having a high negative temperature coefficient, sealed into an evacuated or gasfilled tube.[43,44,45,46] Directly heated and indirectly heated types (heated by a separate heater) are available and are shown in Fig. 9. The response time of the indirectly heated type appears to be too long for use in precision stabilizers and will not be considered further. The great advantage of the thermistor bridge is its large figure of merit; this, as will be shown later, approaches infinity, while the unwanted output voltage is somewhat less than that of a lamp. As the author has found this to be by far the most suitable bridge for a precision stabilizer it will be considered in detail.

Figure of merit

A typical voltage/current characteristic for the type of thermistor used is shown in Fig. 10. As the current is increased the temperature of the bead increases causing the resistance to decrease, this decreasing the rate of rise of voltage over the portion AB. Over the portion BC the decrease in resistance is so great that the voltage actually decreases as the current increases, i.e., the thermistor has a negative slope resistance. Beyond point B the thermistor is unstable on a source of constant voltage and would be destroyed due to overheating. By plotting the characteristics to log. scales as in Fig. 11, it is seen that over the portion BC (corresponding to BC in Fig. 10) the characteristic is approximately straight, showing that the thermistor obeys the law $V = k.I^n$. The value of n is now negative, being about -0.5.

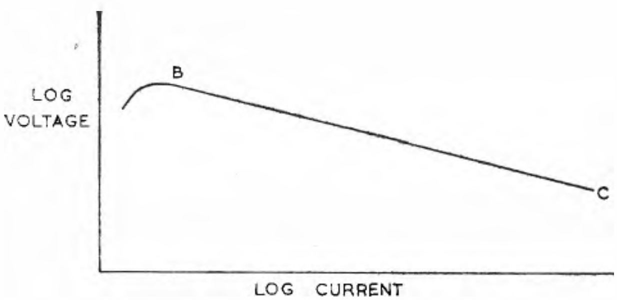


Fig. 11.
Thermistor characteristic plotted to logarithmic scales

For a bridge consisting of three linear arms, r_1, r_2 and r_3, and one non-linear arm, r_4, as shown in Fig. 3, it has been shown[20] that the figure of merit is given by:—

$$\frac{p(n-1)}{(np+1)(p+1)}$$

where $p = r_2/r_1$.

If np is made equal to -1 the figure of merit becomes infinite. This is impossible with devices having a positive index n, but with a thermistor this condition is satisfied by making p approximately 2. This then gives a bridge with an infinite theoretical figure of merit.

This result can be shown in another way from Fig. 12 where the characteristic of the thermistor, r_4, and resistor, r_3, are plotted. The bridge characteristic (or, more correctly, that of arms r_3 and r_4) is obtained by adding together the characteristics of r_3 and r_4 giving the combined characteristic shown. Suppose that the bridge is made to balance at point X (by varying the ratio of r_1 to r_2) corresponding to the horizontal portion of the overall characteristic. Over a small portion around X the characteristic is horizontal and the bridge current, and therefore the bridge output, can change without a change of input voltage. This gives the remarkable result that the bridge output can change without a change in input, i.e., the figure of merit is infinite. In practice the figure of merit is not infinite since the curve is only exactly horizontal at one point, whereas a small portion of the curve must actually be used. In practice a large figure of merit can be obtained and, although difficult to measure, figures around 50 may be obtained.

It will probably be realized that the bridge itself is unstable at a point such as X, but when connected in a suitable stabilizer circuit the whole can be made quite stable and may be operated at any point from,

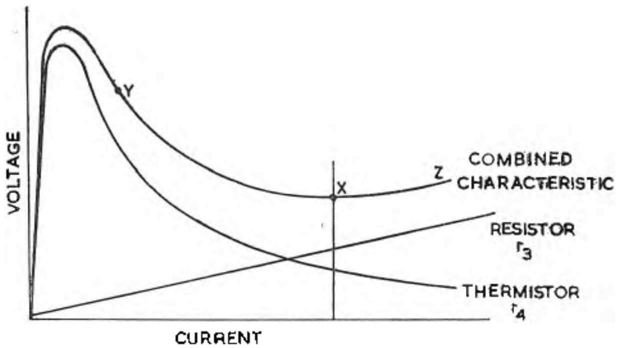


Fig. 12.
Thermistor characteristic, resistor characteristic and combination of the two

say, Y to Z (Z being the limit due to excessive current). If the bridge is operated to the right of X an increase of output voltage, caused by an increase of input voltage, will cause the thermistor current to increase, so increasing the bridge output and tending to correct the output voltage. Since an increase in output voltage is required to increase the bridge current the regulation factor of the stabilizer will be positive and finite. Suppose that the bridge is operated to the left of point X. An increase in input voltage must cause the controller to give a larger output in order to correct for what would normally be a rise in output voltage. In order for this to occur, the thermistor current must increase and this can only be brought about by a *decrease* in bridge voltage. This means that an increase in input voltage results in a decrease in stabilizer output voltage. This result is obtained by a normal type of indicator connected across the input to a stabilizer if the stabilizer amplification M is greater than 1, but would at first sight appear to be an impossibility when the indicator is placed across the output. This property does not appear to be of any practical value and the bridge is normally operated at point X. It is necessary in order to reach the operating point X to get over the peak in the characteristic, and a method of doing this will be detailed later.

Magnitude of Unwanted Voltages

For a 40 volt bridge the value of each unwanted component is about 0.1 volt. For a 0.01 per cent change in input voltage the wanted output voltage would be (assuming a figure of merit of 50) about 0.2 volt and the ratio of wanted to unwanted voltages would be 2. This is an improvement of over 200 times compared with the best lamp bridge. This outstanding improvement in this ratio enables the bridge to be placed across the output terminals of the stabilizer with its corresponding advantages. The bridge is of little use for connecting across the input of a stabilizer as the input-output relationship is not linear and this high figure of merit is not required. The unwanted components may be removed by a similar method to that detailed for lamps, but the problem is now much simpler due to the larger ratio of wanted to unwanted voltages.

Effect of ambient temperature and temperature compensation

Unfortunately, since the thermistor characteristic is brought about by a change in temperature, and its operating temperature is not high, it suffers from changes due to changes in ambient temperature. The amount is large, being about −0.5 per cent per C°, but may be compensated for in a number of ways. The arrangement shown in Fig. 13 has been found most satisfactory. The bridge is supplied from transformer TR_1 through a potential divider formed by r_5 and T_2. T_2 is a large type of thermistor as shown in Fig. 14, consisting of a large block of semiconductor, so that its self heating due to the current is small. When the ambient temperature increases the resistance of T_2 decreases and so decreases the voltage fed to the bridge. This reduction is arranged to be just equal to the reduction in the bridge balance voltage. Thus the bridge is always balanced with the same voltage fed to the primary of TR_1. By this means a change of balance voltage of 5 per cent can be reduced to less than 0.1 per cent for a 10° C. rise in temperature.

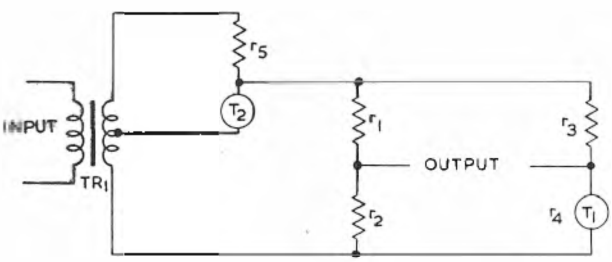


Fig. 13.
Temperature compensated bridge

Response Time

The incremental response time of a thermistor is about 0.3 seconds. This might appear long, but in a complete stabilizer the value is reduced to the order of 0.03 seconds due to the feedback effect when the indicator is placed across the output of the stabilizer.

Advantages :—

1. Extremely high figure of merit.
2. Small magnitude of unwanted voltages.
3. Much larger ratio of wanted to unwanted voltages than any other type of bridge.
4. Electrically and physically small.
5. Not affected by vibration.
6. Due to small size may easily be protected from draughts and rapid changes of temperature.
7. Available in a number of resistance values.

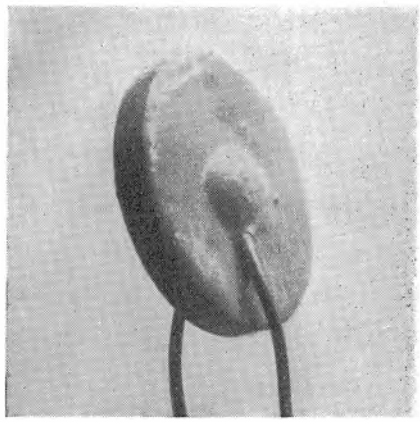


Fig. 14.
Type of Thermistor suitable for use as T2 in the circuit of Fig. 13
Courtesy Standard Telephones & Cables, Ltd.

Disadvantages :—

1. Change of balance voltage with change of ambient temperature, but this may be compensated for to a large extent.
2. Larger incremental response time, but this is not so important when the indicator is placed across the output of the stabilizer, as it is considerably reduced.

From the above it is seen that a thermistor is admirably suited for use in an indicator for a precision stabilizer.

(*To be continued*)

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Microwave Lenses

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Part VI

Conclusions

Production of Shaped Beams

IN many aerial systems, particularly for radar sets, both for use on the ground and airborne, it is desirable to produce a vertical radiation pattern to cover a given volume of sky or area of ground respectively. The usual requirement is that the magnitude of the received power should be the same for all angles of sight for a given height difference between the target and the radar. It follows from the well-known fourth-power radar range equation[32] that the desired vertical pattern is of the form

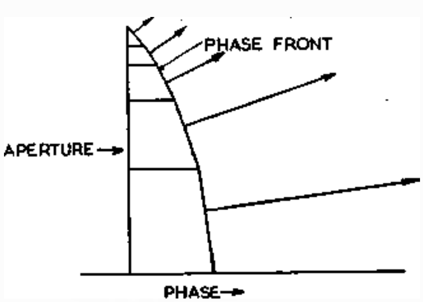
$$P(S) = K \operatorname{cosec}^2 S \qquad (68)$$

where $P(S)$ is the power flux per unit area for constant distance from the radar and for the angle of sight S, and K is a constant depending on the selected constant range and the transmitted power.

Equation (68) implies that the power radiation for zero angle of sight is infinite: this is clearly impossible in practice. It is possible, however, to produce an aerial system which will have a radiation pattern approximating to the required form over a range of angles such as 10° to 80°. Aerial systems of this type have so far usually made use of shaped mirrors or slotted waveguides,[8] but there is every reason to suppose that lenses may be designed to satisfy the requirements equally well.

The essential preliminary to the design of an aerial system which will produce a cosecant squared pattern is the study of the phase and amplitude distribution associated with the pattern. There are two methods of approach due to Chu[33] and Woodward[34] respectively. Chu's method, though by no means rigorous, is probably the more realistic approach to the design of a lens to produce a shaped beam. The basis of Chu's method is to divide the radiating aperture into a large number of elementary horizontal strips and to regard the final radiation pattern as the vector sum of the contributions from each strip. The direction of maximum radiation from each strip will be normal to the phase front existing at the strip. (See Fig. 46.) An approach on these lines will enable the designer, given a known amplitude distribution across the aperture, to derive a suitable phase front to produce the desired beam. A knowledge of this phase front enables a suitable lens to be designed. The general form of lens profiles to produce cosecant squared patterns is shown in Fig. 47.

Fig 46.
Chu's method for obtaining a cosecant squared patterns



Lenses designed to produce cosecant squared beams have not, in general, proved very satisfactory; the reason for the difficulties experienced becomes apparent when the basic assumptions underlying the method of design are considered. It is first assumed that the gain of each elementary strip is proportional to the power flux through it and to its width; the resulting radiation pattern is derived on a basis of ray optics, the direction of each ray being normal to the phase front in the vicinity of the strip producing the ray: the magnitude of the ray depends on the power flux through the strip as shown in Fig. 46.

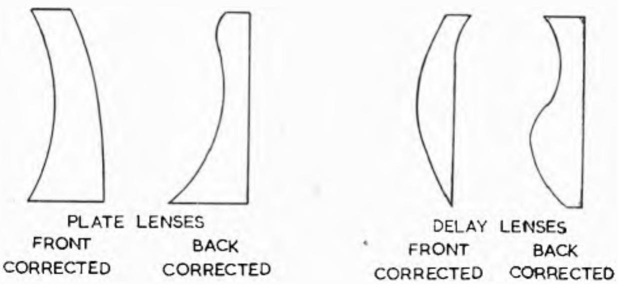


Fig. 47.
Lens profiles for cosecant squared patterns

The above theory is based entirely on the behaviour of each individual ray: in other words, the lens contour has been designed using only the principles of geometrical optics. A more rigorous approach based on physical optics shows that the size of the aperture exerts a considerable influence on the nature of the radiated pattern. In fact, Woodward has shown that if the aperture is ten wavelengths long, then it is only possible to make the radiation pattern have the desired values at ten points. In between these points the pattern will depart from the desired cosecant squared shape as shown in Fig. 48. Periodic variations of this nature occur no matter by what class of beam-forming element, mirror, lens, or slotted waveguide, the cosecant squared pattern is produced. When a lens is used, however, it is found that the departures from the ideal pattern at large angles of sight are worse than when a mirror is used. It has already been explained in Part IV that the reflexions occurring at the surfaces of a lens produce an undesired periodic variation in the amplitude distribu-

* Radar Research and Development Establishment, Ministry of Supply.

tion existing across the outer surface of the lens. The effect of this variation is to produce additional side-lobes at wide-angles. When the lens is designed to produce a cosecant squared pattern, the level of these side-lobes will be comparable with the level of the desired pattern at large angles of sight. Interference effects will therefore cause a deterioration in the performance at these angles. The magnitude of these side-lobes increases as the refractive index departs further from unity. It follows that either the refractive index of the lens must not differ too much from unity, in which case the lens will be undesirably bulky, or steps must be taken to match the lens surfaces. The way in which the radiation pattern of a lens departs from the ideal cosecant squared pattern is illustrated in Fig. 49. A further and unpredictable deterioration may be caused by ground reflexions.

Unfortunately, the practical difficulties of matching lenses have not yet been satisfactorily overcome. The design of a cosecant squared lens must therefore be regarded as a rather unsatisfactory empirical process. However, the manufacture of mirrors to produce shaped beams is a difficult matter, as the surfaces are not surfaces of revolution and therefore cannot be spun; an alternative solution using lens techniques is still desirable.

A satisfactory method of obtaining a cosecant squared pattern, due to R. Shersbie-Harvie, has been described by Lawson.[35] This system uses a small prism to deflect part of the energy radiated by a paraboloid as shown in Fig. 50, and gives very satisfactory results. It has the useful property that the radiated pattern may be adjusted by movements of the prism. In the original model the prism was made of perspex, but the high cost and weight of perspex might now be avoided by constructing the prism from metallic delay material or a metal-plate structure.

The oscillations in the radiation pattern may be reduced in magnitude by increasing the size of the aperture. A further advantage, which is obtained from an increase in the aperture size, is an improvement in the rate of fall-off of the beam below the designed cut-off angle. This causes a reduction in the energy which would otherwise be radiated in unwanted directions. In the case of ground radar systems, improvement in the rate of cut-off below the designed lower limit will also reduce difficulties due to clutter. However, increasing the vertical aperture beyond about twenty wavelengths will not, even though improving the shape of the pattern, produce any great improvement in the range of the radar system. That this is the case may be deduced from a study of the properties of the cosecant squared pattern.

A Comparison of Lenses and Mirrors

In this series of articles an attempt has been made to produce a comprehensive survey of the various types of microwave lens at present in use. It has been shown that lenses perform similar functions to mirrors, and it is therefore desirable at this stage to compare the advantages and disadvantages of lenses and mirrors. Kock, who has been responsible for many of the advances in the development of lenses, states that they are superior to mirrors in three ways: firstly, the tolerances on the dimensions of lenses are not so tight as they are for mirrors, secondly, lenses have superior scanning properties; and, thirdly, the side-lobes in the radiation patterns are smaller for lenses than for mirrors. In this section these three items will be dealt with in turn, and then the ways in which lenses are inferior to mirrors will be considered.

It was shown in Part II that the tolerance on the thickness of a metallic delay lens is $\lambda/16(n-1)$, and the same value applies to all the other types of lens. For the values of refractive index commonly in use, this tolerance is approximately $\lambda/8$ and may be readily met. There is, of course, nothing corresponding to lens thickness in the case of a mirror. On the other hand, the tolerance on a mirror contour is much tighter than on a lens contour, as may be seen by an examination of Fig. 51. If the mirror contour is in error by an amount δx, then the resulting phase error in the aperture distribution is $4\pi\delta x/\lambda$ radians. The same error in the lens contour causes a phase error of less than $\pi\delta x/\lambda$ radians, if the angle θ is less than 30°. It may also be shown that a given magnitude of phase error will have a greater effect on the radiation pattern of a mirror than a lens: this is because the worst phase errors occur on the axis of a mirror but at the edges of a lens. An improvement of at least a factor of eight in the tolerance on the contour may therefore be claimed for a lens. It follows that the supporting framework may be more easily constructed for a lens than for a mirror of the same aperture size. At a wavelength of 3 cm., the tolerances on the contours are approximately 1/32 in. for a mirror, and ¼ in. for a lens.

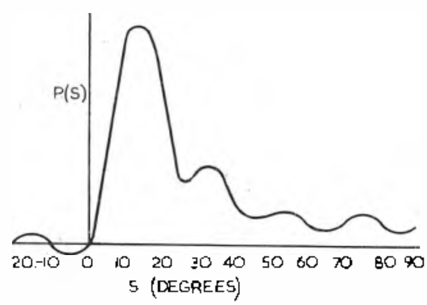


Fig. 48.
Cosecant squared pattern radiated by an aperture ten wavelengths long

This means that it is essential to use a lens when a very narrow beam is desired. For example, a lens of aperture size 480 wavelengths, giving a beam width of six minutes, has been constructed in the U.S.A. The contour of the equivalent mirror would have to be figured to an accuracy of 1 part in 15,000 of its largest dimension: even the differential heating caused by solar radiation would cause errors in excess of this.

The superior scanning properties of lenses are due partly to the fact that the primary feed does not get in the way of the radiated beam, and partly to the increased freedom of design resulting from the lens having two surfaces. The improvements resulting from the second factor were considered in Part IV and will not be discussed further. It was stated in Part IV that the scanning properties of a lens improve as the focal number or numerical aperture (F) is increased. This fact is, of course, well known to students of optics, since the scanning property of a microwave lens is exactly the same thing as the field of view of an optical lens. In order to illuminate satisfactorily the aperture of a system having a large focal number, say $F = 2$, the primary feed must have a radiation pattern whose half-power beam width is some 30°. The radiator, which forms this feed, must therefore have linear dimensions of at least twice the free-space wavelength. If such a radiator were placed in front of a mirror it would intercept an appreciable fraction of the reflected beam. Most of the paraboloids in common use have focal numbers of approximately 0.25, and hence require primary feeds whose dimensions need only be a fraction of the wavelength. The scanning angle of such paraboloids is, however, limited to about 3°. When it is desired to provide a moving beam by moving the feed, the mirror compares even less favourably with the lens, because the scanning mechanism is also in the way of the reflected beam. It is possible to provide some improvement by offsetting the feed system as shown in Fig. 52, but this has an adverse effect on the gain.

The reduction of the size of the side-lobes radiated by a lens occurs for the same reasons as the improvement in the scanning properties. Silver has shown theoretically that one of the effects of the beam obscuration, caused by the feed of a mirror system, is an increase in the side-lobes. When a lens is used it is not difficult to keep the level of the first and largest side-lobe at least 25 db. below that of the main beam. The corresponding level for a parabolic cylinder (a " cheese "), which is the best of the mirror systems in this respect, is between 18 and 22 db. On the other hand, for the reasons given in Part IV, the level of the farther-out side-lobes from a lens is higher than expected, but is still not sufficiently high to cause serious trouble. The side-lobes of a mirror system are increased when steps are taken to remove the mismatch of the primary feed caused by a wave being reflected back into the feed: if a zone plate is inserted in a cheese, for example, the first side-lobe rises to a value only 12 db. below that of the main beam. The methods of overcoming this mismatch of the primary feed in the case of a lens, e.g., stepping one-half of the lens a quarter wavelength with respect to the other, or moving the primary feed off axis (*cf.* Part IV) have practically no effect upon the side-lobe level. It is possible to design a mirror system with a side-lobe level comparable to that of a lens, but the construction becomes very complicated and one of the greatest advantages of a mirror, i.e., simplicity, is lost.

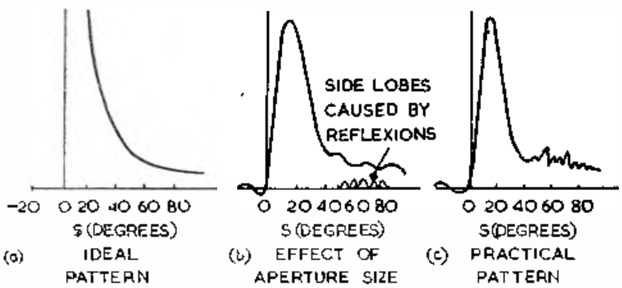


Fig. 49.
Departure of radiation pattern from cosecant squared ideal

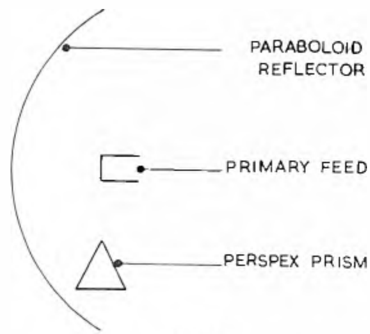


Fig. 50.
Alternative method of obtaining a cosecant squared pattern

A property of aerial systems, related to their side-lobe levels, is the back-to-front ratio, which is effectively a measure of the lobe radiated in the opposite direction to the main beam. This ratio is of great importance in the design of repeater links, in which each station has two aerials placed back-to-back and connected together by an amplifier as in Fig. 53. The fraction of the radiated power from the transmitter aerial, which is fed into the receiver aerial, is the product of the back-to-front ratios for the two aerials. The power gain of the amplifier is limited to a value less than this coupling factor between the aerials, since otherwise the system will become unstable. As a numerical example, suppose both aerials are paraboloids—a typical value of the back-to-front ratio is 100, or 20 db. The coupling factor is therefore 40 db. and so the gain of the amplifier must be less than this. The aerial system with the best back-to-front ratio is the electro-magnetic horn, for which the value is about 50 db. Unfortunately, horns have to be very long in relation to their aperture size in order to obtain the necessary constancy of phase across the aperture, e.g., a horn of 2 ft. aperture, for use at a wavelength in the 10 cm. band, has to be 12 ft. long.[36] The length may be reduced, however, by introducing a correcting lens in the aperture, and this does not appreciably affect the back-to-front ratio. Lens-corrected horns have been used in the U.S.A. as the aerial systems in repeater links, and their performance has been up to expectations.

The principal disadvantage of a lens system is the reduction of gain caused by the reflexion and spill-over losses described in Part IV. Most lenses so far constructed have had gains of between 2 and 3 db. lower than for mirrors of equivalent size. It is hoped to recover part of this loss by the development of suitable matching techniques. Lens-corrected horns are free from reflexion and spill-over losses, and are among the most efficient aerial systems available.

It is not an easy matter to decide whether to use a lens or a mirror as the solution to any given design problem. In general, it may be said that lenses are to be preferred when either side-lobe suppression or wide-angle scanning is the most important requirement. Mirrors are simpler and more compact and should be used when the gain of the aerial is the primary consideration.

A Comparison of the Various Types of Lens

A variety of types of lens is now available, the four main categories being metal-plate or waveguide lenses, metallic delay lenses, rod lenses and path-length lenses. The solid dielectric lens is of limited practical use because of its weight and cost and the difficulty of obtaining large masses of dielectric of the necessary homogeneity. The special systems described in the last article are all designed to meet specific requirements and will not be considered in this comparison.

The metal-plate lens is the earliest and best known type, and as a result there is now available a fairly large amount of practical experience relating to its design. A disadvantage of this type is the narrowness of the band of wavelengths for which a given lens may be used: this disadvantage should not be overrated, however, as, in any practical aerial system, there is almost certain to be some element, such as a matching unit between the oscillator and the waveguide feeder, which will limit the bandwidth at least as much as does the lens. Constructional tolerances may be met, at any rate for wavelengths of 1 cm. and greater. At the smaller wavelengths the number of spacers required to maintain the necessary accuracy of plate spacing may become sufficiently large to have an effect on the radiation pattern of the lens. Such spacers may also increase the reflexion losses from the lens surfaces, and their positioning is therefore a matter for careful study. This aspect of metal-plate lens design is one about which not much is yet known. The egg-box form of construction avoids the deterioration of the radiation pattern arising from the spacers, in effect, by increasing their number to such an extent that the side-lobes produced are not radiated into real space: in other words, the scattered energy is in the form of a storage field rather than real side-lobes. On the other hand, the egg-box construction results in larger reflexions from the surfaces. Egg-box lenses are rather difficult to construct and because of this have not yet been used to any great extent. It is, of course, essential to use an egg-box lens if it is desired to have a metal-plate lens which will operate for any plane of polarization.

Metallic delay lenses are undoubtedly the best of the four types considered here from the electrical point of view. They have wide-band properties, they are very light and they may be easily constructed for the wavelengths at present in use. The strip type in particular behaves almost exactly as predicted by theory, so that the design of any particular lens may be carried out on paper. This form only operates for one plane of polarization; further theoretical work is required to obtain an equally good design procedure for lenses which will work for any plane of polarization. It seems that 3 cm. is about the smallest wavelength for which it will be possible to construct delay lenses: for smaller wavelengths it is difficult to obtain expanded dielectrics sufficiently thin to act as the spacers. Unfortunately, the use of delay lenses in this country is at present limited to aerial systems, in which the gain is of secondary importance. This is because the only expanded dielectrics which are commercially available are too lossy.

Rodded lenses represent an attempt to combine the good electrical performance of the delay lens with the superior mechanical properties of the metal-plate lens. Rodded lenses are similar to metal-plate lenses as far as band-width and polarization are concerned: they are considerably lighter, however, and it is easier to match the surfaces. As yet, there is no experimental evidence to confirm the theoretical improvement in performance over the metal-plate lens.

Path length lenses are also too recent for an accurate assessment of their advantages and disadvantages to have been made. The apparent advantages are the very wide band over which they may be operated, and the very easy tolerances on both the plate spacing and the profile. A major drawback of the inclined plate form is that it produces an asymmetrical amplitude distribution in the direction of the electric vector: this will, of course, cause a corresponding lack of symmetry in the radiation pattern. The serpentine lens has not yet been studied in detail—it appears likely, however, that the plates will have to be spaced so closely together as to make the lens excessively heavy.

The above summary of the advantages and disadvantages of the various types of lens leads to tentative suggestions as to the choice of type which will most satisfactorily meet any given requirement. Almost the only case in which a solid dielectric lens is to be preferred is for the correction of the phase front across the aperture of an E-plane sectoral horn.

The metal plate lens is very suitable for general service requirements such as beaconry, military radar, etc., in which robustness and ease of replacement of damaged elements is of importance. If the rodded lens fulfils the theoretical expectation of better performance it is likely to replace the metal-plate lens in such applications.

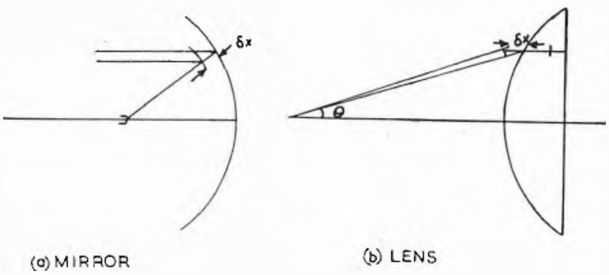


Fig. 51.
Effect of contour errors

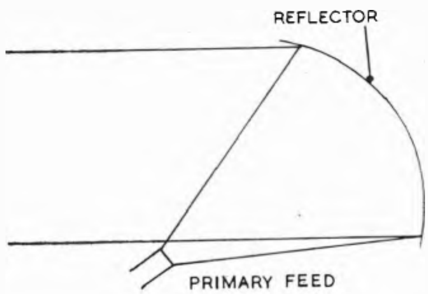


Fig. 52.
Mirror with off-set feed

The great simplicity and ease of construction of the inclined plate path-length lens, together with its broad-band properties, make it an ideal lens for use in microwave repeater links and communication systems. The poor side-lobe suppression in the electric plane may be of little importance if vertical polarization is used, since interference is usually picked up only by the horizontal side-lobes. The serpentine path-length lens is not considered to be developed sufficiently for immediate use.

As has already been stated, delay lenses may at present only be used in aerial systems whose gain is not the most important property. Such is often the case for aerial systems required for laboratory use, and these aerials may be constructed very simply from Dufaylite and metal-sprayed paper. The only tools required in this construction are a pair of scissors and a tin of bakelite varnish. The broad band-width of such lenses is extremely useful in laboratory applications: for example, the equipment used by the authors for measuring radiation patterns employs the same delay lens as the aerial system for investigations at any wavelength between 3 and 12 cm.

Possible Improvements in Lens Systems

There is considerable scope for improvements in the existing types of lens and it is very probable that further new types will be developed. In general, most of the work so far done has been a reproduction of optical techniques in the microwave field. The recent developments, described in the last article, represent, however, a breakaway from conventional optical methods and it is likely that future developments will show even less resemblance to the lenses of optics.

The most pressing requirement, as far as the existing types of lens are concerned, is to achieve some simple method of reducing reflexion losses. Most of the methods of doing this, which were discussed in Part IV, so complicate the design and construction as to make their use impracticable in most cases. Until reflexion losses have been reduced the gain of lens systems will almost certainly be inferior to that of mirrors of the same aperture size. Once a simple method of overcoming these losses has been obtained it will become possible to use refractive indices which are higher in the case of delay materials and lower for metal plate and rodded dielectrics. This will lead to a decrease in the thickness of lenses, with a corresponding reduction in weight.

One of the greatest advantages of lenses over mirrors has been stated to be the much lower side-lobe levels which may be achieved. This improvement in side-lobe level has been obtained without any particular attention being paid to the problem. The fact that lenses have two surfaces leads to a greater freedom of design than in the case of mirrors. Use of this additional freedom may be made in controlling the amplitude distribution across the outer face of the lens. It should therefore be possible to obtain distributions which will result in even smaller side-lobes. A complementary problem is the design of the primary feed: as yet, only feeds such as open-ended wave guides, horns and dipoles, which are in common use elsewhere, have been used. A more careful choice of the feed may lead to further improvements in the amplitude distribution.

An indication of the present trends in microwave lens design has already been given in Part V. Several of the systems described there make use of media in which the refractive index is not homogeneous, a possibility not available in the optical field. These systems have all used some form of spaced conducting sheets and, as yet, focusing in only one plane has been achieved. The introduction of rodded dielectrics offers the possibility of another non-homogeneous medium, obtainable by varying the constants of the lattice structure, the rod diameters, or even both. The last of these suggests the feasibility of constructing non-homogeneous lenses, which will have two-dimensional radiating apertures. An alternative method of obtaining a two-dimensional aperture is to combine two cylindrical systems which focus independently in the two planes. As yet only a few of the ways in which non-homogeneous media may be used, have been investigated: for example, the variation of refractive index, used in the construction of the half fish-eye, is only one of many others which will produce a linear phase-front from a point source. It is probable that some of these

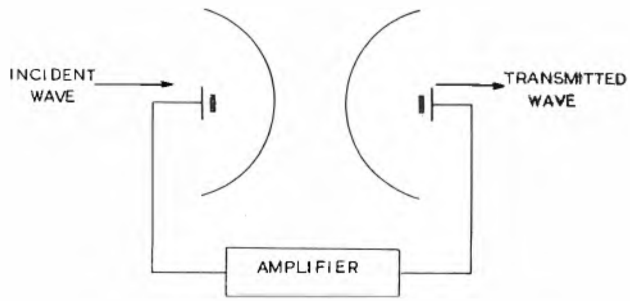


Fig. 53.
A repeater station

may be even more advantageous from the point of view of the properties of the radiation pattern.

Lenses have only been used for wavelengths in the centimetre and millimetre bands. It is certain that the development of even shorter wavelengths will lead to a much more widespread use of lenses. It is also possible that lenses, particularly of the rodded type, may find applications at metre wavelengths. A rodded lens for use at such wavelengths would consist of curtains of wires. The structure would not be dissimilar to the linear arrays at present in widespread use, except, of course, that the number of elements would be greatly increased. On the other hand, there would no longer be the necessity for careful phasing of each individual element as in linear arrays.

The non-homogeneous types such as the half fish-eye could also be extended to the metre band if horizontal polarization were required. The conducting sheets could be replaced by suitably supported areas of wire netting: the spacing between the wires need not be less than about a tenth of the wavelength. Such a system designed to have a gain of 16 db. at a wavelength of 20 metres would occupy a ground space of about 100 yards by 50 yards and, like the rod lens, would be free from phasing difficulties.

Although the subject of microwave lenses has only been studied extensively over the last few years, considerable progress has already been made. There are still many problems to be solved, and it is probable that a rapid rate of progress will be maintained during the next decade.

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Acknowledgment

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The Dot-Sequential Colour Television System

Three-Gun Tri-Colour Cathode-ray Tube

The manner in which the colour screen produces a colour picture is best understood by considering first the operation of the three-gun tri-colour cathode-ray tube. An apertured mask is interposed between the three guns and the dot-phosphor screen in such a manner that the electrons from any one gun can strike only a single colour phosphor no matter which part of the raster is being scanned. The mask comprises a sheet of metal spaced from the phosphor screen and containing 117,000 holes, or one hole for each of the tri-colour dot groups. This hole is so registered with its associated dot group that the difference in the angle of approach of the three oncoming beams determines which colour is excited. Thus, three colour signals applied to the three guns produce independent pictures in the three primary colours, the pictures appearing to the eye to be superimposed because of the close spacing of the very small phosphor dots.

Single-Gun Tri-Colour Cathode-ray Tube

The operation of the single-gun tri-colour cathode-ray tube is analogous to the operation of the three-gun tri-colour cathode-ray tube in that the beam from the single gun is magnetically rotated so that, in effect, it occupies, in time sequence, the three positions of the three guns in the three-gun cathode-ray tube. Thus, when the beam is in a position corresponding to the green gun of the three-gun cathode-ray tube it excites only the green phosphor dots and is at this particular time modulated only by the green component of the video signal. A short time later the beam has been rotated to a position corresponding to the red gun of the three-gun cathode-ray tube and is modulated by the red component of the video signal to excite red phosphor dots. A third position similarly produces the blue picture. Sampling is automatically provided by rotating the beam synchronously at sampling frequency.

Receiver for the Three-Gun Tri-Colour Cathode-ray Tube

Video signal from a conventional black-and-white television receiver is applied simultaneously to the three internally-connected control grids of the three-gun cathode-ray tube. Another signal, derived from the video amplifier, is used to actuate an automatic colour phasing and sampling synchronization circuit which produces a local 3.58-megacycle sampling wave. The latter is applied through an amplifier tube and appropriate delay lines to three gating tubes which supply three sampling pulses, differing in phase by 120 degrees at 3.58 megacycles, to the three cathodes of the cathode-ray tube. Thus, each gun is turned on in time sequence corresponding to the original sampling process at the transmitter and the beam current from each gun excites only one of the three phosphor colours.

A radio-frequency type anode voltage supply provides a potential of 18 kilovolts for the cathode-ray tube final anode, 9 kilovolts for the electrostatic converging electrode and approximately 3.5 kilovolts for the parallel-connected first anodes which produce initial electron-beam focus. A small auxiliary power unit provides heater and H.T. power for the other added circuits.

Receiver for the Single-Gun Tri-colour Cathode-ray Tube

Video signal from the output of the video amplifier of a conventional black-and-white television receiver is applied to the control grid of the single-gun cathode-ray tube in the conventional manner. Here again, as in the previous receiver, another signal from the video amplifier actuates an automatic colour phasing and sampling synchronization circuit which produces a local 3.58 Mc/s. signal which is locked in step with the transmitter sampler. Circular deflexion of the beam, which produces sampling automatically, is provided by a small deflexion yoke having two sets of coils which are fed with quadrature currents at sampling frequency to produce a rotating field. Service adjustment of colour phasing is provided by mechanical positioning of this yoke. The amplitude of the circular deflexion is adjusted to produce the proper convergence angle as required by the mask and phosphor-dot screen. The duration of the sampling period is controlled by a signal having a frequency three times the sampling frequency, which is injected into the C.R.T. cathode circuit. The amplitude and phase of this 10.74 Mc/s. signal are determined by the alignment of a filter circuit which utilizes the third harmonic of the circular-deflexion driver tube.

Potentials of 18 kilovolts for the final anode and 2.7 kilovolts for the electrostatic focus electrode are derived from the kick-back voltage on the horizontal-deflexion output transfomer just as in conventional black-and-white receivers. A small auxiliary power unit provides heater and +B power for the other added circuits.

Convergence of the circularly deflected beam is produced by a magnetic lens in this single-gun cathode-ray tube instead of the electrostatic method employed in the three-gun version.

—Condensed from *R.C.A. Review*, June, 1950.

The Circuit Analogue of Muscle

By D. R. WILKIE, M.R.C.P.

(Department of Physiology, University College, London)

THE analogue circuit described in this paper was devised originally to facilitate some calculations in muscle physiology. However, the ideas it embodies may also be of interest to servo-engineers and others who must include human operators in their circuits; for where the operator must respond by performing mechanical work, the properties of his muscles may be the limiting factor in deciding the performance of the whole system.

The Physiology of Muscle

When a muscle becomes active it tends to shorten and to develop tension. The rate of shortening V depends on the tension exerted P, as shown in Fig. 1.

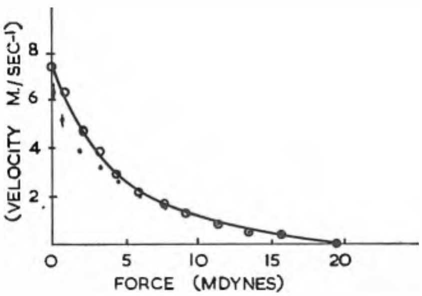


Fig. 1.

The relationship between force and velocity in muscle.
These results were obtained on human muscle.

Abscissa : Velocity of movement of hand, metres/sec.

Ordinate : Force exerted, dynes × 10^6.

Points : Observed results. Each is the mean of 30 observations (six times), whose standard error is drawn as a vertical bar through the point. This is often too short to be visible.

Circles : Results after correction for inertia of the arm and apparatus.

Curves : Drawn from $(P+a)(v+b)=(P_0+a)b$ with $a/P_0=0.20$.

The reduction in force associated with increased speed of movement is part of our everyday muscular experience; however, muscular contraction can be most conveniently studied in isolated muscles, particularly those of small cold-blooded animals such as the frog. Experiments of this type (Hill 1938, Katz 1939) have shown that the force-velocity curve can be fitted by a hyperbolic equation: $(P+a)(V+b)=(P_0+a)b$, where P_0 = tension at zero speed, a and b are constants. The shape of the curve depends on a/P_0 only.

When movements are made voluntarily, the state of affairs is more complex, since in general several muscles act at once through a lever system, and they are not excited to a constant degree throughout the movement. However, if a simple joint is chosen for study, and if movements are made with the greatest possible effort, these difficulties can to some extent be overcome. Studies on maximal flexion of the human elbow (Wilkie 1950) have shown that muscles follow the same hyperbolic relationship within the body as outside it.

The $P:V$ relationship appears to be obeyed at every instant during activity. This makes it possible to predict how the velocity will rise if the muscle is opposed by any specified mechanical system of masses, forces and compliances. Many of the differential equations which arise are non-integrable: the equivalent circuit here described was devised to avoid the labour of numerical integration.

The special case considered here is that of a muscle in the arm pulling against a constant external force F applied at the hand. The mechanical system may be represented by Fig. 2. M is the inertia of the forearm and of any external apparatus; G, the series compliance, is located at least partly in the tendons.

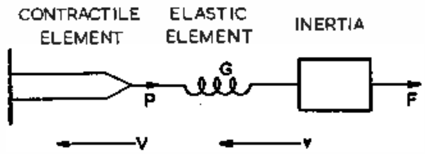


Fig. 2.

Diagram of a muscle pulling against an external load

The equation of motion of M is

$$v+GM.d^2v/dt^2=(P_0+a)b/(F+M.dv/dt+a)-b$$

which is non-integrable. G, M, F, P_0, a and b were known experimentally from other investigations while experimental velocity-time curves of pulls against different forces were available for comparison with the calculated curves.

Analogue Circuit. Theory.

If voltage be taken to represent force, and current to represent velocity, inductance corresponds to inertia and capacitance to compliance.

The curved force/velocity characteristic of muscle can then be imitated by the curved voltage/current characteristic of a non-ohmic resistance, e.g., by a metal rectifier biased to a suitable part of its curve. In Fig. 3, if curves A and B are the same shape, the relationship between voltage E and current I is the same as that between force P and velocity V, in both magnitude and sign, for E and P are positive in the opposite direction from I and V.

The analogue circuit of the whole system is shown in Fig. 4. When the key is closed, very little current

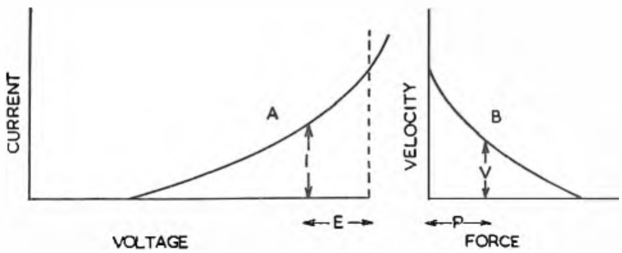
Fig. 3.

Diagram illustrating analogy between muscle and rectifier

Curve A : Voltage-current curve of rectifier.

Curve B : Force-velocity characteristic of muscle.

For significance of symbols see text.



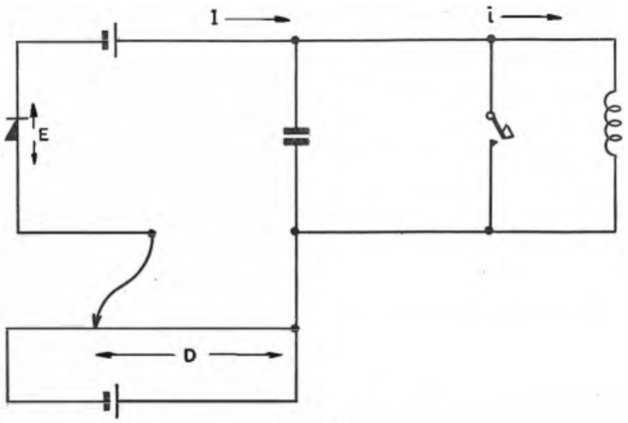


Fig. 4.
The analogue circuit representing Fig. 2.

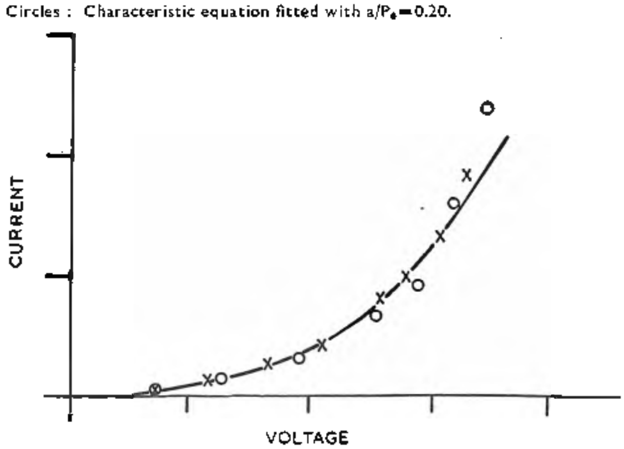
flows through the inductance, but as soon as it is opened the current in the coil rises in exactly the same way as did the velocity of the load in the corresponding mechanical case, for $E=D+L.dI/dt$ (D represents the constant applied force F) and $I = i + C.dE/dt$. Since the rectifier has been chosen so that it obeys the characteristic equation of muscle, $I=(E_0+a)b/(E+a)-b$. Combining these equations to eliminate E and I

$$i+LCd^2i/dt^2=(E_0+a)b/(D+L.di/dt+a)-b$$

which is exactly analogous to the mechanical equation above. In order to solve the equation for known values of D, L, and C it is only necessary to determine practically how the current in such a circuit does rise.

The series compliance has been treated so far as though it were constant: actually it can be shown that the compliance decreases with increasing tension, just as a knitted stocking becomes more difficult to stretch the more it is extended. The variation in compliance is only important at low tensions and no attempt has been made to imitate it in the equivalent circuit. A non-linear compliance could be devised. The capacitance of a capacitor microphone increases with increasing charge, and feedback arrangements could be used to raise its capacitance to a suitable level. However, a better solution would be to alter the system of analogies, making current equivalent to force, voltage to velocity. The analogue of compliance is then inductance, which can more easily be made a function of current.

Fig. 5.
Rectifier curve compared with characteristic equation of muscle
Ordinate : Current in mA.
Abscissa : Voltage applied.
Curve : Rectifier characteristic.
Crosses : Characteristic equation fitted with $a/P_0=0.33$.
Circles : Characteristic equation fitted with $a/P_0=0.20$.



Design

The non-ohmic resistance is the element which dominates the practical design of the analogue circuit. It must have a characteristic of suitable shape and its absolute resistance must be chosen with care, since this governs not only the scaling of mechanical into electrical units, but also the effect on the circuit of stray resistances and capacitances not allowed for in the equations.

The voltage-current curve of the rectifier finally chosen is shown in Fig. 5. The scaling thus defined was:—

10^5 dynes = 78.7 mV   1 metre/sec. = 3.08 mA
1 kg. = 2.55 mH 10^{-4} cm./dyne = 0.39 μF
1 sec. = 1 m.sec.

Stray resistance in the circuit never amounted to more than 2 per cent of the rectifier resistance. No change in the rectifier curve was detected after several months' use.

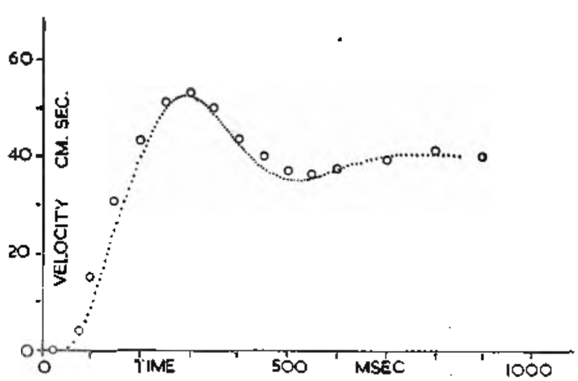
Display Arrangements

In order to adjust D, the steady current through the inductance was measured by a galvanometer shunted by 0.05 ohm to function as a low-resistance milliammeter. Changes of current were displayed on an oscilloscope by amplifying the volt drop across the same shunt with a capacitance-coupled amplifier of time constant about 0.5 sec. When making photographic records the single-stroke time-base was triggered by a Keith Lucas pendulum, which started the analogue circuit running about 0.5 m.sec. later: for visual display the analogue circuit could be actuated repetitively by a relay energized from the 50 c/s. mains. The high writing speed then required was obtained by expansion of the X sweep.

Frequency Response and Linearity of Amplifier

When C is removed and L replaced by a resistor, a square wave can be applied to the amplifier input.

Fig. 6.
Electrical solution (circles) compared with arithmetical solution (dots). Each dot represents one stage of computation.



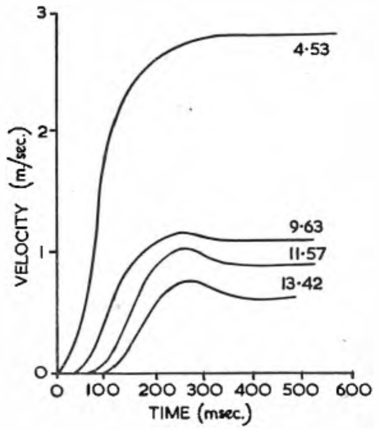


Fig. 7.

Velocity-time curves obtained electrically (a) compared with those found experimentally in muscle (b).

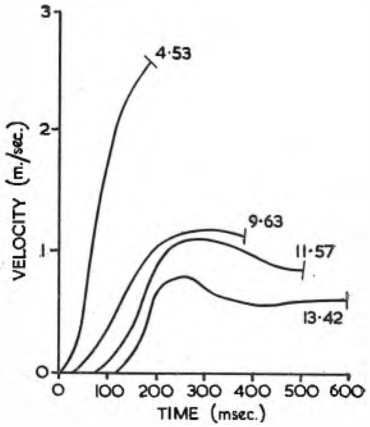
N.B.—The calculated curves have not been fitted to the experimental ones. All the constants required for solution were determined independently.

Ordinate : Velocity of hand in metres/sec.

Abscissa : Time in m.sec.

The number appended to each curve indicates the tension F in dynes x 10^6 against which the movement was made.

The curves have been displaced from their common origin. The bar which ends each experimental curve marks the anatomical limit of the movement. In the faster movements there is not enough time for the maximum velocity to be reached.



The wave recorded rose to 90 per cent of its full height within 40 μsec. The ratio (current change)/(height of photographic record) was constant to within 2 per cent.

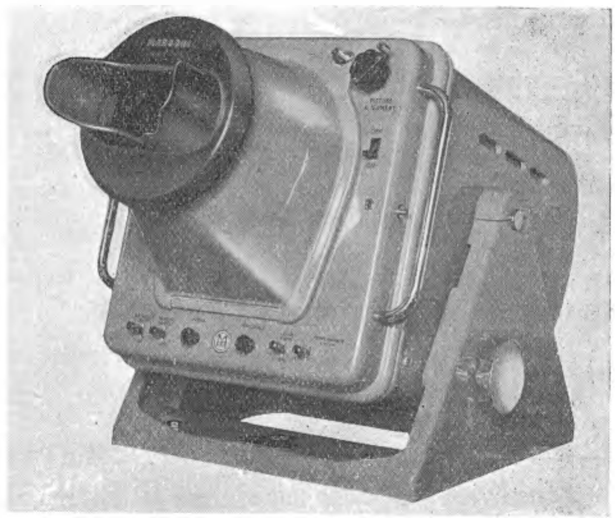
Accuracy

As a check on the accuracy of the calculating circuit an arithmetical solution of the equation was computed in a single case and compared with its electrical counterpart. The close similarity of the two curves may be judged from Fig. 6. The simplified technique of calculation now made it possible to check the original premises on which the equation was based. In **Fig. 7** curves obtained experimentally on a human subject are compared with those calculated from the equation, the constants required for solution having been obtained from independent investigations.

References

Hill, A. V. (1938) *Proc. Roy Soc. B*, *126*, 136.
Katz, B. (1939) *J. Physiol.*, *96*, 45.
Wilkie, D. R. (1950) *J. Physiol.*, *110* 249.

New Marconi Marine Radar



NEW marine radar equipment, the "Radiolocator IV", introduced by The Marconi International Marine Communication Co., Ltd., is capable of detecting and locating suitable targets, such as high land masses, at distances as great as 40 miles, a range so far achieved by no other commercial marine radar installation.

A single switch starts and stops the set, and it is in full operation within one minute of switching on. A gyro-compass repeater may be used for "North-up" presentation of the display, and remote-display units, up to two in number, may be fitted without any alteration to the existing wiring. These remote displays are completely self-contained, and are provided with their own range rings, bearing indicator, etc., for independent observation of targets. The main display unit takes up rather less than 2 cu. ft. of space, and the remote display units are of the same dimensions.

In the "Radiolocator IV", two pulse lengths are used, one for the 1 and 3 mile ranges, and a somewhat longer one for the 10 and 40 mile ranges, resulting in higher definition on the first two and brighter illumination of targets on the latter. The pulse length is automatically changed when switching over from the 3 mile to the 10 mile range or back again.

Built-In Performance Indicator

The aerial scanner of the "Radiolocator IV" installation is fitted with a built-in performance indicator, which, apart from effecting a considerable saving in fitting time and in first installation costs, ensures that all "Radiolocator IV" installations are provided with a standard check on performance, regardless of the structural differences of individual ships. The scanner is also fitted with de-icing apparatus which may be switched on and off as necessary.

Advances in technique have made possible a considerable reduction in the number of valves employed. The "Radiolocator IV" uses only half as many valves as its predecessor, a fact which means a substantial increase in reliability, since the majority of radar breakdowns at sea are traceable to valve failures.

Price £1,950, with all requisite test equipment, and with a full stock of necessary expendable spares.

45 Mc/s. I.F. Strip Design

THE designers of airborne radio and radar equipment are continually being pressed to reduce the size and weight of their equipment. The progress in the design of I.F. amplifiers for radar systems in the past few years is an interesting example of the manner in which valve and component developments are reflected in the improved performance and reduced size of electronic equipment.

The photograph shows a war-time I.F. unit which, together with its pre-amplifier, was extensively used in R.A.F. equipment, in comparison with examples of modern units based on miniature and sub-miniature valves. The reduction in size is quite striking, and one immediate advantage arising from this is that the whole I.F. unit can be mounted in close proximity to the R.F. mixer, thus disposing of the separate head amplifier which is otherwise required.

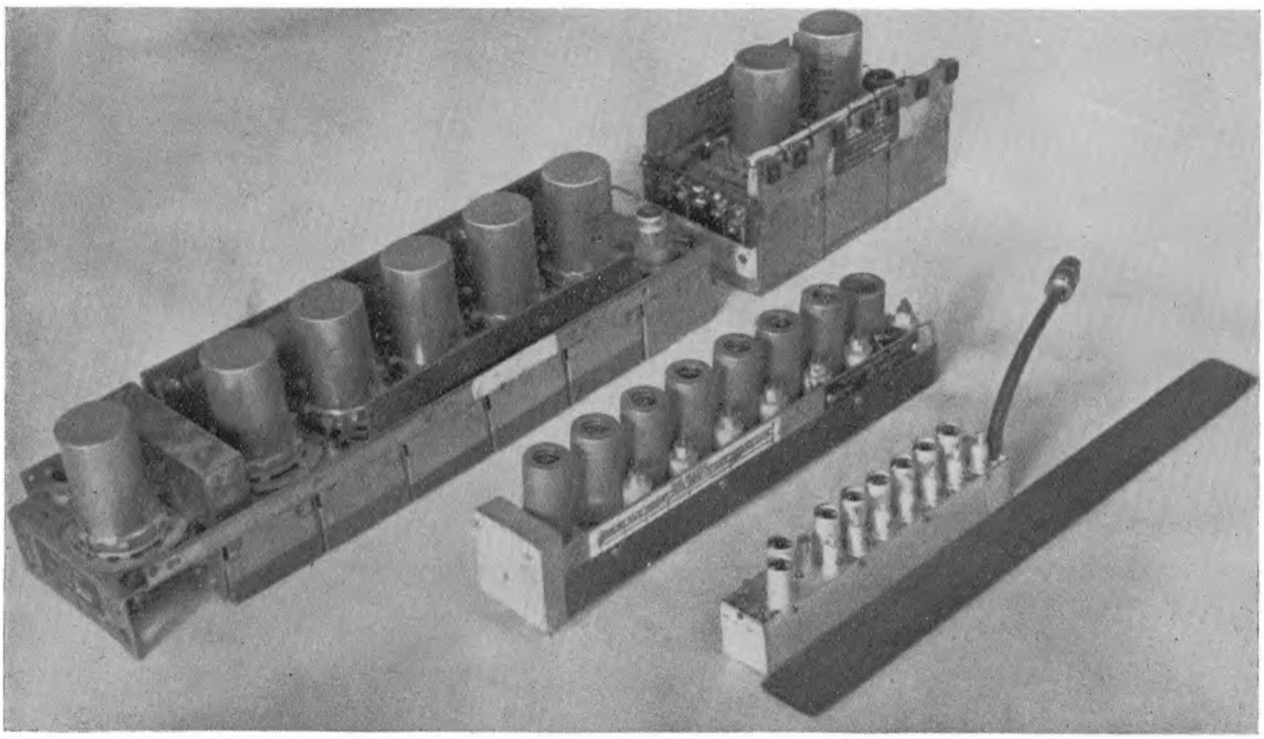
The miniature and sub-miniature amplifiers illustrated are basically similar in circuit arrangement and mechanical layout although there are various detailed differences.

The first stage in each case consists of a suitable R.F. pentode strapped as a triode and neutralized. The neutralizing capacitor for this application is a specially developed miniature component and is roughly one quarter the size of a normal air-spaced trimmer. The triode stage is followed by a normal pentode. Under these conditions it is found that the second stage contributes about 10 per cent of the total noise output from the system and a further small improvement in the noise factor could, therefore, be achieved by use of the cascade or a similar circuit. At the frequencies in question the small decrease in noise factor obtained is not worth the additional complication. Noise factor measurements on various valves have shown that the 6AK5 strapped as a triode is the best type of valve in current use, although the 6J6 with both halves in parallel is only about 0.5 to 0.8db worse.

Considerable thought has been given to the inter-stage coupling arrangement. One important consideration is that the lining-up procedure both in production and subsequent servicing should be as simple as possible. A satisfactory gain-frequency characteristic for both bandwidths is secured by the use of dust-iron cored circuits, stagger-tuned in pairs. In the miniature strip the tuned circuits are in the grid of each stage, the required damping being provided by the anode resistor of the previous stage. In the sub-miniature strip the principle of staggered pairs is retained but the single tuned circuits are replaced by " unity " coupled transformers.

A very high degree of stability is obtained in the miniature and sub-miniature units although there is no interstage shielding. By far the biggest factor in stability problems is the reduction of earth currents to the absolute minimum by careful layout. It is interesting to note that in spite of reduced length, the sub-miniature strip gave less design trouble from the stability point of view than the miniature version. This is to be expected if the theory which regards the chassis plus its cover as a wave-guide below cut-off is valid, since the decrease in overall length is accompanied by a reduction in cross-sectional dimensions and a consequent substantial increase in the cut-off frequency of the box.

(Communicated by E. K. Cole, Ltd.)



Electronics at the British Association

(From Our Special Correspondent)

ALTHOUGH there were many incidental references to the science of electronics at the annual meeting of the British Association for the Advancement of Science, which ended in Birmingham on September 6, the subject properly claimed only two papers; both of which were well received.

Professor A. Tustin (Department of Electrical Engineering, Birmingham University) devoted his attention to the progress that has been made in automatic control systems. The principles and limitations of all automatic control systems were analysed; and he showed that such systems might be regarded as power amplifiers with feed-back, and that the fundamental limitations to accuracy and speed of control arose from energy storages in the system. Mr. L. Bainbridge-Bell (Admiralty Signals and Radar Establishment, Witley, Surrey) spoke on the need for clearer circuit diagrams, and some principles and articles were detailed.

Present Progress in Automatic Control Systems

The development of automatic machinery and of mechanization, said Professor Tustin, was continuing more rapidly than ever before. Most people believed that to be a good thing and to be likely, in some not-too-distant future, to ensure a good life for mankind. "I am afraid that I cannot regard this result as certain. It seems equally possible that advancing mechanization may lead to the dissipation of the earth's limited resources in a joyless urban existence, to a blind proliferation of population, to intensified economic competition and to ever greater horrors in war."

When one spoke of automatic operation, said Professor Tustin in referring to electronic controls, it was usually the automatic carrying out of a predetermined sequence that was meant; and mechanisms of that sort continued to be developed of constantly increasing complexity. He exampled the automatic telephone exchange and a modern digital calculating machine, which essentially performed a pre-arranged sequence of counting operations corresponding to an arithmetical calculation. But during and since the war progress had been particularly marked in the distinct class of error-actuated automatic devices, and special consideration was given to that category of control. He explained that it differed from the fixed-operation type in that the operation was not predetermined; it depended on circumstances. The mechanism operated so as to bring about a predetermined *result*, independently of varying circumstances and disturbances. Because of the dependence of the operation on the result, which was contrived by causing any error from the *required* result to operate the control in such a way so as to minimize the error, such control systems were called "error-actuated" or "monitored" controls. They were also "feed-back controls," since they necessarily included feed-back of information as to the magnitude of the error.

It might at first be thought, said Professor Tustin, that automatic control systems could be made indefinitely more and more accurate by simply increasing the amplification, so that a smaller error would suffice to provide the required variations of output. That would be the case if the responses to error were instantaneous, but in practice time delays occurred and the corrective action persisted in the same direction after the error had been fully corrected; thus the error was reversed and appeared in the opposite direction. Time delays were shown to be the enemy of accurate control; if they existed it was impossible to raise the amplification indefinitely without cumulative oscillation occurring. It was because all electric circuits were inductive that a power amplifier, such as an electric generator or a sequence of electric generators, could not be made to have an instantaneous response.

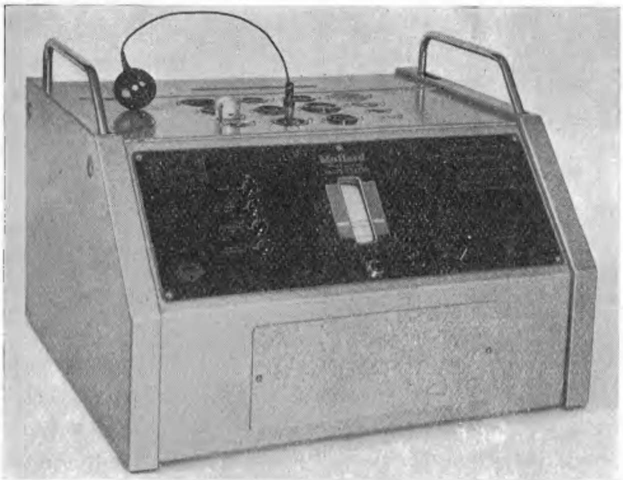
Simplified Schematics

The need for clearer circuit diagrams was emphasized by Mr. L. Bainbridge-Bell. Earlier users of electronic circuits had little difficulty as the apparatus had such a well-defined path of flow from cause to effect that it was difficult to draw a confusing diagram; but the development of radar with its various time-measuring and pulse-shaping circuits, often involving multiple feed-back paths, had made the situation intolerable. The advent of those complicated circuits, in which it was difficult to find where a particular circuit was thought to start, coincided with the need for quick (war-time) training of operators and servicing engineers. Those circuits had set the pattern for much peace-time electronic apparatus, so that the problems of clarifying complicated diagrams was present in ever increasing degree. Some guidance already existed, and in B.S. 530 there was a code of practice. Mr. Bainbridge-Bell mentioned a few artifices which he had found useful. The human eye would easily follow a line formed of smooth curves across a number of other patterns, particularly if those patterns were rectangular. The quantitative study of the time saved (or the accuracy gained) by the use, in moderation, of curved lines would be interesting and, he thought, a profitable study for the psychologist. A study of the results might persuade more people to break away from the mechanical draughtsman's conventions. Arrows, with their barbs drawn on the circuit lines, were useful, even when the "cause to effect" bow went from left to right. They livened-up the diagram and could emphasize important paths.

The solution was (1) that readers should develop a critical faculty and voice their discontent when confusing diagrams were presented to them; (2) that in universities, students should be given guidance on the subject, and lecturers should draw their diagrams carefully. Mr. Bainbridge-Bell suggested that in commercial firms a man with the requisite technical knowledge should act as adviser to the drawing office. He exhibited a "classical example" of utter confusion which sought to reveal the circuit of a motor car's electrical equipment.

Electronic

A selection of new electronic equipment and accessories exhibited at



Mullard High-speed Valve Tester

(Illustrated above)

HIGH-SPEED, accurate valve testing by non-technical personnel is now possible by means of a new, automatic valve tester introduced by Mullard Electronic Products, Ltd., to be known as the Mullard High-speed Valve Tester.

The new instrument has been specifically designed to satisfy all the exacting requirements necessary for the accurate testing of valves in dealer's shops and service departments in accordance with approved test specifications.

An interesting feature of the new instrument is the provision of a cathode ray tube and a coloured scale for visually indicating the condition of the valve under test. If the valve is within the specified limits, the spot on the screen is deflected to the green part of the scale. If, however, the valve fails to meet the test requirements, the spot is deflected to the red part of the scale. A reading in the intermediate yellow portion of the scale indicates that the valve, although still giving reasonable performance, will probably need early replacement.

Correct testing conditions are applied automatically. The selection of the correct test voltages is quite simply performed by inserting in a multiple gate switch, a punched card which corresponds to the type of valve under test. Numbered cards are included to cover the popular radio and television valves of all makes, and other cards are available on request.

It is a compact and solidly constructed instrument, with all internal components readily accessible for servicing by removing one or more sections of the case.

The top panel carries a complete range of valve holders to accommodate all current types of valve. A flying lead to top-cap connexions is also provided.

The tests include the following: filament or heater continuity, electrode insulation with and without H.T. applied, heater-cathode insulation, grid current, emission and electrode open circuit. These various tests are applied by simple operation of a rotary switch and four push buttons.

The valve tester has been designed to operate from a 50 c/s, 180-260-volt A.C. supply.

The instrument is less than 12 inches high and requires a bench space of not more than 16 inches square, and is easily transportable.

Mullard Electronic Products, Ltd.,
Century House,
Shaftesbury Avenue, W.C.2.

Portable Radioactive Ore Detector

THE Ekco "Prod" is an audible type portable counter which has been reduced to a simple form and comprises: a Geiger Muller tube, a polarizing battery, special headphone matching transformer, a small resistor and a pair of headphones.

The amplitude of the clicks is adequate, even under extreme conditions, as the pulses are shaped and amplified by the specially designed output stage incorporated in the "prod."

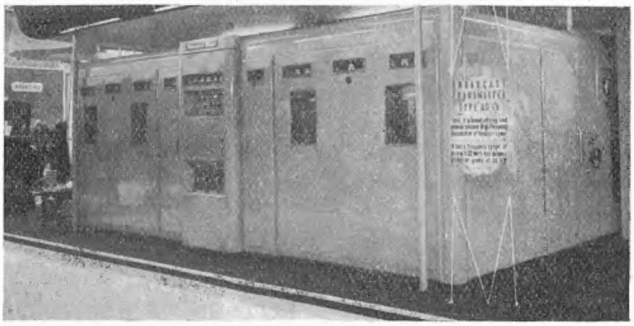
There are no controls, no interconnecting cables, except for the headphones, and no electronic circuitry to go wrong. It weighs only 3¼ lb. and may be carried in one hand like a wand, and used as a probe, or slung from the body. Since the instrument does not require to be switched on it is ready for use at all times, the only replacement required being the battery, the life of which depends only upon the climate. In temperate climates the life is between six and nine months.

The life of the Ekco halogen-quenched Geiger Muller tube is not affected by use. Another advantage of this tube is that the polarizing voltage is reduced from over 1,000 volts to less than 400 volts. This, apart from the reduction in size and weight, means a considerable saving in the cost of replacement batteries.

E. K. Cole Ltd.,
Ekco Works,
Southend-on-Sea,
Essex.

Standard 20 kW Short-Wave Equipment

THE "Standard" Type D.S.15 Radio Transmitter is a 20 kW short-wave equipment, which may be employed in conjunction with the appropriate external Drive Unit for



Equipment

The National Radio Exhibition, Castle Bromwich, from September 6—16.

a number of services, such as :— broadcasting and telephone transmissions (A.3) ; high-speed telegraph transmissions with either frequency-shift (F.1) or "ON/OFF" Keying (A.1). (In the latter case, facilities exist for the carrier signal to be either steady for straight CW telegraphy, or "wobbled" about the mean value to correspond to MCW telegraphy) ; reduced carrier single or independent sideband transmission (A.3a or A.3b) ; and the transmitter may also be used for frequency modulated "facsimile" transmissions (F.4).

Operating frequencies are controlled by accurately ground quartz crystals mounted in a thermostatically controlled oven. The R.F. exciter mounts up to six crystals. Alternatively, the oscillator may be self excited.

The final R.F. stage employs an inverted or grounded grid amplifier.

Also shown on this stand was a diagrammatic map of the new radio link explaining the part played by the Company in its execution.

Standard Telephones & Cables Ltd., Connaught House, Aldwych, London, W.C.2.

D.C. Insulation Tester

(Illustrated top right)

THIS portable 10 kV Insulation Tester was designed to utilize an electronic technique.

Audible indication of the degree of ionization in the test specimen is obtained by amplifying the A.C. components of the ionization current passing through the test specimen and applying them to a small loudspeaker. When testing a batch of components, a faulty specimen is readily detected by an increase in ionization noise compared with the normal ionization level obtained from good samples. The use of the ionization amplifier is restricted to those specimens that are isolated from earth.

It should be noted that the examination of ionization phenomena has only recently been introduced as an additional technique for insulation testing ; so far, it has not yielded sufficient data to enable it to be relied on for quantitative insulation measurement.

The test voltage is a stepless adjustment from 500 to 10,000 volts D.C. The resistance ranges, as calculated from current readings at 10 kV, are in three positions of 5,000 to 250,000 megohms, 500 to 25,000 megohms and 50 to 2,500 megohms. The accuracy is within 10 per cent up to 2,500 megohms and within 20 per cent up to 100,000 megohms. Indication of insulation is given by a $2\frac{1}{2}$ in. meter—calibrated to read applied volts in kV, and current through specimen in micro-amps.

The English Electric Co. Ltd., Industrial Electronics Department, Stafford.

Appliance Switches

FOUR new types of appliance switches have been added to the list of Bulgin components.

The List No. S.377 is a two-pole on/off lever switch having all-insulated moulded thermo-setting structure. It is for 6 amps at 250 volts peak maximum, fitted with 6 B.A. screw terminals for cables up to 3/20 S.W.G. or equivalent. It has dry insulation resistance of not less than 400 $M\Omega$ at 500V., D.C.

List types S.560, S.561 and S.360 are similar in construction. They are very robust, having a push-push action single-pole on/off for A.C./D.C. appliances. They are for 250 V 2A peak maximum, tested at 1 kV peak. The first is fitted with an erinoid knob, the second with a polished nickel-plated knob and the third with a rubber foot knob.

A. F. Bulgin & Co. Ltd., By-Pass Road, Barking, Essex.

R.F. Heating Equipment for Cabinets

(Illustrated below)

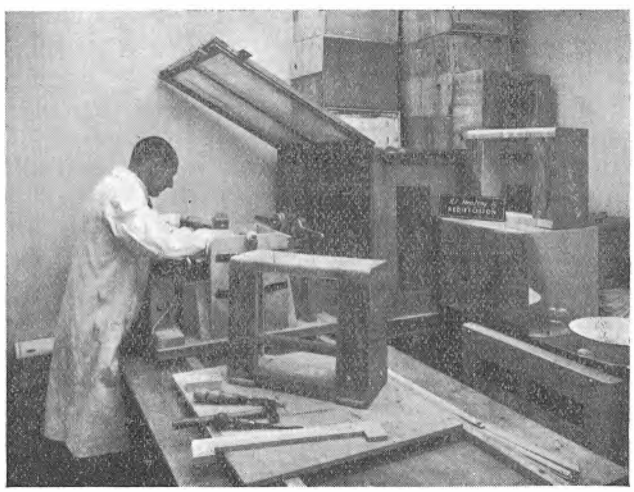
J. B. Manufacturing Cabinets Ltd. use an R.F. heating equipment for making cabinets. The wood parts are covered with a synthetic resin glue at the correct joinings and put into a wooden jig, which is in turn put into a screen cage.

The screen cage uses a Redifon R.F. heater of 7 kilowatts, and has a time cycle of 3 minutes. The main advantages are that shapes are fabricated from lamina and curves are made without any strain or buckling.

The illustration shows the operator setting the jig, with a completed cabinet in the foreground.

The glue used is exceptionally strong, and cabinets can be bounced on the floor when set without breaking.

J. B. Manufacturing Co. (Cabinets) Ltd., 86, East Palmerston Road, London, E.17.



LETTERS TO THE EDITOR

(We do not hold ourselves responsible for the opinions of our correspondents)

A Photographic Timer

Dear Sir,—Mr. C. H. Banthorpe's circuit (March issue) is interesting but unduly expensive considering the accuracy is still limited to the accuracy of the resistors used in the "C.R." circircuit.

A much cheaper device using neons will give the same (if not greater) accuracy, no valves or transformer being required.

I have recently built two such timers, one for A.C. and one for D.C. operation, which are very reliable. The circuit may be of interest. (See below.)

Notes:

Dissimilar neons are necessary.

Neon "X" stabilizes charging voltage. The one used stabilized at 165V.

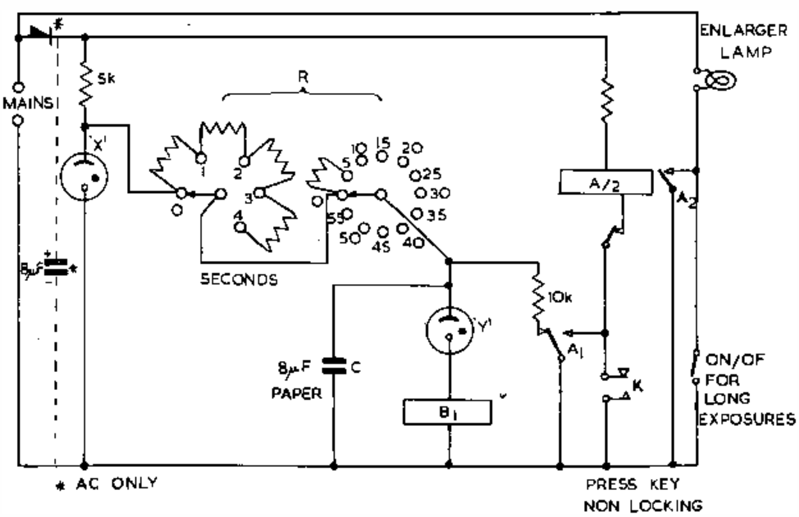
Neon "Y" has a striking voltage less than the stabilized voltage, i.e., 120V—145V, cv188 is suitable.

Operation:

1. On pressing *K* relay *A* operates via *B*1 and
2. Locks via *A*1 also shunt 10k removed from *C*
3. Lamp circuit completed via *A*2
4. Capacitor *C* (8μF paper of good insulation) charges via selected resistance *R*
5. Neon "*Y*" strikes and relay *B* operates
6. Circuit of relay *A* breaks at *B*1 releasing relay *A* and switching off lamp.

Choice of values for *C* and *R* depend on range required. Suitable values are *C* = 8μF paper good insulation, for 1 second approximately *R* = 90k, for 5 seconds approximately *R* = 490k.

Yours faithfully,
R. F. Wolfenden.
Hendon,
London, N.W.4.



Mr. Banthorpe replies:

I like Mr. Wolfenden's timer, and he has obviously given it considerable thought.

I have had some experience of a similar circuit, and neons are an obvious choice if extreme economy is necessary.

Unfortunately neons vary rather a lot, and unless one can select them from several, the resistors may have to be different in different units, particularly as the striking potential of neon "Y" is not much below neon "X" and it will thus strike on the flat part of the charging voltage across *C*.

The resistors would have to be changed if either neon was replaced by another with a different striking potential, and unless relay *B*1 is of high resistance, neon "Y" may have a short life.

Mr. Wolfenden mentions a striking potential of between 120V-145V for neon "*Y*," but these values would make a time difference of over 1.6 : 1.

The position may be improved somewhat by tapping up the 5k resistor in series with neon "*X*." The hard valve circuit is, of course, much more repeatable.

I have found that a timer similar to mine has been described before[1], in rather more detail, and the writers found theirs satisfactory too. There are probably a great number of satisfactory timers in use, and they are a real help in the darkroom.

I think Mr. Wolfenden's circuit may be of considerable interest to those who require a timer of extreme economy, and who would be prepared to "tailor" the resistors to the particular neons used, and would not mind changing them if they had to replace a neon.

[1] "A Timer for Photo Printing" by N. Phelp and F. Tappenden. *Electronic Engineering*, October 1946

Time-Calibrated C.R.T. Traces

Dear Sir,—It was with interest that I read Mr. O. S. Puckle's adaptation of Mr. C. H. Banthorpe's time marker circuit for television receiver spot wobbling, in the June issue of Electronic Engineering. There are, however, two points in the deflector plate theory which appear misleading:—

(i) A plate spacing of ¾ in. has been assumed, but this is really an arbitrary figure and cannot be realized in present-day television tubes, whose neck diameters are some 35 mm. Calculations suggest that Mr. Puckle's figure of ¾ in. is, anyway, too small when taken in conjunction with the plate length and sensitivity quoted; if the plate spacing is increased to the 35 mm. required by current tubes, the necessary deflecting fields would, however, still be obtained.

(ii) It is difficult to see how the electron beam, at least in commercial television tubes, can be suitably deflected by externally applied electrostatic fields, since the internal wall of the neck, which is coated with graphite and connected to the final anode, has a fairly low resistance. Thus it is impossible to introduce deflecting fields within the tube neck by placing deflector plates outside.—Yours faithfully,

L. S. Allard.
Research Laboratories,
The General Electric Co., Ltd.,
Middlesex.

The Influence of High-Order Products in Non-Linear Distortion

Dear Sir,—As an originator of the 1937 R.M.A. specification for testing harmonic distortion in broadcast receivers, I was very interested to note Mr. Shorter's article on this subject in the April issue.

A series of tests upon the minimum per cent harmonic audible on pure tones gave figures varying from some 5 to 10 per cent for 2nd harmonic to the order of 0.1 per cent for, say, the 9th harmonic. The figures obtained varied considerably with the fundamental frequency and the loudness level, and were checked against the curves for masking of a high tone by a lower tone, as given by Harvey Fletcher. I have not got the actual figures by me now, but can say that they indicated that while no simple xn or xn^2 weighting system was really correct, either of these would be a great improvement over the usual R.M.S. sum.

The R.M.A. specification test was proposed only after a simple piece of test apparatus working on these lines had been in use over a considerable period. This merely comprised a filter for the fundamental plus a single *R-C* section to attenuate the lower harmonics proportionally, together with an amplifier and output meter.

Although in some cases a more drastic weighting was indicated by the tests, there were certain practical difficulties (e.g., correct weighting varies greatly with fundamental frequency) and it was felt that it would be unwise to standardize on a system so radically different from current practice without exhaustive tests and considerable practical experience of its correlation with audible distortion on actual programmes.

More recently I have made attempts to correlate intermodulation distortion with per cent harmonic. If we assume a power series ($aV+bV^2+cV^3+$ ... etc.) characteristic, we can easily calculate all the harmonic and intermodulation tones present for any given values of the coefficients a, b, etc., provided we limit the number of terms in the series to, say, eight, or assume a strong convergence in the coefficients. However, tests on actual valves under various Class A and Class AB conditions gave a per cent of harmonic relative to that of corresponding intermodulation tones which did not agree well with the calculations: this can only mean that the valve characteristic is not at all accurately representable by a power series with strongly convergent coefficients.

Insofar, however, as the analysis is valid, it does suggest that the frequently advocated measurement of the first intermodulation tone will not tell us any more (i.e., is not any more indicative of the presence of high order curvature in the characteristic) than a measurement of 2nd harmonic, and it certainly cannot tell us anything about the odd-order distortion terms which may be present.

I conclude, therefore, that the R.M.A. test could well be more widely used, and that further research should be undertaken into the relative merits of harmonic measurements with heavier weighting, and of intermodulation measurements.

M. V. Callendar.
Southend-on-Sea, Essex.

Mr. Shorter replies:

I have studied the letter from Mr. M. V. Callendar relating to my article published in the April issue of Electronic Engineering, and would like to make the following comments:—

I share Mr. M. V. Callendar's views on the need for caution in the introduction of new test methods. The purpose of my article was not to propose any one system of distortion measurement, but to demonstrate one of the shortcomings of the current practice and to indicate possible lines of approach for the future.

It is a pity that the 1937 R.M.A. specification for distortion testing has received so little attention. If the method there prescribed had been generally adopted at the time, much valuable experience would have been gained and we might now be in a better position to consider further refinements in technique.

His observations on intermodulation tests are of great interest. The value of such tests depends very much on their being intelligently applied. An incomplete analysis of intermodulation products is not only subject to the same kind of limitation as an incomplete analysis of harmonics, but, as Mr. Callendar points out, can be misleading in the matter of odd-order distortions.

I entirely agree with the suggestions contained in Mr. Callendar's last paragraph, to which I would add (a) that in the present state of the art, the amplitudes of individual distortion products should be quoted whenever the degree of linearity of different systems is being compared, (b) that any distortion measurements should be supplemented by an oscillographic inspection of the waveform for discontinuities.

Checking Television Interlace

Dear Sir,—On the occasion of the recent Radio Show at Castle Bromwich the Television Society organized a visit of members and friends to the Sutton Coldfield transmitting station, thus taking a rare opportunity of bringing together members from different centres.

I was lucky enough to be included in the party, and, apart from the hospitality extended to us by the B.B.C. staff, the main impression I carried away was of the perfectly interlaced pictures on the control room monitor screens. From these "ideal" pictures it was obvious that the quality of interlacing can be quite easily checked by eye.

By staring fixedly at one small area of the cathode ray tube face the general pattern of the lines will be observed; this will reveal if there is any *slight* tendency to pairing. If the lines are equidistant, and single lines can be followed by the eye as they appear to move up the tube face at a rate of about 1.25 inches a second then perfect interlacing has been achieved. It will also be noticed that the moving lines seem to be separated by twice their usual spacing. — Yours faithfully,

E. Turner.
London, N.13.

BOOK

Electronic Valves—Book I. Fundamentals of Radio-Valve Technique

By J. Deketh, translated by F. G. Garratt. 535 pages. Philips Technical Library, distributed by Cleaver Hume Press, Ltd., London. 1949. Price 25s.

THE radio-valve industry has an extremely good record in respect of the effort which is applied to research and development. Consequently it is in continuous production of development engineers who wish to understand the fundamental principles of radio-valves. In this book the author has done a very great deal to meet this demand. He starts by explaining very clearly the factors controlling electron currents in valves, and shows great skill in avoiding unnecessary mathematics, demonstrating the three-halves-power law, for example, quite rigorously without the use of calculus or more than the simplest of algebra.

The emission of electrons from metals and composite surfaces is similarly dealt with in a simple but satisfying manner. The statement that the exponential form of the diode-current in the retarding-field region shows that this current " can never become zero " might be queried; with −10 volts in the anode, an ordinary diode would require to exist for ten million times as long as the earth has existed so far for there to be a good chance of one electron passing, and hence over practical periods the current will in fact be exactly zero. The voltage range over which such currents are of practical interest is, however, made quite clear.

Actual valve construction is extremely well described for conventional modern types, including those with pressed glass bases, but not including the disk-seal valves used for ultra high frequencies. No mention is made of klystrons, magnetrons or allied devices, but excellent descriptions are given of frequency-changers and of many established forms of multi-grid valves.

The performance of valves in circuits and the tricks by which the best can be got out of them are very well dealt with; the fund of practical experience available within the Philips laboratories shows to excellent effect.

The translation is outstandingly good. The only objection noticed by the reviewer is the use of the term " hail effect " for the fluctuation of current due to the discrete nature of electrons rather than the more usual " shot effect."

Further books in this series will be looked forward to with great interest.

J. H. Fremlin.

Electronic Valves—Book II. Data and Circuits of Receiver and Amplifier Valves

Translated by G. Ducloux. 409 pp. 532 figures. Philips Technical Library, distributed by Cleaver Hume Press, Ltd., London. 1949. Price 21s.

" DATA and Circuits of Receiver and Amplifier Valves " is Book II of a series published by Philips' Technical Library.

This book contains a survey, together with descriptions and data, of the valves brought out by Philips in the years 1933 to 1939. The information covers the use and application of modern receiving, amplifying and rectifying valves, as well as electronic tubes for current stabilizers and regulators, etc. Tables are incorporated of other receiver type valves, cathode ray tubes and special valves, and a large number of receiver circuits are also included, as well as details of measuring instruments for laboratories and workshops.

In general the valve descriptions are brief, but newer principles are more fully treated, and some circuit diagrams are included. The data concerning each valve is given mainly for application purposes.

The opening chapters of this work contain hints on the use of valves and tubes, and undesirable connexions or wrong applications can be avoided by perusal of this section.

" Data and Circuits of Receiver and Amplifier Valves " should interest all those concerned with radio receivers, their associated equipment and electronic tubes.

Electronic Valves—Book III. Data and Circuits of Modern Receiver and Amplifier Valves

Translated by G. Ducloux. 213 pp. 267 figures. Philips Technical Library, distributed by Cleaver Hume Press, Ltd., London. 1949. Price 12s. 6d.

THIS book provides a review giving full descriptions, including applications and circuits, of the receiver/amplifier and rectifier, all glass types of valves developed in 1940 and 1941, and forms a supplement to Book II. Many receiver circuits using the valves mentioned are included, and the book also con-

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tains a description of measuring and auxiliary instruments used both in plant and laboratories as in 1942.

Electronic Valves—Book IV. Application of the Electronic Valve in Radio Receivers and Amplifiers

By B. G. Dammers, J. Haantjes, J. Otte and H. Van Suchtelen. Translated by S. H. Alexander. 416 pp. 256 figures. Philips Technical Library, distributed by the Cleaver Hume Press, Ltd. Price 35s.

THIS book, the fourth of a series on electronic valves by staff of the Philips Organisation, is concerned with the R.F. stages of the receiver. It is a translation of an original in Dutch, but only in a few places in the text is this fact betrayed. A full list of symbols is given in the introduction; except that S instead of g is used for conductance they are similar to the symbols used in English publications. The same is true of circuit symbols, resistance only differing from normal English practice. The subject of the book has been treated by English authors fairly thoroughly from the standpoint both of technician and technologist so that we should not expect to find much that is new, and this indeed is the case. There are five sections dealing respectively with R.F. and I.F. amplification, frequency changing, oscillator tracking, distortion and detection.

The first section starts with the fundamental characteristics of single and double tuned circuits and it draws attention to their presentation in generalized form. The logarithmic ratio ordinate scale, which is employed, could with advantage have been replaced by the linear db scale, and it is surprising that no reference is made to the original work carried out by Beatty (*Wireless Engineer*, October 1932). The authors also seem to be unaware of Sturley's extension of Beatty's curves to band pass circuits with couplings other than by mutual inductance (*Wireless Engineer*, Sept. and Oct. 1943). An excellent feature of this section and, indeed, of the whole book is the generous sprinkling of numerical illustrations though English readers may feel that undue emphasis is placed on an I.F. of 125 kc/s. The use of tuned circuits in association with valves, and the combination of circuits to provide optimum selectivity in V.F. stages are discussed.

The title " Frequency Changing " of the second section, which occupies well over a third of the book, is somewhat a misnomer as a large proportion of the text is devoted to oscillator problems. A short introduction deals with the calculation of g_c and then we are switched to a discussion on the properties of oscillator circuits and the stabilization of oscillator amplitude over the tuning range. The parallel fed oscillator is the subject of the next part, which is followed by discussions on squegging, interaction, electron transit time and frequency drift.

Section 3 shows how tracking error is estimated and oscillator circuit components calculated.

The causes of modulation hum, modulation distortion and cross modulation are considered in the next section, and methods of measuring these valve curvature effects are outlined. Whistles and whistle charts are also dealt with.

The method of presenting detection is conventional and section 5 mainly concerns itself with the diode and its effect on the associated circuits. Reference is made to the distortion which arises when one sideband is missing and to demodulation due to a strong undesired carrier.

It is very probable that this book fulfils a need in the country of its origin, but it is difficult to foresee a large market for it in this country where it has to compete with a number of well established books covering much the same ground.

K. R. STURLEY.

Electronics in Engineering

By W. Ryland Hill. 274 pp. 1st Edition. McGraw-Hill Book Co., Inc., 1949. Price 30s.

THE increasing use of electronics in industry makes very necessary a reliable text-book to which non-electronic engineers can turn to learn something of this branch of technology. The volume under review is an attempt to meet this need. Professor Hill has produced an elementary book of wide scope which contains a number of very misleading statements. Page 141 shows the frequency response of a simple R-C amplifier, with no special precautions to ensure good frequency response, to be flat from (approximately) 100 c/s to 100 kc/s. The name " damplifier " is suggested (humourously) for the cathode-follower, but its unique impedance transformation property is nowhere mentioned. On page 182, we are told that the Q of a tuned circuit depends almost entirely on the resistance of the coil since " practical capacitors are nearly lossless," while on the page

BOOK REVIEWS (Continued)

opposite a typical value of Q for a 500 μh air cored Litz wire coil is shown to be 800 at 800 kc/s. Any reader with no practical knowledge of electronics will certainly find himself at a disadvantage if he relies on this book for information on quantitative aspects of components and apparatus.

On the positive side, the style is free, the diagrams clear and the layout and presentation excellent. Some unusual topics for a book of this level are included, e.g., electronic control circuits, transducers, and the general groundwork is briefly covered. Each chapter concludes with a series of problems to which, unfortunately, no answers are given.

K. G. Lockyer.

Giant Brains ; or Machines That Think

By E. C. Berkeley. 270 pp. Chapman & Hall, 1949. Price 32s.

THE least satisfactory feature of this book is the title, which suggests a naïve or sensational attitude; actually, although informal in style, it is serious in intent and gives a large amount of information about the whole subject of automatic computation. Most of the book is descriptive, giving accounts of the more important large-scale computers so far built or planned.

The contents of the book can be summarized as follows. After introductory chapters, in which the author indicates the field to be covered, discusses what he means by "thinking" and gives short accounts of language and number systems and the storage and transmission of information, there is a detailed description of a hypothetical machine which might be called a basic computer. This is a simple relay-controlled device—the author christens it Simon—working in the scale of 2, which can perform elementary arithmetical and logical operations; this chapter is really an excellent introduction to the theory of one type of automatic computer. These are then the six descriptive chapters, which include accounts of the famous if already obsolescent ENIAC, the MIT differential analyzer, standard commercial punched-card machinery (e.g., IBM in America, Hollerith or Powers Samas here) and an interesting machine built at Harvard in 1947 to solve logical problems. In each case the account is about as complete as could be reasonably expected: the physical construction of the machine is described, also the method of representing and manipulating numbers, and the working out of a simple sum is shown. The chapter on the logical calculator contains an interesting short account of Boolean algebra, the algebra of logical relations. The two final chapters are less factual. In "The Future" the author tries to forecast the developments yet to be made—not only super-calculators, but also electronic translators and typists, even, maybe, an electronic cook. In "Social Control" he discusses the impact of these machines on society. An appendix gives an excellent collection of references to all aspects of the subject—from papers in the Royal Society Proceedings to snappy articles in "Time."

To the English reader, the almost complete omission of reference to development outside America is a little depressing. To be sure, the Americans have plunged into this subject with remarkable if typical enthusiasm; but surely the Cambridge EDSAC merits more than a mere note in the bibliography? More surprising still, the work of F. C. Williams and his Manchester team on the cathode-ray memory tube gets no mention at all.

The whole book is exceedingly interesting and informative, and the author has done a service in writing it. The advent of machines which, once given a set of instructions, will carry out a tremendous computation at about a thousand times the speed of a human computer, is a matter of prime importance, which should be appreciated throughout the scientific world. The reviewer would rather not call these machines brains, nor say that they think: they perform well-defined operations in response to precise instructions, just as does an automatic telephone exchange or a desk calculating machine—the difference in degree is enormous, but not in kind.

A good (if not socially very valuable) illustration of this difference in degree is provided by the recent publication of the value of π to 2035 decimals: this was calculated by the ENIAC in the course of two week-ends. Nevertheless, the machine did not "think" of the way to do this; just as a pianola can play a Beethoven sonata, but only if it is given the right roll.

J. Howlett.

NOTES FROM THE INDUSTRY

The Polytechnic Electrical Engineering Courses

In the session 1950-1951 the Polytechnic offer a wide range of courses in the electrical engineering field, providing a thorough training in the principles and techniques of the subjects.

The telecommunications courses are approved for the award of the "Ordinary" and "Higher" National Certificates granted by the Institution of Electrical Engineers in conjunction with the Ministry of Education, and are also recognized by the City and Guilds of London Institute as preparing for their examinations.

The courses in radio and television service work prepare for the Radio and Television Servicing Certificates awarded jointly by the City and Guilds and the Radio Trades Examination Board.

The radio, television and measurements laboratories are well equipped with modern apparatus, most of which has been specially designed for advanced experimental work. These laboratories are available for those wishing to prepare theses for the Higher University Degrees, or to pursue an advanced study of a particular subject.

Full details and a prospectus can be obtained on application to the Polytechnic, Electrical Engineering Department, 309 Regent Street, London, W.1.

Record Demands for Space at 1951 British Industries Fair

The 1951 British Industries Fair will be held in London, at Earls Court and Olympia, and in Birmingham at Castle Bromwich from 30 April to 11 May.

The response to application forms for space is already an all-time record and indicates that this will be the largest and most comprehensive trade fair ever organised in Britain. The forms were sent out during the middle of July and by the end of July twice as much space had been applied for as at the corresponding period for any previous Fair. The organisers are arranging to extend the stand space as far as the limits of the fair buildings and the interests of the exhibitors will allow. The closing date for applications was 19 August and every effort will be made to accommodate all requests received though sizes of individual stands may have to be limited.

Congress of International Scientific Film Association

The Fourth Congress of the International Scientific Film Association is being held in Florence from 14 October-22 October, 1950.

As in the case of the three previous Congresses held in Paris, London and Brussels, there will be a Festival of Scientific Films during the period of the Congress. A special committee of the British Scientific Film Association has been formed to recommend films for submission to this Festival.

Apart from the Festival of Films, there will be specialist group meetings of the Medical, Research Film and Industrial Film Committees, and also the business meetings of the General Assembly of the Congress.

Engineering Education in Technical Institutes and Colleges

The Regional Advisory Council for Higher Education have published a pamphlet on engineering education in the area, in collaboration with the Engineering and Allied Employers (London and District) Association.

The pamphlet gives general information about the various types of courses available (e.g., for engineering crafts, National Certificates and so on) for students from the time they leave school. An outline is also given of facilities available for students who are qualified to follow courses for university degrees and research. There is also a list of the colleges in the region concerned with the courses offered during the session 1949/50, which it is thought may be a useful guide on which to base inquiries regarding present or future facilities.

Copies can be obtained from the offices of the Regional Advisory Council, 20 Fitzroy Square, London, W.1, price 1s., post free.

A Studio Without Engineers

In the past the BBC frequently has wished to include in its programmes a topical interview with some celebrity just disembarked from one of the transatlantic liners. This has meant the transport of an engineer and a duplicate set of outside broadcasting equipment from Bristol or some nearer point where the outside broadcasting staff were working at the moment. The dislocation in the planning and staffing of outside broadcasts has been unavoidable, and the cost in transport and engineer hours has been considerable.

To meet the difficulty, the BBC Engineering Division is experimentally installing an unattended set of studio equipment in the Guildhall at Southampton. The equipment will normally be locked up, but is installed in such a way that it can be switched on and send its programme via the G.P.O. lines to Bristol, without anyone being present other than the celebrity and the programme representative responsible for the interview.

The programme representative will switch on the equipment, which is installed in duplicate, and the microphone output with the necessary amplification, and electrical corrections, will, it is hoped, appear at Bristol, where it will be switched into the broadcast chain on a prearranged cue.

Greece Orders British Telecommunication Equipment

As part of the complete reconstruction of the Greek telecommunication system, the Greek P. T. & T. authorities have ordered from Standard Telephones and Cables, Ltd., their main overseas radiotelephone link system.

This will provide telephone circuits to the most modern standards, with the U.K. and the U.S.A.

The equipment to be supplied covers the complete system and includes two 4 kW Single-Sideband transmitters for simultaneous working to London and New York, with a 40 kW power amplifier for use when transmission conditions are poor, two single-sideband receivers, equipment for connecting four speech circuits to the telephone network including a privacy system and aerials, power plant, etc.

The value of the installations, work on which will commence early next year, approximates to £80,000.

Radiomobile Limited A Change of Name

Recently the title of Radiomobile, Ltd., the company responsible for the manufacture and distribution of "His Master's Voice" Automobile Radio Equipment, changed to S. Smith & Sons (Radiomobile), Ltd.

Rome Ultrasonics Convention

We regret that a misprint appeared on page 393 of the September issue of ELECTRONIC ENGINEERING. The first column, line 24, should read "crystals, an accuracy of 0.05 per cent being attained" and not "crystals, an accuracy of 0.5 per cent being attained."

MEETINGS THIS MONTH

INSTITUTION OF ELECTRICAL ENGINEERS

All London meetings are held at the Institution of Electrical Engineers, Savoy Place, W.C.2.

Date: October 5. Time: 5.30 p.m.
Inaugural Address as President.
By: Sir Archibald J. Gill, B.Sc.(Eng.).

Measurements Section
Date: October 10. Time: 5.30 p.m.
Chairman's Address.
By: G. A. V. Sowter, Ph.D., B.Sc.(Eng.).

Radio Section
Date: October 18. Time: 5.30 p.m.
Chairman's Address.
By: C. F. Booth, O.B.E.

Informal Meetings
Date: October 23. Time: 5.30 p.m.
Discussion: Radio Interference and the New Wireless Telegraphy Act.
Opened by the President.

District Meeting
Date: October 11. Time: 7 p.m.
Held at: The Electricity Showrooms, Southern Electricity Board, 37, George Street, Oxford.
Lecture: Present Problems in Television.
By: G. H. Watson.

East Midland Centre
Date: October 17. Time: 6.30 p.m.
Held at: Gas Demonstration Theatre, Nottingham.
Informal Lecture: The Operation and Maintenance of Television Outside-Broadcast Equipment.
By: T. H. Bridgewater.

Cambridge Radio Group
Date: October 17. Time: 6 p.m.
Held at: The Cambridgeshire Technical College, Cambridge.
Chairman's Address.
By: F. H. Townsend.

North-Eastern Radio and Measurements Group
Date: October 16. Time: 6.15 p.m.
Held at: King's College, Newcastle-on-Tyne.
Chairman's Address.
By: Professor J. C. Prescott, D.Eng.

North Midland Centre
Date: October 3. Time: 6.30 p.m.
Held at: British Electricity Authority, Yorkshire Division, 1, Whitehall Road, Leeds.
Chairman's Address.
By: W. F. Smith, B.Sc.

North-Western Radio Group
Date: October 25. Time: 6.30 p.m.
Held at: The Engineers' Club, Albert Square, Manchester.
Chairman's Address.
By: A. G. S. Smith.

South Midland Radio Group
Date: October 2. Time: 6 p.m.
Held at: James Watt Memorial Institute, Great Charles Street, Birmingham.
Lecture: Economies in the Planning, Design and Operation of a Sound Broadcasting System.
By: R. T. B. Wynn, C.B.E., M.A.

THE TELEVISION SOCIETY

Engineering Group
Date: October 13.
Held at: C. E. A., 164, Shaftesbury Avenue, London, W.C.2.
Lecture: The English Electric Television Receiver Model 1550.
By: D. Feuings, B.Sc.

INSTITUTION OF POST OFFICE ELECTRICAL ENGINEERS

Date: October 3. Time: 5 p.m.
Held at: Institution of Electrical Engineers, Savoy Place, W.C.2.
Lecture: London Station Cable and Wireless Telegraph Section, L.T.R.—An Engineering Survey.

SECRETARIES OF ASSOCIATIONS

INSTITUTION OF ELECTRICAL ENGINEERS

The Secretary, Institution of Electrical Engineers, Savoy Place, W.C.2.

Cambridge Radio Group
G. E. Middleton, M.A., University Engineering Laboratory, Cambridge.

North-Eastern Radio and Measurements Group
G. A. Kysh (Asst. Sec.), Carliol House, Newcastle-on-Tyne, 1.

North-Western Radio Group
A. L. Green (Asst. Sec.), 244, Brantingham Road, Chorlton-cum-Hardy, Manchester, 21.

South Midland Radio Group
W. H. Brent, B.Sc., Regional Director's Office, Midlands Region (G.P.O.), Civic House, Great Charles Street, Birmingham, 3.

BRITISH INSTITUTION OF RADIO ENGINEERS

The General Secretary, 9, Bedford Square, London, W.C.1.

Scottish Section
A. A. M. Turnbull, A.M.Brit.I.R.E., 68, Lauderdale Gardens, Glasgow, W.2.

West Midlands Section
R. A. Lampitt, A.M.Brit.I.R.E., 20, Northfield Grove, Merry Hill, Wolverhampton.

INC. RADIO SOCIETY OF GT. BRITAIN

General Secretary, New Ruskin House, Little Russell Street, W.C.1.

BRITISH SOUND RECORDING ASSOCIATION

Richard W. Lowden, "Wayford," Napoleon Avenue, Farnborough, Hants.

TELEVISION SOCIETY

Engineering Group
G. T. Clack, 10, Tantallon Road, Balham, London, S.W.12.

RADAR ASSOCIATION

Technical Committee: J. C. Warren, 57, Gallows Hill, Kings Langley, Herts.

INSTITUTION OF POST OFFICE ELECTRICAL ENGINEERS

W. H. Fox, A.M.I.E.E., Engineer-in-Chief's Office (T. P. Branch), Alder House, E.C.1.

INSTITUTION OF ELECTRONICS

Lecture Sec.: W. Summer, 31, Beech Road, Bournville, Birmingham, 30.

North-West Branch
L. F. Berry, 105, Birch Avenue, Chadderton, Lancs.

SOCIETY OF RELAY ENGINEERS

T. H. Hall, M.Brit.I.R.E., 23, Dalkeith Place, Kettering, Northants.

SOCIETY OF INSTRUMENT TECHNOLOGY

L. B. Lambert, 55, Tudor Gardens, London, W.3.

By: R. G. Griffiths.
Date: October 18. Time: 5 p.m.
Held at: Conference Room, 4th Floor, Waterloo Bridge House, S.E.1.
Lecture: Some Aspects of Auto-Exchange Maintenance Organisation.
By: G. R. Stapley.

BRITISH INSTITUTION OF RADIO ENGINEERS

London Section
Date: October 18. Time: 6.30 p.m.
Held at: London School of Hygiene and Tropical Medicine, Gower Street, London, W.C.1.
Lecture: Signal Sources for Television Testing.
By: D. W. Thomasson.

North-Eastern Section
Date: October 11. Time: 6 p.m.
Held at: Institution of Mining and Mechanical Engineers, Neville Hall, Westgate Road, Newcastle-on-Tyne.
Lecture: High Fidelity Reproduction.
By: H. J. Leak.

West Midland Section
Date: October 25. Time: 7.30 p.m.
Held at: Works Canteen, R.G.D. Co., Ltd., Pale Meadow Works, Bridgenorth, Shropshire.
Lecture: Loudspeaker Design.
By: A. E. Falkus.

North-Western Section
Date: October 5. Time: 6.45 p.m.
Held at: College of Technology, Manchester.
Lecture: Education in Industry.
By: Dr. P. F. R. Venables.

SOCIETY OF INSTRUMENT TECHNOLOGY

Teeside Section
Date: October 18.
Lecture: Temperature Measurement at the National Physical Laboratory.
By: J. A. Hall.

Midland Section
Date: October 20. Time: 7 p.m.
Held at: The Arden Hotel, New Street, Birmingham.
Lecture: Earthquakes and How They are Recorded.
By: H. V. Shaw.

North-West Section
Date: October 10.
Lecture: The British Scientific Instrument Manufacturers' Research Association.
By: A. J. Philpot.

South Wales Section
Date: October 4.
Lecture: Some Electronic Developments of Instruments.
By: Group Captain Hockey.

SOCIETY OF RELAY ENGINEERS

Date: October 3. Time: 2.30 p.m.
Held at: The Engineers' Club, 17, Albert Square, Manchester, 2.
Lecture: Medium Voltage Feeders and their Associated Transformer Equipment for Long Distance Transmission of Audio Frequency Power.
By: K. A. Russell, B.Sc.

INSTITUTION OF ELECTRONICS

Date: October 17. Time: 6.30 p.m.
Held at: Reynolds Hall, College of Technology, Manchester.
Lecture: Ultrasonics.
By: A. E. Crawford.

THE RADAR ASSOCIATION

Date: October 3. Time: 7.30 p.m.
Held at: The Albert Tavern, Victoria Street, London, S.W.1.
Monthly meeting.

East Anglian Meeting
Date: October 28.
Held at: The Lamb Inn, Norwich.
Monthly meeting.